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BASIC ELECTRONICS TECHNOLOGY AND TESTING PRACTICES

CHAPTER 11 TESTING TECHNIQUES AND PRACTICES

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CHAPTER 11

TESTING TECHNIQUES AND PRACTICES

11-1. INTRODUCTION. This chapter explains the testing techniques and practices which you may have to use to operate, maintain, or repair communications, electronic, and associated equipment. These testing techniques and practices make use of the types of test equipment described in Chapter 2 and the fundamental test methods discussed in Chapter 4. The information is divided into 12 sections. Tests that are applicable to several types of electronic equipment, such as radar, radio, and television, are included in Section I. Tests which are applicable to communication equipment, which includes radio, facsimile, telephone, and teletypewriter equipment, are included in Section II. Tests which are normally made on the electronic components of radar and identification equipment are described in Section III. Tests which are peculiar to navigational aids devices are included in Section IV. Television equipment tests are discussed in Section V. Radiac equipment tests are described in Section VI. Tests for synchro and servo equipment are discussed in Section VII. Power supply tests are covered in Section VIII. Tests for electrical machinery are described in Section IX. Tests for infrared equipment are given in Section X. Computer equipment tests are explained in Section XI. Tests for telephone and telegraph equipment are described in Section XII.

SECTION I

GENERAL ELECTRONIC EQUIPMENT TESTING

11-2. GENERAL.

11-3. To ensure operational reliability of electronic equipments and systems, the measurements, tests, checks, and inspections described in this section must be made. These maintenance procedures provide information to establish an acceptable operating condition or reveal undesirable conditions which exist in the equipment tested. In the former instance, testing and inspection prove that the equipments are performing within tolerances, and are a form of maintenance which is preventive in nature. In the latter instance, testing and inspection reveal conditions which are undesirable, and which must be corrected if the equipments are to perform as designed. The result of this testing establishes another form of maintenance, which is corrective in nature. Thus it is necessary to perform both preventive and corrective maintenance to keep electronic equipments in continuous service at optimum or peak performance, which is reasonably close to the standards set by the manufacturers.

11-4. PREVENTIVE MAINTENANCE.

11-5. GENERAL.

11-6. The most effective maintenance work is preventive in nature, potential failures being detected and corrected before they have a chance to develop. Preventive or scheduled maintenance is defined as those measures taken periodically, or when needed, to achieve maximum efficiency in performance, to ensure continuity of service, to reduce major breakdowns, and to lengthen the useful life of the equipment or system. This form of maintenance consists principally of cleaning, lubrication, and periodic inspections aimed at discovering conditions which, if not corrected, may lead to malfunctions requiring major repair.

11-7. INSPECTIONS.

11-8. Inspections fall into two main categories. First, there is the regular visual inspection of the mechanical aspects of the equipments, which is conducted for the purpose of finding dirt, corrosion, loose connections, mechanical defects, and other sources of trouble. Second, there are the functional inspections that are accomplished through periodic tests and through the less-frequent bench tests. To realize optimum results from the regular functional inspections, a careful record of the performance data on each equipment must be kept. The value of these records will be demonstrated in a number of ways. Comparison of data taken on a particular equipment at different times reveals slow, progressive drifts that may be too small to show up significantly in any one test. While the week-to-week changes may be slight, they should be followed carefully, so that necessary replacements or repairs may be effected before the margin of performance limits is reached. Any marked variations should be regarded as abnormal, and should be investigated immediately. Another advantage in keeping systematic records of performance and servicing data is that maintenance personnel develop a more rapid familiarization with the equipment involved. The accumulated experience contained in the records serves as a guide to swift and accurate trouble shooting.

11-9. The actual work of testing and servicing, as well as that of recording performance data, should be done systematically. While a logical sequence of steps is required, this does not imply the rigid necessity of making only a step-by-step progression. Working within the over-all pattern of the procedure, maintenance personnel should analyze the results obtained, to eliminate unnecessary steps.

11-10. CORRECTIVE MAINTENANCE.

11-11. GENERAL.

11-12. Corrective maintenance consists of the location and correction of troubles whenever an equipment or system fails to function properly. The trouble may be corrected by mechanical or electrical adjustments, or it may be necessary to replace one or more parts. As a rule, reports are submitted when a defective part is replaced. These reports are important, because the statistics gathered from them may be used to determine the future stock spares requirements. These statistics may also be used to improve the design of equipments on future contracts.

11-13. TROUBLE SHOOTING.

11-14. Corrective maintenance, for the most part, is concerned with trouble shooting, which can be divided into two phases. The first phase is system trouble shooting. It is based on the starting procedure, and is designed to locate the unit in which the trouble occurs. The second phase is unit trouble shooting, which is designed to locate the trouble in the unit in which it occurs. In some cases it is possible to determine which unit is at fault without following the system trouble-shooting method to isolate the unit. However, quite often it is impossible to determine which unit is at fault until the system method has been employed in whole or in part.

11-15. TROUBLE ISOLATION.

11-16. When abnormal operation has been traced to a particular stage or to a functionally related group of stages, its cause must be further isolated and identified as due to a particular faulty component or group of components. To do this it may sometimes be necessary to disassemble the equipment, either in whole or in part. After disassembly, the trouble may be immediately apparent through a mere visual inspection, whereupon the trouble should be corrected by repair or replacement. If the trouble is not immediately apparent, a more detailed procedure should be followed to isolate and repair or replace the actual circuit component responsible for the failure. This procedure consists of tube checks, point-by-point resistance and voltage checks, waveform analysis, and finally, repair or replacement of the defective component.

11-17. TUBE TESTING.

11-18. Electron tube failures are responsible for the largest percentage of troubles that occur in electronic equipments or systems. However, if a particular system uses a great number of tubes, it is obviously impracticable, as well as poor policy, for you to attempt to locate faults by general tube checking. Only when the fault has been traced to a particular stage should any tubes be tested, and then only those associated with the improperly functioning circuits.

11-19. When replacing a tube in a circuit, note and record the positions of the equipment controls before changing the setting of any of them. Test the new tube for shorts and gas before inserting it into the circuit. If adjusting the controls with the new tube in place does not correct the abnormal condition, return the controls to their original positions, and, unless a reliable tube tester shows the original tube to be defective, reinsert the old tube in the original circuit. After replacing a tube in a circuit, decide immediately whether or not to keep the old tube. If the tube is kept, it should be labeled so that it will not be replaced in the same socket. Do not change tubes indiscriminately; otherwise, tubes whose exact age and condition are unknown (or uncertain) will accumulate. In many high-frequency circuits, the interelectrode capacitance of a tube is a characteristic of a tuned circuit; therefore, when tubes are changed, the tuning of the circuits may be upset. Thus, when certain tube substitutions are made, the unit may have to be realigned.

11-20. RESISTANCE MEASUREMENTS.

11-21. Defective parts can usually be quickly located by measurement of the dc resistance between various points in the circuit and a reference point or points (usually ground), because when a fault develops it will generally produce a change in the resistance values. Point-to-point resistance charts can be used advantageously at this time. The values given, unless otherwise stated, are measured between the indicated points and ground.

CAUTION

Before making resistance measurements, make sure that the power to the equipment under test has been turned off, and discharge all filter capacitors.

11-22. VOLTAGE MEASUREMENTS.

11-23. Since most troubles encountered in equipments and systems either result from abnormal voltages or produce abnormal voltages, voltage measurements are considered an indispensable aid in locating trouble. Testing techniques that utilize voltage measurements also have the advantage that circuit operation is not interrupted. Point-to-point voltage measurement charts which contain the normal operating voltages encountered in the various stages of the equipment are available to the technician. These voltages are usually measured between the indicated points and ground, unless otherwise stated. When voltage measurements are taken, it is considered good practice to set the voltmeter on the highest range initially, so that any excessive voltages existing in a circuit will not cause overloading of the meter. To obtain increased accuracy, the voltmeter may then be set to the designated range for the proper comparison with the representative value given in the voltage charts. When checking voltages, it is important to remember that a voltage reading can be obtained across a resistance, even if that resistance is open. The resistance of the meter (and the multipliers) forms a circuit resistance when the meter prods are placed across the open resistance. Thus, the voltage across the component may appear to be approximately normal, as read on the meter, but may be abnormal when the meter is disconnected from the circuit. Therefore, to avoid unnecessary delay in the troubleshooting procedure, it is good practice to make a resistance check on a "cold" circuit (before applying power), to determine whether the resistance values are normal.

11-24. If the internal resistance of the voltmeter and multiplier is approximately comparable in value to the resistance of the circuit under test, it will indicate a considerably lower voltage than the actual voltage present when the meter is removed from the circuit. For a discussion of the effects of voltmeter loading, refer to paragraph 8-83, Volume VI. The sensitivity (in ohms-per-volt) of the voltmeter used to prepare the voltage charts for Technical Orders is always given on those charts; therefore, if a meter of similar sensitivity is available, it should be used, so that the effects of loading will not have to be considered.

11-25. The following precautions are general safety measures, recommended for the measurement of voltages when working with electronic equipment. You should constantly keep in mind that all voltages are dangerous. It should be recognized that even low voltages or currents can be hazardous and even lethal if unusual conditions should exist. Some of these conditions could be excessive humidity, wet areas, lack of protective matting or other equipment, improper grounding and others. The criterion for electrocution consists of many variables; the real measure of a shock's intensity lies in the amount of current forced through the body, and not the voltage. While any amount of current over 10 MA. is capable of producing painful to severe shock, currents between 100 and 200 MA. are absolutely lethal, refer to paragraph 2-37 - 2-70, Volume I. However, lest these details be misinterpreted, the only reasonable conclusion that can be drawn is that 75 volts are just as lethal as 750 volts. When it becomes necessary to measure high voltages the following precautions should be observed:

a. Connect the ground lead of the voltmeter first. While making measurements, place one hand in a pocket or behind the back.

b. If the voltage to be measured is less than 600 volts, place the end of the test prod on the point to be tested, which may be either positive or negative with respect to ground.

11-26. If the voltage to be measured is greater than 600 volts, proceed as follows: shut off the circuit power, discharge any filter capacitors, and temporarily ground the point to be measured; then connect (clip on) the proper test lead to the high potential point, move away from the voltmeter. Do not come in contact with any part of the equipment while the power is on. This is particularly important when the voltage under measurement is across two points, both of which are above ground potential. If an electronic voltmeter equipped with a polarity reversing switch is to be used, refer to paragraph 8-134, Volume VI.

11-27. WAVEFORM COMPARISON.

11-28. The measurement and comparison of waveforms is considered to be a very important part of the circuit analysis used in trouble shooting. In some circuits (for example, pulse circuits), waveform analysis is indispensable. Waveforms may be observed at test points, shown in the waveform charts, or on schematic diagrams which are a part of the maintenance literature supplied for each equipment. It should be noted that the waveforms given in instruction books are often idealized, and do not show some of the details which are normally present when the actual waveform is displayed on an oscilloscope. However, by comparing the observed waveform with the reference waveform, faults can be localized rapidly. An appreciable departure from the normal waveform indicates a fault that is located between the point where the waveform is last seen to be normal and the point where it is observed to be abnormal. For example, if a waveform is observed to be normal at the grid circuit of a stage, and abnormal at the plate circuit of the same stage, the trouble lies in that stage or possibly the input of the following stage.

11-29. If there is no trouble present in an equipment or system, a waveform observed at a point in the equipment should closely resemble the reference waveform given for the test point. The reference waveforms supplied with maintenance literature are the criterion of proper circuit performance. However, test equipment characteristics or usage can cause distortion of the observed waveforms, even though the equipment or system is operating normally. Several of the most common causes of these conditions are summarized as follows:

a. The leads of the test oscilloscope may not be placed in the same manner as those of the oscilloscope used in preparing the reference waveforms, or the lead lengths may differ considerably. This is particularly significant in the case of shielded test leads, where the capacitance per unit length is a factor.

b. A type of test oscilloscope having different values of input impedance, different sweep durations, or different frequency response may have been used.

c. The equipment operating (and servicing) controls may not have settings similar to those used when the reference waveforms were prepared. This condition is normally to be expected when servicing adjustments are made in terms of their effect on the shape or amplitude of an observed waveform.

d. The vertical or horizontal amplitudes of the reference and test patterns may not be proportional. This will produce apparent differences between the waveforms when actually there is no difference.

11-30. Whether or not a minor waveform discrepancy may be disregarded depends upon the type of circuit being traced. A minor discrepancy is not significant unless the nature of the discrepancy, in consideration of the circuit under test, indicates faulty operation of the equipment. In general, time should not be wasted in searching for faults when relatively minor differences are detected between the reference waveforms and those obtained by test.

11-31. PERFORMANCE TESTING.

11-32. GENERAL.

11-33. Performance testing of electronic equipments and systems comprises certain specific tests for each category of equipment. This test data is included in equipment service manuals, preventive maintenance work cards, facility manuals, and system manuals. In practice, performance checks determine, with minimum effort and maximum accuracy, the operating condition of a complete section of an equipment or system. Performance tests that are common to radio, radar, and television equipment are included in this section. These tests are: receiver noise, gain, and sensitivity measurements, transmitter power output measurements, standing wave measurements, frequency spectrum measurements, and impedance measurements of antennas and transmission lines.

11-34. RECEIVER NOISE MEASUREMENTS.

11-35. GENERAL. Theoretically, it is possible to amplify a feeble electrical signal by practically any desired factor; however, it is still not possible practically to discern an arbitrarily weak signal because of the presence of random electrical fluctuations. or "noise." If the intelligence signal entering the receiver becomes progressively weaker, it subsides eventually into the fluctuating background of noise and is lost. The limit of sensitivity for low-frequency receivers, as for all receivers, is set by random electrical disturbances. However, in this case, the largest random disturbances with which the signal must compete generally originate, not in the receiver itself, but elsewhere in space. Whatever the external noise source, whether from an electrical apparatus or from intersteller space, these disturbances enter the receiver by way of the antenna. The crucial quantity, therefore, is the ratio of the field strength of the signal in the neighborhood of the antenna to that of that of the extrinsic noise or interference. The absolute magnitude of signal and interference power available at the antenna terminals are of little importance; only their ratio, which, for example, might be favorably altered by the use of a directional antenna pattern, determines the ultimate performance. More significantly, it explains why the emphasis for low-frequency receivers involves discrimination against some of the external noise (for example, by greater frequency selectivity) rather than by reduction of the noise inherent in the receiver.

11-36. In the microwave region, substantially all of the noise originates in the receiver, not because microwave receivers are noisier or more imperfect receivers than low-frequency receivers, but because environmental low-frequency noise is enormously greater than high-frequency noise. In fact, such noise in the microwave region is almost wholly negligible; it is the noise that originates in the receiver that interferes with the signal. The inherent noise generated in a receiver (or receiver section of more complex equipment) establishes the minimum limit of signal that a receiver can usefully amplify. Therefore, maximum receiver sensitivity, in most cases, is not determined by the gain of the particular receiver, but by the magnitude of the input circuit noise. The noise is a result of the random motion of the electrons in the antenna and receiver circuits (thermal agitation or resistance noise) and tubes (tube or shot noise). Thermal and tube noise can be considered collectively as receiver noise. Receiver noise exists across the entire radio-frequency spectrum, and its magnitude increases with an increase in temperature. Because the noise is across the spectrum, the noise level increases also with an increase in pass band. In normal operating circuits, only the self-generated noise in the initial amplifier stages is significant, since these stages are subject to maximum amplification. This is true for a-m, fm, television, ssb, radar, etc; although for systems employing fm methods of detection, the noise assumes a greater degree of importance, because receiver noise becomes more appreciable at the higher frequencies at which these equipments usually operate, and because of the much larger pass band generally employed.

11-37. Since receiver noise determines the weakest signal that can be practically detected, its behavior and measurement is of fundamental importance for equipment which may be used to receive very low-intensity signals.

11-38. NOISE FIGURE. An ideal receiver would be one with no noise other than that caused by thermal agitation. The degree to which a receiver approaches this ideal is indicated by the noise figure of the receiver, and may be defined as:

$$\text{Noise figure} = \frac{\text{Signal-to-noise power ratio of ideal receiver}}{\text{Actual signal-to-noise power ratio of receiver output}}$$

For simple test methods, it may be expressed as the ratio of noise power at the input of the receiver required to double the noise output of the receiver; since it is a power ratio it is usually expressed in db's. Noise figures of 2 to 4 db are obtainable for very quiet receivers, but ratios of 6 to 12 db are more typical. This measurement is used primarily for rf receivers, but it can also be applied to devices such as microphones, electromechanical equipment, and photoelectric equipment.

11-39. NOISE GENERATOR METHOD. A noise generator is designed to produce a random noise signal which covers a frequency range in excess of the receiver bandwidth. One such instrument uses what is called a temperature-limited diode, operated at temperature saturation, as the noise-signal source. When a diode is operated under these conditions, the noise produced is proportional to the dc input current. Other types of generators employ thermal noise at elevated temperatures, or use certain types of gas discharges. However, most noise generators up to microwave frequencies are based on the shot noise generated by a temperature-limited diode. Regardless of the type used, the dc input reading of the generator can easily be converted to obtain the

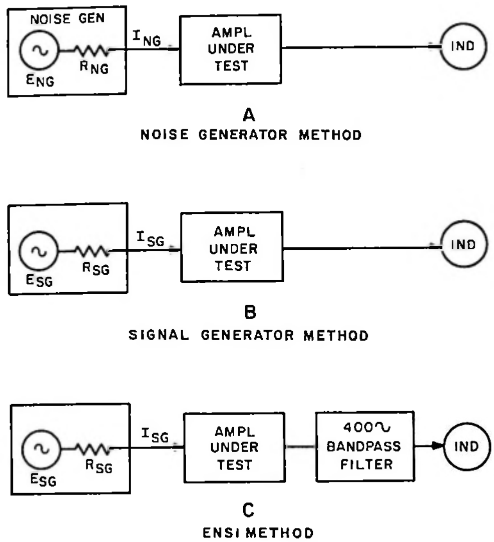


Figure 11-1. Methods for Making Noise Figure Measurements

true noise power. The noise-generator method of determining the noise figure has the advantage that no knowledge of either the gain or the response characteristics of the amplifier is necessary, since the amount of noise from the noise generator is amplified and governed by the effective bandwidth in exactly the same manner as are the thermal and tube noise of the receiver. The noise-generator method of measurement consists of comparing the noise actually present in the receiver with the calibrated output of the noise generator.

11-40. Figure 11-1A shows a block diagram for a noise figure test using a calibrated noise generator. For an accurate measurement, the noise generator output impedance (R_{NG}) is adjusted to be the same impedance as the normal signal source to the equipment under test. In most cases this value is given in the equipment Technical Order, and is the impedance at the transmission line termination from an antenna or antenna multicoupler. For best results, the shortest possible leads should be used between the noise generator and the receiver or converter input. Coaxial leads with connectors in good condition will be used in most cases. Checks should be made for the presence of spurious signals which may affect the output power of the equipment being tested. To prevent feedback when converter equipment is tested, the input and output leads should be separated as much as possible if the frequency difference between the input and output signals is not great. The measurements are ideally performed in a shielded room; if this is not possible, a location should be chosen where a minimum of interference from radio transmitters, radar transmitters, or other electrical devices will be encountered. An advantage of the noise generator method is that it eliminates the necessity of measuring the receiver or converter bandwidth if no spurious signals are present in the output.

11-41. The indicator (an ac voltmeter, db meter, or milliwatt meter) may be connected across either the detector load or the receiver output. A dc meter can also be used as the indicator to measure the dc detector load voltage, or, in the case of fm receivers, to measure the voltage developed across the resistor in series with the first limited grid. The automatic volume control and noise limiter circuits, if provided, should be disabled. Set the receiver gain control so that the receiver noise provides a convenient indication on the output meter, and note this value. Turn the noise generator on and adjust its output until the noise power is doubled. If the receiver or meter detector is a square-law-type device, as is usually the case with detectors for low signal levels, then for doubled power input the detector output power is doubled. If an ac voltmeter is used for the indicator, the noise generator should be adjusted for an output voltage 1.4 times the no-input voltage indication; if a db meter is used, the noise generator should be adjusted for a 3 db increase over the no-input meter indication; if a milliwattmeter is used, the noise generator should be adjusted for twice the no-input reading. The noise figure is then indicated on the output level control of the noise generator. The noise figure in db is 10 times the common logarithm of the indicated noise figure.

11-42. SIGNAL GENERATOR METHOD. This method has an advantage in that sine-wave generators are available at electronic equipment maintenance shops more often than are noise generators. However, the signal generator method is not as practical or accurate as the noise generator method for field measurements. When using the signal generator method, you must take into account the bandwidth (B) and the response curve of the receiver or converter under test. Generally, the bandwidth used is the frequency range between the half-power points of the response curve. For an accurate measurement, the sine-wave generator output impedance (R_{SG}) should be the same impedance as the normal signal source for the receiver or converter under test.

11-43. Figure 11-1B shows a block diagram for a noise figure test using a sine-wave generator. The measuring procedure is similar to that for the noise generator method. First, with no signal output from the sine-wave generator, measure the noise power output of the equipment under test. Then turn the signal generator on, set the output signal at the center frequency of the response curve for the equipment under test, and adjust the output signal level until the test meter indicates twice the power of the no-signal level. The noise figure can then be determined by the following formula:

$$NF\ (db) = 10 \log_{10} \frac{(I_{SG})^2 R_{SG}}{KTB}$$

or

$$NF\ (db) = 10 \log_{10} \frac{E_{SG}^2}{R_{SG}KTB}$$

where:

NF (db) = noise figure in db

I_{SG} = signal generator current

R_{SG} = signal generator output impedance

E_{SG} = signal generator output voltage

B = receiver bandwidth between half-power points

K* = Boltzmann's constant = 1.38×10^{-23} joule (watt-second) per degree Kelvin

T* = absolute temperature in degrees Kelvin

* When noise calculations are being made, the reference standard temperature, T = 290°K (62.6°F) is often used; this value gives KT the convenient value of 4×10^{-21}

11-44. ENSI METHOD. The equivalent-noise-sideband-input (ensi) method of noise level measurement determines the equivalent input voltage of all random noise which appears in the output of the receiver being tested. This test is sometimes used in preference to other methods of measuring noise level because, over a limited frequency range, it is not appreciably affected by changes in the signal input. Figure 11-1C shows a block diagram of the ensi test connections.

11-45. This test can be made without disabling the automatic volume control circuit. The receiver volume control should be set to avoid overloading the audio amplifiers, and the tone control, if present, should be set for maximum high-frequency response. The signal generator is set at the center frequency of the receiver response curve and adjusted for an unmodulated carrier signal output level of approximately 5 mv for receivers having sensitivities less than this value, and 50 mv for receivers having sensitivities greater than 5 mv. An rms voltmeter is connected in a similar manner as for the noise generator method, and used to measure the noise output power (P_N). A 400-cycle bandpass filter is connected between the receiver and the output meter. After the noise power is measured, the signal generator is modulated 30 percent by a 400-cps signal, and the output signal power (P_S) is measured. If the 400-cycle filter is not used, the signal and noise output power can be measured together. The signal output power can then be calculated, if desired, by subtracting the noise output from the combined power output. The noise level in terms of ensi may then be calculated from the following formula:

$$E_N = ME_{SG} \frac{P_N}{P_S}$$

where:

E_N(mv) = ensi

E_{SG} (mv) = signal generator output

P_N = noise power output

P_S = signal power output

M = modulation factor

11-46. FM RECEIVER CONSIDERATIONS. The use of a filter is essential in measuring the noise characteristics of an fm receiver. This necessity arises because the signal-to-noise ratio is improved by a large frequency deviation. Consequently, an fm signal suitably modulated at 400 cps is applied to the receiver, and the output of the receiver is passed through a 400-cycle rejection filter in order to determine the noise output. The reading obtained in the absence of filtering will usually suffice as the measurement of the useful signal. Express the noise characteristic as the ratio of signal voltage (or power) to noise voltage (or power) in db.

11-47. RADAR RECEIVER CONSIDERATIONS. In the microwave frequency range in which radar operates, virtually all noise originates in the receiver. The main sources of noise are the crystal mixer, the preamplifiers, and the local oscillator. Both the noise and sine-wave generator methods of measurement can be applied.

11-48. RECEIVER GAIN MEASUREMENTS.

11-49. GENERAL. When trouble shooting an insensitive receiver, it is often advisable to perform gain measurement of suspected amplifier stages. The voltage gain of any stage is given by the fundamental relationship:

$$\text{Voltage gain (VG)} = \frac{\text{Voltage at output of stage}}{\text{Voltage at input of stage}}$$

That is to say, if V_1 is considered the voltage applied to the stage under measurement, and V_2 is the measured voltage applied to the grid of the succeeding stage, then

$$VG = \frac{V_2}{V_1}$$

It is important, when making gain measurements, that the normal operation of the stage not be disturbed by the test equipment.

11-50. A-M RECEIVERS. The preferred method for determining the gain is by introducing a meter (generally a power or db meter) into a low-impedance portion of the e-

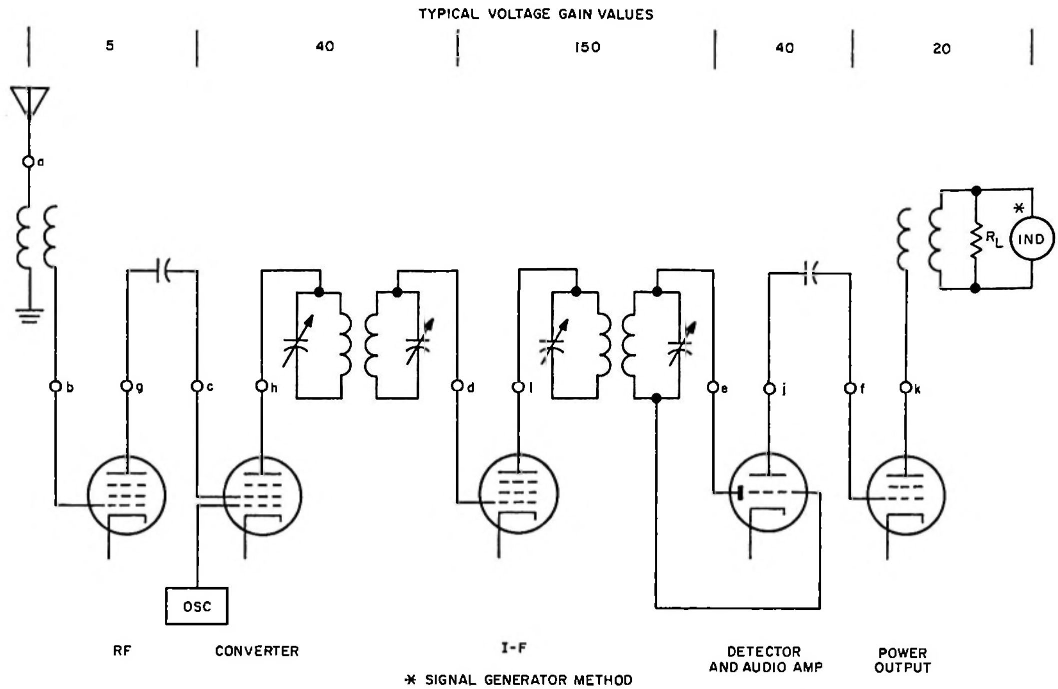


Figure 11-2. Receiver Stage Gain

quipment, such as the output circuit in the case of an am receiver. The regeneration, capacitance, stray inductance, etc, that may result from the connection of test equipment into high-impedance locations may in this way be avoided.

11-51. Figure 11-2 shows the gain for each stage of a simple receiver. If a suitable power meter is available, you can remove the normal output load and connect the meter directly across the secondary winding of the output transformer. If you use an ac voltmeter or a db meter, it may be connected across the normal output load, or a noninductive resistor (R_L), which is used in place of the normal output load. A signal generator modulated 30 percent at 400 cps, and set to the proper frequency at an output level that will not overload the receiver, is connected successively to points d and e (to measure the gain of the i-f amplifier), or to the input and output, respectively, of any other stage under measurement.

CAUTION

A blocking capacitor should be placed in series with the signal generator output when a signal is applied to high-voltage sources, such as the plate circuits of amplifier tubes. If this precaution is not taken, the signal generator output attenuator may be damaged.

11-52. Another method of measurement (one that is especially suitable for signal generators with outputs calibrated in db) is as follows: The generator output is adjusted so that equal receiver outputs are obtained when the generator is connected, for example, first to point a and then to point b. The difference between the generator settings indicates the gain of the stage (between a and b) directly in db. This method of testing is especially suitable for determining the gain between any two points in a receiver. The procedure is as follows: The automatic-volume-control lead is disconnected from the second detector, and connected to a fixed bias of a value equal to that obtained from the detector when a signal of a desired strength is received. The generator output is then applied successively to points a, b, c, etc, as shown in figure 11-2, which also shows gains typical of an a-m radio receiver.

11-53. When measuring the gain of the converter stage, several special considerations should be taken into account. First, it must be remembered that the output frequency of the stage is different from that of the applied signal. That is to say, the tuned frequency should be inserted at point c and the i-f frequency at point d. The resulting gain figure must still be regarded with caution, since a certain amount of oscillator current flows in the signal grid circuit. This current influences the actual conversion gain of the stage. A less misleading procedure would be to measure the combined gain of the converter and the immediately preceding rf stage of the receiver. In this case, the signal generator may be fed at point b, or, in the case of high frequencies, at the antenna coil (point a). In the latter case, the gain figure obtained will combine the apparent antenna coil gain and the conversion gain. At these high frequencies, the antenna coil gain will probably be low as a result of damping caused by the input impedance of the rf amplifier. The converter gain can then be determined by subtracting the rf stage gain from the combined gain figure. This method has the disadvantage that serious detuning, regeneration, or even oscillation may occur when test probes are connected to a high-frequency stage. Therefore, you must recognize any peculiar meter indications. Low-capacitance test probes should be used, and you must select a meter which has an adequate frequency range.

11-54. FM RECEIVERS. In the measurement of gain in an fm receiver, the presence of limiting is a complicating factor. An accurate method of measuring the gain of rf and i-f stages is to use auxiliary equipment. This equipment can incorporate an ordinary a-m detector followed by audio frequency amplification. A meter of some kind should be placed in the output circuit of the final amplifier. If the output of the final i-f stage is applied to the a-m detector rather than the limiter, the gain of any preceding stage can be determined in the manner already given for a-m receivers. For this method, of course, you must use an a-m signal generator.

11-55. The easiest method of measuring gain in most fm receivers is to determine the grid current of the first limiter stage by measuring the voltage drop across a resistor in the grid circuit. Another procedure is to connect an ac vacuum-tube voltmeter to measure the output of the final i-f stage, with the limiter tube removed from the receiver. However, this arrangement may cause the gain measurement of the final i-f stage to be inaccurate unless the input impedance of the meter is satisfactorily isolated from the stage. Measuring the gain of an af stage is accomplished by the same methods used for a-m receivers.

11-56. MINIMUM DISCERNIBLE SIGNAL MEASUREMENTS.

11-57. Minimum discernible signal (mds) measurements are measurements generally confined to pulse-type receivers; they provide an indication of receiver sensitivity. The mds measurement actually denotes the

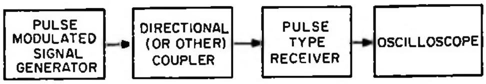


Figure 11-3. Minimum Discernible Signal Measurement

weakest signal that will produce a visible, or, in the case of transponders, a usable pulse output. Because the maximum possible sensitivity is dependent upon the amount of receiver noise, a minimum discernible signal measurement will preclude the necessity for making a noise figure measurement.

11-58. Figure 11-3 shows the block diagram of the mds test. The signal output of a calibrated, pulse-modulated signal generator is applied to the input coupler of the receiver under test. The signal generator pulse width and pulse repetition rate are adjusted to a suitable value for the receiver under test. You may then connect either the A scope component of the equipment, if included, or a synchroscope to the receiver output, and adjust the receiver gain control for a receiver noise level just below saturation. The power output of the pulse-modulated signal generator is then set for a power output of 1 milliwatt, or any other reference (such as the older standard of 6 milliwatts). The pulse-modulated signal generator attenuation is then increased until the pulse displayed on the indicator is at the point of disappearing. The attenuation resulting from the connecting cables and coupler is added to the attenuator dial reading, and this figure, after conversion to power, is employed in the following formula as P_2 to determine the mds power:

$$P_{mds}\text{ (db)} = 10 \log_{10} \frac{P_2}{P_1}$$

where:

P_{mds} = minimum discernible signal power

P_1 = pulse-modulated signal generator reference power

P_2 = total power attenuation

11-59. In certain types of equipments, such as transponders, the received signal is amplified and employed to key a transmitter (usually an integral part of the equipment). For this equipment it is sometimes convenient to measure the mds during normal operation by feeding in, from a pulse-modulated signal generator, a pulse of a power magnitude just sufficient to cause consistent operation of the transmitter. This value is then used during preventive and corrective maintenance procedures to determine whether the equipment is operating properly.

11-60. STANDING WAVE MEASUREMENTS.

11-61. Standing waves present on transmission lines and waveguides are indicative of poor or imperfect impedance match between a transmitter or receiver and its antenna. When this condition occurs, the transfer of energy between these units and their respective antennas becomes inefficient. For this reason, standing-wave measurement tests are included in the scheduled preventive maintenance procedures for equipment which is applicable to this type of test. A low standing-wave ratio is indicative of a properly matched system, and is a prerequisite for good performance of all communication and radar transmission lines which are intended to match the load and source impedance, such as long coaxial and waveguide transmission lines. Procedures for checking the current or voltage variations which are the components of the standing waves are dependent fundamentally upon the frequency of the system. Various methods for determination of the standing wave ratio are included in Section VI of Chapter 10.

11-62. FREQUENCY SPECTRUM MEASUREMENTS.

11-63. The amplitude of a group of frequencies, which make up a signal, plotted against frequency is termed the frequency spectrum of the signal. Various types of equipment provide a visual (or meter) indication of the spectrum. Those equipments which are employed for observing or measuring segments of the audio band are called wave analyzers, distortion analyzers, sound analyzers, etc; those employed for observing large segments of the radio-frequency spectrum are called panoramic adapters or signal analyzers and those employed for observation of the spectra of radio-frequency oscillators are called spectrum or pulse analyzers.

11-64. The wave analyzer is a laboratory-type instrument, and is used to determine the frequency components present in an audio-frequency wave. The panoramic adapter is used as a functional equipment for monitoring segments of the radio-frequency spectrum. Although it will provide relative information regarding signal strength, pulse information, etc, its use is generally confined to countermeasures and other monitoring purposes. Spectrum analyzers are most often employed for use with pulse-type equipment.

11-65. When a radio-frequency carrier wave is modulated, whether by audio intelligence or pulse signals, the resulting waveform incorporates the sideband components, both above and below the carrier frequency for conventional modulation, and on one side of the carrier frequency for single-sideband modulation. These sideband components occur at rf frequencies. A common conception of a pulsed output is a single frequency which is turned on and off for periods of standard duration (similar to that of a cw telegraph signal). The output of pulse transmitters does not consist of only a pulsed fundamental frequency, and, therefore, should be considered as a fundamental frequency that is modulated by the waveform of a pulse. The fundamental frequency with sideband frequencies is collectively called the spectrum. The spectrum analyzer resolves the rf signal into its frequency components and displays the results on the screen of a cathode-ray tube so as to show amplitude plotted against frequency. The use of the spectrum analyzer in the microwave region (the region in which radar operates) is especially valuable. Good performance of pulse-type transmitting equipments depends on the use of properly shaped pulses, to produce a transmitter output which is devoid of spurious frequency components. Good performance of these equipments also depends upon proper settings of the local oscillator and waveguide plumbing. Some of the performance measurements and checks that can be performed with the analyzer are as follows: observations of the operation of local oscillators with respect to their transmitters, checks of automatic frequency control operation, and measurements of frequency and power. Detailed information and specific applications pertaining to spectrum analysis are given under paragraph heading 11-722.

11-66. IMPEDANCE TESTING OF ANTENNAS AND TRANSMISSION LINES.

11-67. The amount of current that flows in an antenna is one of the most important factors affecting the performance of transmitter equipment. Thus, in order to secure the maximum radiated power from a transmitter of a given power, as much of the radio-frequency energy generated as possible must be efficiently transferred to the antenna. Also, for optimum reception, the maximum transfer of energy from the antenna to the receiver must occur. Efficient transmission and reception conditions prevail whenever the transmitter (or receiver) is properly matched to the transmission line and the transmission line is properly matched to the antenna.

11-68. Normally, the antenna and transmission lines are installed as an integral part of the equipment, and performance tests concerning impedance match consist primarily of taking standing-wave measurements. However, in certain instances, it will be found that an undesirably high standing-wave ratio will be obtained, caused by a change in antenna impedance. This could be the result of a new antenna installation, or the erection of a structure in the proximity of the antenna, so that the structure influences the antenna characteristics. In practice, the antenna matching metwork is generally varied to match the new antenna characteristics, since the transmission line is designed to match the equipment impedance. This can best be done by making a series of standing-wave-ratio checks and antenna-matching adjustments until an acceptable standing-wave ratio is reached. It must be understood, however, that the antenna does have a specific impedance at a given frequency, and that, when necessary, this impedance may be determined by use of an rf impedance bridge.

SECTION II

COMMUNICATIONS EQUIPMENT TESTING

11-69. GENERAL.

11-70. Efficient and accurate testing procedures are indispensable for reliable operation of communication equipment. You must use trouble-shooting, adjustment, and alignment techniques during corrective maintenance procedures. In addition, measurements should be made periodically, while the communication equipment is functioning properly, to disclose any gradual decline in the quality of equipment performance, enabling you to prevent many incipient failures.

11-71. Prior to making any measurements, you should check the equipment power source, eg, power-line voltage and frequency, battery condition, etc. The test equipment used should be of the type recommended by the equipment technical manual or other reliable source of information, such as Equipment Component List 665 or the manufacturer's instruction book. When specific test equipment is not listed in the above publications, or if the recommended equipment is not available, you may use Chapter 9, Volume VII, to determine which of the available test equipments is most suitable for the test to be made.

11-72. COMMUNICATIONS RECEIVER TESTING.

11-73. GENERAL.

11-74. Communications receivers are composed of a series of selective rf and af circuits, each stage of which is designed to amplify the output of the preceding stage. The lowered efficiency of any one tube, or a change in any one circuit parameter, usually results in lowered over-all efficiency of the receiver. The sensitivity of the receiver may also be decreased by the misalignment of the successive circuits, each of which may function in a suitable manner as a unit. The sole function of a communications receiver is to receive (selectively) a weak signal; therefore, an objective over-all test on sensitivity is the most significant single check that can be made on the condition of a receiver.

11-75. Some receivers are provided with a built-in output meter; others have an output meter equipped with a cord and plug to facilitate testing. The only other requirement for a sensitivity check is a standard signal for the excitation of receivers on the various bands. During radio silence this signal may be provided by a calibrated signal generator with a dummy antenna coupled directly to the receiver input. When it is permissible to operate transmitters, the output of the signal generator may be fed into a central radiating antenna at the station, and receivers may be calibrated from the signal received by their own antennas. Any decrease in sensitivity should be corrected by following the appropriate remedial procedures discussed in this section.

11-76. In addition to periodic checks on sensitivity, routine physical inspections must be made of the receiver and accessory units. Lubrication and cleaning schedules recommended by the manufacturer and outlined in the preventive maintenance cards must be followed. Tubes should be tested sparingly, because frequent insertion and removal weakens the socket contacts and causes noisy (or intermittent) operation. Electron-tube life generally extends to several thousand hours; therefore, in equipment which is continuously operated and on which periodic sensitivity tests (or operation records) are made, tubes should be checked only when poor performance indicates such a need. When tubes are replaced in rf circuits, the circuits should be realigned, if necessary, to achieve normal sensitivity. Before new tubes are used, it is good practice to check them on a transconductance-type tube tester.

11-77. Methods of adjustment and servicing each type of equipment are discussed in detail in the technical manual or manuals furnished with each equipment. You should proceed methodically in locating receiver faults, first testing the most accessible (or vulnerable) parts. Previous experience and trouble-shooting charts aid in isolating the trouble. While the receiver is on the test bench, it should be thoroughly cleaned and inspected so that parts close to failure may be detected beforehand. Since receivers could operate for many years with reduced sensitivity before a complete failure occurs, the preventive maintenance schedule for each receiver should be followed.

11-78. SENSITIVITY MEASUREMENTS.

11-79. GENERAL. Sensitivity measurements are convenient over-all measurements which provide quantitative information regarding receiver performance in the field. The sensitivity of a radio receiver can be defined as the input carrier voltage with standard modulation required to develop a standard value of test output.

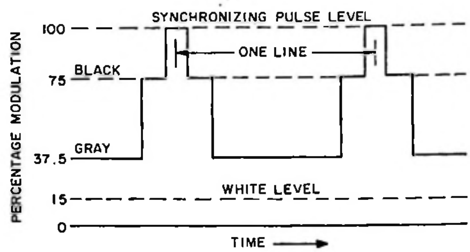


Figure 11-4. IRE Standard Waveform for the Sensitivity Measurement of Television Video Reception

11-80. The standard modulation for the following types of receivers, as established by the Institute of Radio Engineers, are:

Amplitude-modulation broadcast receivers: 400 cycles, 30 percent

Frequency-modulation broadcast receivers: 400 cycles, 22.5-kc frequency deviation

Television receivers (black and white 6 mc):
sound channel—400 cycles, 7.5-kc frequency deviation
video channel—waveform shown in figure 11-4 or 400 cycles, 30 percent amplitude-modulated

11-81. Sensitivity measurements require the application of an accurately calibrated signal to the antenna input terminals of a receiver, through an impedance network (dummy antenna) which approximates the impedance characteristics of the antenna with which the receiver is designed to be used. The dummy antenna simulates normal operating conditions and insures that the receiver has the proper impedance match, and that the signal current during testing is equivalent to the signal current obtained from a real signal of equivalent magnitude. The loudspeaker (or headset) is replaced by a suitable re-

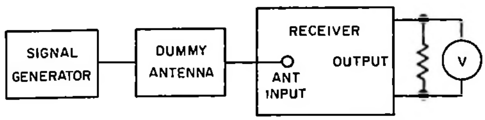


Figure 11-5. Typical Equipment Arrangement for Radio Receiver Testing

sistor (with the resistance equal to the normal load impedance at the frequency of modulation), or an indicating device is connected directly to the detector circuit and the power output measured. The sensitivity is then a measurement of the required input signal for a standard power output. Note that the output power may be measured at the receiver output or at the detector, since virtually no noise is generated in the audio stages. Figure 11-5 shows a typical test equipment arrangement for the measurement of sensitivity.

11-82. Sensitivity measurements, unlike noise-figure measurements which are a ratio of the signal-to-noise power ratio of an ideal receiver to the signal-to-noise power ratio of the receiver under test, are relative measurements which are arbitrarily calculated as the 30-percent, 400-cps, modulated voltage input required to raise the noise level of the receiver in volts by 20 db, and are measured in microvolts or in db below one volt. This arbitrary reference was selected because a change of 20 db represents a 10-time change in voltage. Therefore, the voltage developed by the receiver across a given resistor with zero-volt input to the receiver can be raised by increasing the receiver input in microvolts for a 10-time increase of voltage across the resistor. However, the voltages employed must be well below the value in which limiting occurs. The output signal level for various types of receivers varies with their function and will be discussed when the sensitivity measurements of the different receiver types are mentioned in the following paragraphs.

11-83. Sensitivity measurements of single-sideband (ssb) receivers are determined in a manner similar to that used for other amplitude-modulation equipment. However, certain considerations must be taken in account when performing measurements on this type of equipment. For example, the frequency stability requirements of single-sideband equipment operating in the high-frequency region are on the order of 0.2 to 2 parts per million. Frequency errors greater than 30 cps can cause voice transmissions to be unintelligible with certain signal-to-noise ratios. Intelligibility decreases with an increase in frequency error, even with a high signal-to-noise ratio. When frequency conditions are simulated for the sensitivity measurements, this same high degree of accuracy is still warranted. Therefore, proper measurements for single-sideband receivers should include the use of test equipment designed for use with this type of communication equipment, or test equipment with an accuracy equal to, or better than, the accuracy to be maintained in the receiver.

11-84. It should be noted that sensitivity (and selectivity) may be affected by alignment in all types of receivers. As receivers become more complex, alignment becomes more of a problem. In amplitude-modulation receivers using conventional double-sideband signals, improper alignment may result in the loss of weak signals, through loss of sensitivity, and the inability to select the desired signal. If the oscillator is shifted off frequency, dial error will be introduced, and tracking error produces a varying intermediate frequency, which results in loss of signal over portions of the frequency range of the receiver. In frequency-modulation receivers, the discriminator tuning becomes somewhat critical, and in phase-modulated receivers, phasing of the carrier must be correct, adding to the alignment problem. When automatic volume control and automatic frequency control are added to a receiver, proper alignment procedures should be followed, or serious errors may be introduced, further complicating alignment. In equipment employing crystals, as either reference generators, oscillators, or filters, the alignment must center around the crystals, since, for all practical purposes, the frequencies of crystals are not variable. All of these factors must also be considered in the alignment of single-sideband receivers. In addition, the oscillator frequencies must be precisely adjusted because demodulation is directly affected by the oscillator frequency and any associated error. Filters must also be considered, since they affect the bandpass and rejection of undesired frequencies. Because of the wide variations in circuitry between various models of receivers, the actual alignment procedures and specfications provided in the applicable technical manual must be closely followed before performing the sensitivity check on a receiver.

11-85. DUMMY ANTENNAS. In the 15 to 30, 000-kc range, a typical standard dummy antenna for a high-impedance-input receiver consists of a 20-microhenry inductor shunted by a series-connected, 400-picofarad capacitor and a suitable 400-ohm resistor, with the shunt combination in series with a 200-picofarad capacitor. A standard dummy antenna circuit is shown in figure 11-6. This unit should be enclosed within a properly designed grounded shield and used with a signal generator having a resistive output impedance not exceeding 50 ohms. This dummy antenna acts like a 200-picofarad capacitance at low frequencies, like a complex impedance in the 1-mc region, and like a 400-ohm resistance at frequencies of 2 to 30 mc. For the measurement of low-impedance-input receivers of 50 to 70 ohms nominal impedance, a signal generator with

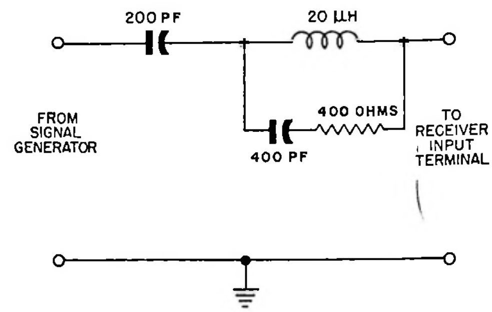


Figure 11-6. Standard Dummy Antenna Circuit

a 50-ohm output may be directly connected without the use of an external dummy antenna. Other generator impedances may require special dummy-antenna networks to load the generator and the receiver properly while allowing the equivalent induced antenna voltage to be accurately known.

11-86. CONDITIONS FOR SENSITIVITY MEASUREMENTS. For measurement of sensitivity, the receiver is adjusted for the type of reception desired, and circuits such as tone controls, audio filters, agc, silencer, noise limiter, etc, are placed in or out of operation, as required, or are set at the appropriate control positions, as discussed later. The power-line voltage applied to the receiver should be well within the normal recommended operating range. The receiver output terminals should be properly loaded. At the headphone or audio-line terminals, unless otherwise specified in the technical manual for the equipment, the load should be a 600-ohm noninductive resistor (such as one of the composition type), capable of continuously dissipating the maximum receiver audio power output that can be produced at these terminals. High-impedance headphones may be used in shunt with the load for monitoring the output. Low-impedance phones may load the output appreciably, and may have to be removed when measurements are being made.

The output voltage should be measured with a high-impedance audio-frequency voltmeter, capable of accurate indication from 0.1 volt to 100 volts, that will not appreciably load the output circuit. Although some receivers are equipped with audio-output meters, the meters provided may not indicate the required standard noise levels with sufficient accuracy.

11-87. CW AND FACSIMILE RECEIVER SENSITIVITY. For determination of cw (A-1) reception sensitivity, some means must be provided to set the output beat note of the receiver to a standard 1000-cps frequency with reasonable accuracy (about 1000 ± 50 cps). In some receivers, the 1000-cps "sharp" audio filter provided has a bandwidth narrow enough to allow satisfactory adjustment of the beat note by centering the tone in the pass band. The 1000-cps internal tone modulation frequency of certain signal generators is also accurate enough, and can be zero-beat against the output beat note. Alternatively, the output of a calibrated audio oscillator and that of the receiver may be fed independently to the deflection amplifiers of an oscilloscope to give the circular or elliptical Lissajous pattern characteristic of identical frequency.

11-88. For determination of both keyed cw and facsimile (A-1 and A-4) reception sensitivity, you should set the cw (beat-frequency) oscillator to <u>on</u> and the receiver audio gain at maximum, and the agc, silencer, noise limiter, and output limiter to <u>off</u>. If not otherwise specified in the receiver technical manual, you should set audio filters or tone controls for maximum audio range. The antenna trimmer normally should be peaked at the high-frequency end of each band, and not reset at other frequencies. The signal generator is used unmodulated. Following these initial adjustments, the following settings are typical: The rf gain control is adjusted to produce 60 microwatts of noise at the receiver output (0.19 volt across 600 ohms) with the receiver tuned to the desired frequency, but with no input signal applied from the signal generator. The signal (carrier only) is then applied, and is tuned as nearly as possible to center on the noise of the over-all rf pass band of the receiver, with the cw oscillator frequency control adjusted to the side of zero beat that produces the higher output with a 1000-cps beat note. The input-signal voltage is then adjusted to produce a 6-milliwatt output (1.9 volts), resulting in a +20-db output signal-to-noise ratio. The receiver sensitivity, in terms of input-signal voltage, is then read from the signal-generator voltage calibration. Other reference levels may be used, but the 20-db voltage relationship should be maintained.

11-89. A-M RECEIVER SENSITIVITY. For determination of voice-modulation (A-3) reception sensitivity, a carrier modulated 30 percent at 400 cps must be applied. The rf gain control should be set at maximum, with agc <u>on</u> and the cw (beat-frequency) oscillator <u>off</u> (this condition may be automatically established by the reception selector control provided on some receivers). All other controls except the af gain control should be set as indicated for cw (A-1) reception. Typical settings are as follows: Both the input-signal level and the af gain control are progressively adjusted until the receiver output noise level is 0.6 milliwatt (0.6 volt across 600 ohms) with signal-generator modulation <u>off</u>, and the signal-plus-noise output is 6 milliwatts (1.9 volts across 600 ohms) with modulation <u>on</u>, which produces a +10-db ratio of output signal-plus-noise to noise (10.4 db signal-to-noise ratio). The receiver sensitivity, in terms of input voltage, is then read from the signal-generator voltage calibration.

11-90. MCW RECEIVER SENSITIVITY. For tone-modulation (A-2) sensitivity measurements, the carrier should be 100-percent modulated with a 1000-cycle tone. The receiver af gain control should be set at maximum with the agc control off. Then, with the generator modulation off, the receiver rf gain control should be adjusted for a noise output of 60 microwatts (0.19 volt across 600 ohms). The generator modulation should be turned on and the generator output varied until a signal-plus-noise output of 6 milliwatts (0.6-volt across 600 ohms) is obtained (20-db output signal-to-noise ratio).

11-91. If the available signal generators cannot be used modulated 100 percent, because of excessive frequency modulation or other limitations, an approximate sensitivity measurement may be made by employing 30-percent modulation to produce a 6-milliwatt output with a 10-db output signal-plus-noise to noise ratio. This procedure may give somewhat erroneous results, as detector modulation distortion or modulation clipping by built-in noise limiters may be much less at 30-percent than at 100-percent modulation of the carrier.

11-92. FSK RECEIVER SENSITIVITY. The receiver, frequency-shift keying (fsk) converter, and teletypewriter must all operate satisfactorily to produce proper copy in fsk operation. If the receiver checks satisfactorily for cw (A-1) sensitivity, only the additional switching for fsk reception and any special fsk filters in the receiver could produce poor fsk operation, so far as the receiver proper is concerned. Therefore, the receiver may be checked for fsk sensitivity by initially checking its standard cw sensitivity. If this proves to be normal, switching to fsk operation will allow the output beat frequencies and audio output level to be checked, to insure that they meet the requirements of whatever audio-frequency-type fsk converter is employed.

11-93. The output which the receiver can produce for an i-f-type converter (if this facility is provided) may be checked with an electronic voltmeter, capable of good accuracy at the intermediate frequency, and with a range of 0.001 volt to at least 10 volts. The receiver and converter technical manuals should be consulted for standards of receiver output in this case.

11-94. FM RECEIVER SENSITIVITY. The procedures for measurement of fm (F-3) receiver sensitivity are analogous to those for a-m receivers. However, an fm signal generator must be used. For broadcast receivers, the signal generator should be modulated 30 percent of 75 kc maximum frequency deviation (22.5 kc) at 400 cps.

11-95. Because noise is predominantly amplitude-modulated and the limiting action inherent in fm receivers tends to reduce the noise level, a convenient performance measurement is "quieting-signal sensitivity". The quieting-signal sensitivity is the smallest unmodulated carrier voltage that will provide an output 30 decibels below the output that would be obtained from the standard modulation (30 percent of maximum frequency deviation). The test is made by applying a signal of mean value, 30-percent-modulated at 400 cps, to the receiver with the volume control adjusted to give a convenient output below audio distortion. The modulation should then be switched off and the signal intensity reduced to the least value which will produce a 30-db rise in the indicated output with standard test modulation, as compared with the indicated output with the unmodulated carrier. The results are generally expressed in microvolts for 30-db quieting, or in decibels below 1 watt.

11-96. PULSE-MODULATION RECEIVER SENSITIVITY. Continuous-wave generator methods of measuring sensitivity do not provide an accurate indication of the ability of a receiver which is designed for the reception of pulse-modulated signals to receive weak pulse transmissions. A better method of determining the sensitivity of a pulse-modulation receiver is by performing a minimum-discernible-signal measurement. This type of measurement consists of measuring the power level of a pulse whose level is just sufficient to produce a visible receiver output (as described in paragraph 11-637). However, because of the relatively wide bandwidths associated with pulse-modulation receivers, a still better performance indication can be obtained by determining quantitatively how much noise is inherent in the receiver, since noise is the limiting factor in the determination of maximum sensitivity. This method of checking sensitivity utilizes a noise generator, as the most desirable test equipment, for a signal source. The noise in the receiver is related to a calculable noise figure (see paragraph 11-38).

11-97. RESERVE GAIN. Reserve gain for all types of reception may be determined by measuring the ratio of noise output at standard gain (the gain condition used in measuring standard sensitivity) to noise output at maximum gain, provided that maximum gain does not produce any substantial degree of output overload or saturation. If saturation is approached or reached at maximum gain, the setting for standard gain should be noted, and the reserve gain determined with the aid of a gain control calibration curve, which can be obtained by other measurements.

11-98. GAIN VARIATION WITH CHANGE OF FREQUENCY. Gain variation over each band for any condition of reception may be determined by adjusting for standard gain (as for sensitivity measurements) at the high-frequency end of each band, and then noting the input-signal voltage required at various frequencies over the band to produce the same 6-milliwatt, 400-cps output.

11-99. DETERMINATION OF RESPONSE CURVE. A graph which shows the response curve of a receiver or amplifier can be obtained by plotting frequency horizontally from left to right and response amplitude

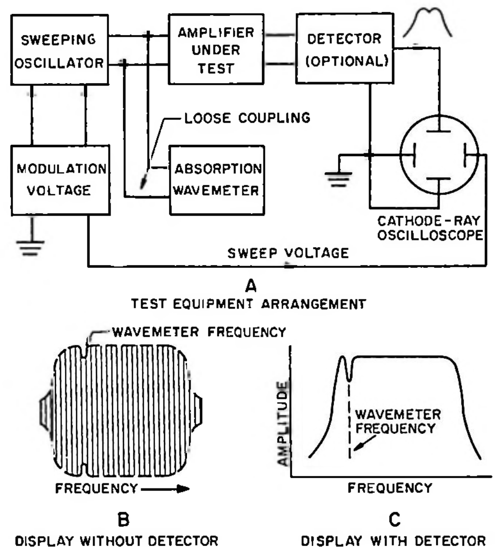


Figure 11-7. Equipment Arrangement To Obtain a Visual Response Curve

vertically, if a sufficient number of points are determined. However, such a method of obtaining a response curve might prove to be laborious and time-consuming; with such a method, alignment for a specified response curve, when repeated curves are necessary, would be a rather drawn-out process.

11-100. A visual response curve may be produced by applying an fm signal to the receiver circuit being checked, and employing an oscilloscope as an output indicator; figure 11-7 shows such an equipment arrangement. The fm signal varies sinusoidally with frequency, usually at a 60-cps rate, about a center frequency. The amount of frequency variation (deviation) about the center frequency is determined by the sweep-width or deviation control of the generator. For a total deviation of 6 mc, the frequency varies both above and below the center frequency by 3 mc. When this changing frequency is applied to the circuit under test, the instantaneous output amplitude is always pro-

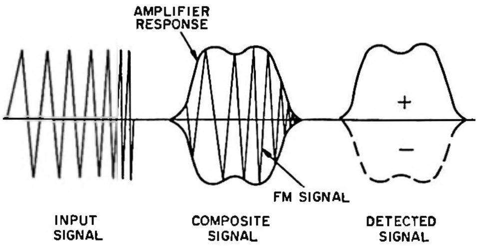


Figure 11-8. Visual Response Curve Composite Waveform

portional to the response of the circuit to the frequency at that instant. Thus, the original frequency-modulated input signal is changed in passing through the amplifier, so that the output signal from the amplifier consists of an fm signal which is also amplitude-modulated (see figure 11-8). Since, for equal deviations, the positive and negative portions of this envelope are symmetrical, it is necessary to observe only one side of the envelope. The signal is also varying at an rf rate, and cannot be seen on the oscilloscope without first being detected. The composite signal is therefore detected by either the second detector or the discriminator of the receiver. After detection, only the modulation remains, and this appears on the oscilloscope in the form of the response curve of the amplifier. Note that the detector polarity determines whether a positive or negative output occurs; the response curve will appear inverted if the output is of negative polarity. However, the response curve may be used in whichever position it appears at the output of the detector.

11-101. To determine the correct deviation, it is important to remember that a response curve can be more easily checked when it covers a large portion rather than a small portion of the oscilloscope screen; therefore, the generator deviation (sweep width) control should be operated at a position which produces only enough deviation to accommodate the pass band of the circuit being checked. It is poor practice to use maximum deviation at all times, or to use a sweep width much greater than required, except for some special objective. For example, a-m circuits should be operated with about 20-kc deviation, fm circuits with about 1-mc deviation, television i-f circuits with about 10-mc deviation, and broad rf circuits with a broad pass band with full deviation.

11-102. While the response curve shows the frequency discrimination of the circuit under test, some means of determining the actual frequency at any definite point along the curve is necessary. This can be accomplished by use of an absorption-type wave meter loosely coupled to the generator output, which causes a reduction of amplitude at the point of resonance, thereby placing a notch in the curve at the point at which the wave meter is tuned. A separate marker generator can also be used to furnish a cw signal which is combined with the fm sweep output signal. At the point on the response curve corresponding to the frequency to which the marker generator is tuned, a pip is produced which is the result of the beat note between the marker generator and the sweeping generator at that frequency.

11-103. SELECTIVITY AND BANDWIDTH MEASUREMENTS.

11-104. GENERAL. Selectivity is the property which enables a receiver to discriminate against transmissions other than the one to which it is tuned. It is usually expressed in the form of a curve obtained from a plot of the strength of a standard modulated-carrier signal that is required to produce a constant (standard) output, versus off-resonance frequency. Figure 11-9 shows a typical selectivity curve with the carrier signal strength at resonance used as a reference.

11-105. The bandwidth of a receiver is usually used to define that portion of the selec-

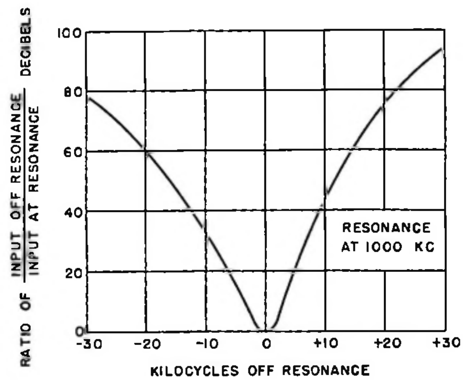


Figure 11-9. Selectivity Curve of Typical A-M Receiver

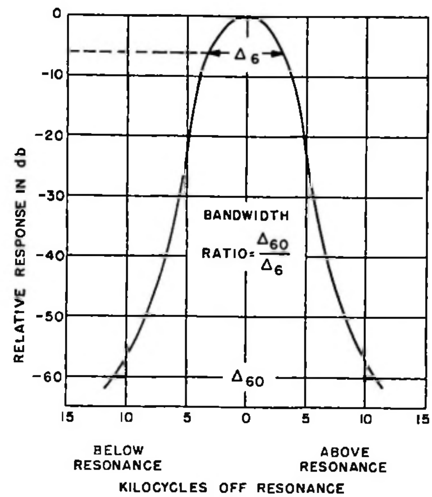


Figure 11-10. Receiver Response Curve

tivity curve that represents the frequency range over which the amplification is relatively constant. For most receivers, the bandwidth represents the usable portion of the curve, and has a direct relation to the fidelity of the modulated intelligence. Practically, the bandwidth is measured at the half-power (3-db-down) points, or, for certain applications, at the 6-db-down points, and is represented by the frequency range between the two points on a response curve expressed as relative response in db versus frequency, as shown in figure 11-10. However, the bandwidth at the 3-db (or often the 6-db) points, when compared with the bandwidth at the 60-db-down points, gives a good indication of the selectivity of the receiver, since the character of the skirts of the curve becomes apparent. This comparison is referred to as the bandwidth, or selectivity ratio. In most receivers, the over-all bandwidth is determined by the i-f amplifiers; therefore, bandwidth is sometimes considered as fundamentally an i-f characteristic measurement.

11-106. OVER-ALL SELECTIVITY. Since the rf stages of a receiver are also of some importance in determining the selectivity, and of fundamental importance in determining the image rejection characteristics, the selectivity is most often plotted as over-all selectivity. The term over-all selectivity usually refers to the frequency selectivity of a receiver as measured from (and including) the antenna to the input terminals of the final detector. It does not normally include any elements of the audio system. The over-all selectivity of a superheterodyne receiver may be quite difficult to measure accurately with the equipment available in most operating installations, especially at frequencies above 1 mc. If the lowest signal frequency is at least several times that of the lowest intermediate frequency used in the receiver, the over-all selectivity is very likely to be practically the same as the lowest i-f selectivity. In such instances, the i-f selectivity curve may suffice, and is much easier to measure.

11-107. BANDWIDTH. Bandwidth and sensitivity curves can be obtained with similar test equipment and connections. Figure 11-5 shows one method of test for audio-modulation receivers. For unmodulated test of cw receivers, or narrow-band (less than 5 kc at the 6-db-down points) receivers, a high-impedance voltmeter should be connected across the second-detector load for the output indicator.

11-108. When making bandwidth measurements, the receiver avc should be disabled (grounded), connected to a source of fixed bias, or turned off, and the volume control set to maximum. The signal generator is set at the receiver frequency, and, if applicable, modulated 30 percent at 400 cps. The output of the generator is set to a value which will place the receiver well below the point of overload (no limiting action), and the receiver rf gain is adjusted for a convenient reference receiver-output voltage. The signal generator output is then increased by 3 db (1.4 times the voltage of the original setting) for measurement at 3-db-down points, or 6 db (2 times the voltage of the original setting) at the 6-db-down points. The generator frequency is then varied to one side of resonance until the receiver output gain indicates the reference level; the procedure is then repeated on the other side of resonance. The bandwidth is the total frequency displacement between the 3-db points (or 6-db-down points), whichever is applicable.

11-109. NARROW-BAND RECEIVER SELECTIVITY. For a receiver with an rf or i-f bandwidth of less than about 5 kc at 6 db down (eg, vlf and lf receivers), the measurement can be made as follows: The receiver should be adjusted for standard mcw (A-2) sensitivity conditions (agc off), with a high-impedance dc voltmeter connected to read the voltage across the final-detector diode load. (It may be necessary to connect a 1-megohm isolating resistor between the "high" end of the diode load and the "high" lead of the voltmeter, to prevent regeneration or other undesirable effects.) The signal generator is used unmodulated, and the signal voltage is increased from the standard input in steps of about 1.4, 2, 3, 5, 10, 100, and 1000 times the standard input, in turn. At each step, the frequency of the signal is adjusted to produce the same detector diode voltage as previously obtained with the standard input at resonance. The signal-generator frequency-vernier dial reading is taken for each step at both sides of resonance. The reading recorded should always be obtained by approaching from the same direction of dial rotation for all readings, to minimize error resulting from signal-generator dial backlash. The signal-generator frequency dial can be calibrated with higher precision than is afforded by its markings in the range covered, by first noting the vernier-dial settings for the two frequency-dial markings nearest the limit frequencies of the selectivity curve data, and then dividing the difference in frequency by the corresponding number of vernier divisions to obtain kc per division. A curve can be plotted of "times resonant input voltage" on semi-log paper (or "db above resonant input" on linear paper) against "kc off resonance" on a linear scale. The points of greatest interest are those defining the 3-db-down, 6-db-down, and 60-db-down bandwidths, which also determine the 60/6-db bandwidth ratio (or selectivity ratio).

11-110. WIDE-BAND RECEIVER SELECTIVITY. With receivers having 5 kc or more bandwidth at 6 db down on the selectivity curve, selectivity measurements may be made by use of a modulated carrier, provided that the rate of attenuation at the skirts of the curve is not too high (not more than about 6 db per kc). This means that, in general, trf and single-conversion superheterodyne receivers designed for operation above 500 kc may be measured with a carrier modulated 30 percent by a 400- or 1000-cps tone. The procedure is the same as for unmodulated carrier selectivity measurements, with the same receiver conditions, except that the output measurement is made at the audio output terminals of the receiver, just as for mcw (A-2) sensitivity (agc off). The audio output is kept at a standard level, as the input-signal frequency is varied, by adjustment of the signal-generator carrier output voltage.

11-111. I-F SELECTIVITY. The selectivity of i-f amplifiers may be measured in the same manner as over-all selectivity, except that it may be desirable to disconnect the input-signal circuit from the preceding frequency converter (mixer) to prevent that circuit from loading the signal generator. It will then usually be necessary to provide a grid-return resistor for the mixer (about 10,000 ohms), as well as a coupling capacitor (about 1000 pf) from the mixer input grid to the signal-generator output, in order to prevent dc return through the generator system. The oscillator should be disabled, by removing the oscillator tube or its supply voltages, if it appears to produce interference with the measuring signal. If this is done, the mixer output impedance will be changed somewhat, producing some change in i-f selectivity; however, the change is usually negligible.

11-112. FM RECEIVER SELECTIVITY. Selectivity measurements on fm receivers are somewhat complicated by the limiting action inherent in fm equipment, since once the incoming signal attains a certain amplitude, further increases of signal strength do not affect the power output of the receiver. Therefore, limiting action invalidates the procedure of measuring the selectivity of rf and i-f stages by the use of a meter connected in the output of the receiver. (Conceivably, the signals applied to the fm receiver could be made so small that limiting action does not occur, but noise considerations make this procedure impracticable.) The procedure for making selectivity measurements on fm equipment incorporating the grid-leak limiter is as follows: A vacuum-tube voltmeter is connected across the grid-leak resistor, or an ammeter is connected in series with the resistor, in order to indicate the receiver output. With no signal applied, the meter reading is noted. This reading corresponds to the noise generated by the stages preceding the limiter. The fm signal generator is next set up to the receiver frequency and adjusted for a suitable frequency deviation, usually 30 percent of 75 kc maximum frequency deviation (22.5 kc) at 400 cps for broadcast receivers. The generator input is increased until the reading on the vacuum-tube voltmeter or ammeter is appreciably greater than the noise level. For an average receiver, a signal of 50 microvolts or less should be satisfactory; for a sensitive receiver, a 5-microvolt signal should be adequate. The reading of the meter under this condition is the reference value to be observed throughout the remainder of the measurement. The signal generator is now detuned by a suitable amount from the resonant frequency, and its output is increased until the meter in the limiter circuit indicates the reference value. This is repeated at frequencies further removed from resonance. The ratios of the amplitudes of the signals at the off-resonant frequencies to the amplitude at resonance are plotted against frequency, as is done for a-m receivers.

11-113. For receivers incorporating detectors other than discriminator types (usually not containing limiters in which the limiter current could be measured), it is necessary to improvise a simple detector circuit external to the receiver. An acceptable arrangement is to connect a high resistance across a germanium diode, such as a 1N34, and then to couple the positive side of the diode to the secondary of the final i-f stage by means of a capacitor. The negative side of the diode must be connected to the low side of the secondary. The values are not critical; the resistor is usually between 0.5 and 5 megohms, and the capacitor between 100 and 300 pf. If a vacuum-tube voltmeter is connected across the resistor, or a milliammeter is connected in series with the resistor, the procedure is the same as that

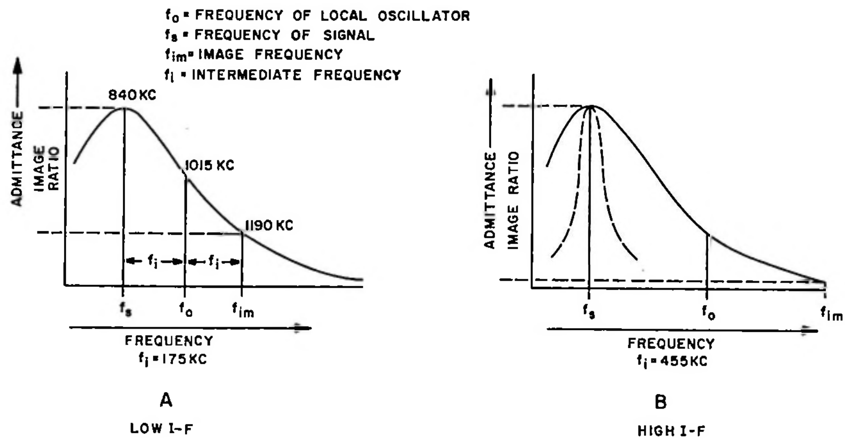


Figure 11-11. Image-Frequency Response of a Superheterodyne Receiver

described for receivers incorporating grid-leak detectors.

11-114. IMAGE RESPONSE. A superheterodyne receiver is responsive to a frequency which differs from the desired signal by twice the intermediate frequency. A signal at this unwanted frequency is called an image. The image rejection ability of the receiver is a direct function of the input selectivity of the receiver; in general, the image rejection capability is proportional to the image ratio. The image ratio is defined as the ratio of input voltage that produces a standard output at the image frequency to the input voltage required at the resonant frequency of the receiver to produce the same output. Figure 11-11 illustrates the image-frequency response in a superheterodyne receiver. Figure 11-11A illustrates a receiver tuned to 840 kc, with the local oscillator operating at a frequency of 175 kc above the tuned signal. The image frequency, 1190 kc, may not be entirely suppressed. Figure 11-11B shows the curve of a similar receiver tuned to 840 kc, but with the local oscillator operating at 1295 kc. In this case the image signal is at 1750 kc, which results in a more favorable image ratio and in greater image suppression. For the described cases, the greater separation between signal and image frequency (350 kc and 910 kc, respectively) provides a greater image ratio.

11-115. PRIMARY IMAGE REJECTION. In order to perform a test for image rejection, it is necessary to know whether the local oscillator is above or below the incoming signal. If above, the image frequency is higher than the incoming signal by twice the i-f frequency; if below, it is lower than the incoming frequency by twice the i-f frequency. The image ratio is determined by finding the sensitivity of the receiver at the image frequency and dividing this figure by the sensitivity at the resonant frequency. The primary image rejection ratio provides a simple criterion of the preselector (front-end) alignment condition. To determine the primary image ratio, the receiver is first adjusted for standard mcw (A-2) sensitivity conditions. The signal generator is then tuned to produce maximum response at the primary image frequency (twice the i-f away from resonance, on the same side as the oscillator), and the input-signal voltage is adjusted to produce standard output. The ratio of the image input voltage to the standard sensitivity input (usually expressed in db) is the image rejection at that desired signal frequency. In a good receiver the image ratio will be 60 db or more. However, if the values obtained for this ratio over each band are within 3 db of the technical manual values, and the sensitivity is normal, the front-end alignment is probably good.

11-116. CROSS-MODULATION TEST. A cross-modulation test is, in a sense, a selectivity test. However, it has the distinct advantage of determining the effect of "interfering" signals with the automatic volume and gain controls in operation, which is the situation that will normally be encountered during receiver operation. This test requires the use of two signal generators. One signal is set at the frequency to which the receiver is tuned, while the other is tuned away from resonance in order to represent the interfering signal. This technique has the advantage of closely approximating the condition that exists when the receiver is tuned to a station close to the frequency of a strong local signal. Initially, the desired signal is adjusted so that a suitable output is produced. A common procedure is to set the desired signal at a standard input voltage, such as 50 microvolts or 5 millivolts (depending upon the receiver sensitivity), and to adjust the volume control of the receiver for a standard output, making certain not to overload the receiver. After the output is noted, the modulation of the desired signal is removed. Then the other generator is turned on. It is set up to provide a signal modulated 30 percent at 400 cps, and is tuned through various radio frequencies on either side of the desired signal. Usually, increments of 10 kc or 1 mc are selected, depending upon the band in which the receiver operates. In each case, the amplitude of the interfering signal is increased until the standard output is once more produced by the receiver. From the data obtained, a curve can be plotted in the manner described for selectivity in paragraph 11-109.

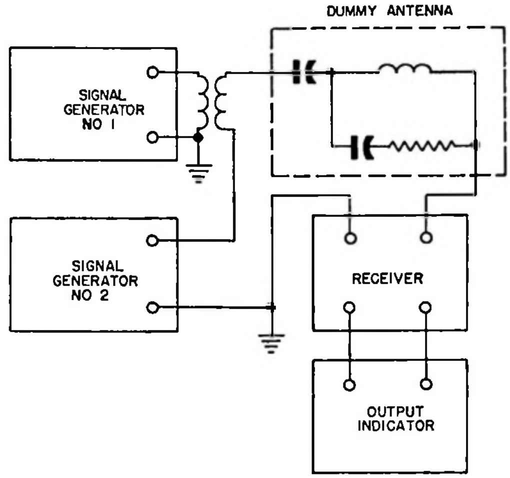


Figure 11-12. Cross-Modulation Test

11-117. Figure 11-12 shows one method of coupling the two generators to the dummy antenna of the receiver. The only restriction of the coupling transformer is that the inductance of the secondary winding must be small as compared with the inductance of the dummy antenna. A one-to-one voltage transfer is frequently employed. It is also possible to utilize the secondary winding as the usual 20-microhenry inductance of the dummy antenna, as shown in figure 11-6, provided that loose coupling (less than 30 percent) exists between the primary and the secondary. This precaution minimizes the effect of changes in the constants of the dummy antenna. The reactance of the primary winding must be approximately equal to, or greater than, the output impedance of the generator unless only small signals are required.

11-118. BLOCKING INTERFERENCE TEST. The cross-modulation test described in paragraph 11-116 gives no information concerning the attenuation of a desired signal in the presence of a strong interfering signal. This type of apparent signal loss occurs when the combined strength of the desired signal and the interference is sufficient to cause overloading of the receiver, or, in the case of a poorly designed avc circuit, to cause this circuit to reduce the output of the receiver. In order to determine whether the blocking effect exists, perform the cross-modulation test described in paragraph 11-116, and record the amplitude of the input signal from signal generator No. 2 at each frequency in the test; then turn off the modulation of the generator. Next, turn on the modulation of generator No. 1, and measure the output of the receiver. Plot these values of output against frequency on the cross-modulation graph. If straight lines are obtained rather than curved lines, no blocking is present.

11-119. MODULATION DISTORTION MEASUREMENTS.

11-120. The distortion produced in the radio-frequency, intermediate-frequency, and detector stages of a receiver can increase significantly as a result of increases in the percentage of modulation. For example, the distortion generated by a square-law detector becomes prohibitive at percentages of modulation greater than about 50 percent, and for this reason such detectors are rarely employed in modern communications receivers.

11-121. One method for determining the distortion of a receiver in terms of the percentage of modulation is to connect a signal generator, by means of the dummy antenna shown in figure 11-6, to the receiver input, and to connect a suitable resistor across the receiver output, as shown in figure 11-5. A distortion meter should be connected across the resistor. The distortion meter will not respond to the resonant frequency, which is suppressed, but will provide the rms value of the other components of the distorted output signal, and thereby provide an indication of the amounts of distortion for calibrated percentages of modulation. The signal generator should be modulated at 400 cps and set for an output on the order of 50 microvolts. The receiver volume control should be adjusted for a low-level output on the order of 50 milliwatts, and this level should be maintained throughout the test. The reason for maintaining this low power output level is to keep the distortion contributed by the audio section to a low, constant level. The percentage of modulation at the generator is then increased in convenient steps from 10 to 100 percent, and the results plotted on linear graph paper, with the modulation percentage appearing horizontally and the values of distortion vertically. This test should then be repeated for different rf gain settings to determine whether the rf and i-f amplifiers affect the modulation.

11-122. TUNING DIAL CALIBRATION.

11-123. GENERAL. Tuning dial frequency calibration can be checked against any signal whose frequency is accurately known, such as that of radio station WWV and standard a-m, fm, and tv broadcast stations. Some receivers have built-in crystal calibrators, which give signals spaced throughout the working range of the receiver. If none of these signals is available, it will be necessary to obtain a heterodyne frequency meter such as Frequency Meter Set SCR-211-( ), the output of which may be fed to the receiver antenna terminals and tuned in as an unmodulated cw signal. When using signal sources having relatively high output levels, such as Frequency Meter Set SCR-211-( ), you must ascertain that the receiver is not tuned to receive the test signal at the image frequency. With any of these means, the tuning dial error should be carefully observed. If the error is excessive (more than about ±1 percent) and shows a definite progression with frequency, the receiver may require realignment.

11-124. TUNING DIAL BACKLASH. Tuning dial backlash is usually best determined in the cw (A-1) reception condition, by first tuning in a cw signal to receiver resonance with rotation of the dial in one direction, and adjusting the beat-frequency oscillator, on the side of zero beat that gives the greater output, until a 1000-cps beat note (accurate within ± 5 cps) is obtained. The vernier tuning dial (if provided) is then read as accurately as possible, or the tuning knob is marked in some suitable way. Following this, the signal is again tuned in, approaching it this time from the opposite direction of dial rotation, until the same 1000-cps indication is obtained on the same side of zero beat as before. The difference in vernier-dial divisions or in angular position of the tuning knob is the backlash.

11-125. The zero-beat output from the receiver might be used as a reference for this measurement instead of 1000 cps, which would allow dispensing with the oscilloscope and audio oscillator. However, the audio response of most receivers at very low frequencies is not very good, which usually makes accurate determination of zero beat by ear or output meter quite difficult.

11-126. RESONANT OVERLOAD MEASUREMENTS.

11-127. The resonant overload characteristic (desired output vs signal carrier input voltage) should be determined with the receiver adjusted for standard sensitivity conditions for the type of reception desired (A-1, A-2, etc). The receiver output voltage for increasing values of signal-generator input from 0.1 microvolt to maximum is recorded. In addition, in mcw operation, the output noise level at each input-signal level with modulation off but carrier on should be recorded. These readings, when plotted on log-log paper, can be interpreted to indicate linearity of receiver gain, residual hum, hum modulation, etc. If resonant overload curves are obtained for different audio-gain-control (agc on) and output-limiter-control (agc off) settings, respectively, the capabilities of the agc circuit and of the limiter may be determined (for steady-state signal conditions). If resonant overload curves are obtained at different silencer or squelch-control settings over the working range, silencer characteristics and effectiveness of operation may be determined. These may be compared with technical manual data on these characteristics.

11-128. FREQUENCY STABILITY MEASUREMENTS.

11-129. OSCILLATOR FREQUENCY DRIFT MEASUREMENTS. The use of a converter tube in a superheterodyne receiver in place of a separate mixer and local oscillator often presents a problem of oscillator stability. This consideration is especially important in the case of short-wave receivers, whether the tube is a pentagrid, triode-hexode, or heptode. One cause of frequency instability is the effect of electrostatic or space-charge coupling between the two sections of the tube, which increases considerably with an increase in frequency. There are, however, additional causes of oscillator frequency instability. The most important is variation of the signal-grid bias voltage. Such variation occurs continually in response to changes in the avc voltage. Variations of the anode and screen-grid voltages are additional causes of frequency drift. These variations are usually due to fluctuations of the power-supply voltage, which reflects any change in the filter output resulting from a fluctuating current drain. The last condition explains the "flutter" effects encountered in short-wave reception.

11-130. Generally, it is inconvenient to investigate separately the effect of grid-bias variations, since the plate and screen voltages would have to be obtained from a well-regulated supply. The usual procedure,

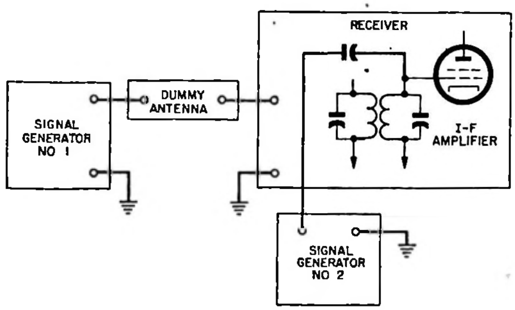


Figure 11-13. Oscillator Frequency Drift Measurement, Equipment Arrangement

therefore, is to conduct a combined test by operating the receiver in the normal manner. The signal input is varied in suitable intervals from 10 microvolts to 0.1 volt, and at each value the oscillator frequency shift is measured. In general, it is advisable to perform the test on all short-wave bands, but at the high-frequency end of the highest frequency band, the check is essential.

11-131. THE SIGNAL GENERATOR METHOD. As shown in figure 11-13, two standard signal generators are used; one generator is coupled to the receiver in the normal manner through a dummy antenna. The other generator is coupled loosely to the grid of the first i-f amplifier by means of an extremely small capacitor (about 5 pf). This precaution minimizes loading effects.

11-132. To perform the measurement, first adjust signal generator No. 1 to the frequency at which the check is to be made. This should be at the high end of the band. Set the output of this signal generator to 10 microvolts. Then tune signal generator No. 2 to approximately the intermediate frequency. A beat note should be heard in the output of the receiver. Carefully adjust generator No. 2 for zero beat, so that it supplies a signal at the intermediate frequency of the receiver.

11-133. Increase the output of signal generator No. 1 from 10 microvolts to 0.1 volt in convenient steps. At each input voltage, retune signal generator No. 2 for zero beat, if necessary. Record the frequency difference between the new setting and the original setting. This value is the drift, and represents the amount by which the oscillator frequency has changed. You can then draw a curve of the results on semilog paper, plotting the input voltages horizontally on the logarithmic scale, and the frequency changes vertically on the linear scale.

11-134. In the test described, the gradual increase in signal strength causes the avc voltage to increase also. This action not only changes the bias on the converter grid, but, in addition, lowers the current drain of the other tubes in the avc circuit, thereby affecting the voltage of the power supply. Consequently, the voltages applied to the screen and anode of the converter are also changed. The test, therefore, shows the composite effect of all three conditions influencing frequency stability.

11-135. If desired, a similar test can be made of oscillator instability resulting from line-voltage variation. An acceptable procedure is to vary the line voltage 10 percent on either side of the normal value, and to measure the frequency drift by the method described previously. If the heaters of the tubes are operated from a constant voltage source, the test can be made to show more tangibly the effect of plate-supply variations on frequency stability.

11-136. Frequency-drift measurements of frequency-modulation receivers are generally unnecessary because of the automatic-frequency-control circuits that are usually incorporated in these receivers. However, when it is desirable to make an afc circuit check, the oscillator drift may be performed as described under paragraph heading 11-157.

11-137. MECHANICAL EFFECTS. The effect, on a received signal, of mechanical stress caused by inclination, shock, and vibration is of particular importance to equipment used for mobile applications. You can determine whether mechanical stress causes excessive frequency instability by checking receiver performance while the receiver is subjected to these conditions during use, or by applying pressure on various surfaces of the equipment and jarring the equipment during the bench tests. In either case, the effect of mechanical displacements of the receiver on the pitch of the output beat note at various receiver input frequencies should be checked under the conditions of cw (A-1) reception. Relative comparisons can be made between different receivers of the same model, by this procedure, if you suspect that a defect has developed in one of them. You may also use this method to detect microphonic elements in communication equipment.

11-138. WARM-UP FREQUENCY DRIFT. Frequency changes which occur while the receiver equipment is warming up may be measured by setting a Frequency Meter Set SCR-211-( ), or similar frequency meter, to give a 1000-cps beat-note output from the receiver (in A-1 reception condition) when the receiver is first turned on. Then, as the receiver drifts, the frequency meter is readjusted to produce this same beat note. When the output frequency becomes essentially stable, the total frequency drift is indicated by the difference between the final and the original frequency settings of the frequency meter dial. If the drift is small, the frequency meter dial setting may be left fixed, and the change in beat-note frequency observed instead. If the receiver is not designed for A-1 operation, a signal that will heterodyne with the intermediate frequency may be injected into the final detector through suitable loose coupling means, and the frequency meter used as directed above.

11-139. GAIN CONTROL EFFECTS. Frequency stability with gain change may be determined by feeding the maximum unmodulated carrier output of the signal generator to the receiver (as adjusted for cw (A-1) or fsk (agc off) reception), and then turning the rf gain control from maximum gain down until the output signal can just be heard. The change in output beat-note frequency should be less than 100 cps with a well-designed receiver for input signals below 30 mc, and less than 10 cps for input signals below 1 mc. A similar test for receivers not designed for the above modes of operation can be made by using a heterodyning voltage injected into the second detector as described above. For such receivers, greater frequency changes are usually tolerable (up to 10 percent of the 6-db down bandwidth of the over-all selectivity curve).

11-140. SIGNAL STRENGTH EFFECTS. Frequency stability with input-signal voltage change may be determined by noting the beat-note change during a cw (A-1) or fsk (agc off) resonant overload measurement, made as previously described. The same beat-note tolerances as above will apply. This test may be made on receivers without beat-frequency oscillators by using auxiliary beating means as described below.

11-141. NOISE (INTERFERENCE) MEASUREMENTS.

11-142. Because of the increased number of electronic and electromechanical equipments required to meet modern military needs, the elimination or suppression of noise and radio interference produced by these equipments has assumed greater importance. Interference, it should be remembered, not only may restrict or prevent vital communications, but also may divulge the position of a military unit to the enemy.

11-143. Various test equipments, called _radio test sets_ or _noise meters_, are available to aid you in measuring and locating interference. However, you must follow a systematic and logical procedure to locate the offending noise. If the receiver antenna is introducing the interference, disconnecting the antenna at the receiver input terminals will generally cause the noise to disappear or diminish considerably. Noise being conducted through the power line will, of course, not be affected by the above procedure, except for that type of interference that requires cross modulation with a carrier to make itself evident. Also, noise generated within the receiver itself will not be altered. The above test is important when the sound of the interference appears similar to tube noise or power-supply hum. Caution must be observed when using this method on receivers that operate on the higher frequencies, since a short length of antenna lead (or even an unshielded circuit) may be sufficient to pick up considerable interference when the source is close to the receiver.

11-144. When it is established that the antenna is picking up interference, it is necessary to determine the exact source. An effective method is to turn off all equipments operating in the vicinity; if the interference ceases, each individual equipment can then be restarted, one at a time, until the equipment which causes the interference to reappear is located. It is more advantageous to begin with all equipments shut down, rather than stopping individual equipments with others running, because of the possibility that a weak source may be masked by a stronger source.

11-145. A common method of locating a noise source, using a noise meter, is to move about the suspected area with the instrument and observe the intensity on the indicating meter, or listen to the audio level with a headset. Since noise sources usually have a large gradient, it is often possible to proceed to the source of interference by walking in the direction of increasing readings. After locating the offending equipment, it is necessary to determine the particular part of the equipment that is responsible for the disturbance. This is accomplished, if possible, by judicious use of individual switches (or other means of disconnection) on various units of the equipment and by the use of probe antennas. Two types of probe antennas are available: the magnetic type, consisting of a small loop for magnetic pickup, and the electrostatic type, consisting of a length of shielded cable with about 5 inches of the insulated (but shield-stripped) inner conductor extending for electrical pickup. The shield covering the leads to the probe should be connected to the case of the noise meter. A probe antenna is effective only when brought close to the source, and is a great aid in locating the actual source.

11-146. A test equipment of the type used to measure radio interference is essentially a sensitive portable receiver covering a specified range of frequencies and capable of measuring noise levels and field intensities. This type of equipment may be used as an rf voltmeter to measure a voltage between two points. It differs from a conventional receiver, however, in two respects: (1) a time delay is introduced into the avc circuit so that the output meter indicates the noise voltage in terms of the peak (or quasi-peak) value, which is more significant than the average, and (2) the gain of the receiver is adjustable to previously calibrated levels, to ensure uniformity of measurement on all frequencies. A calibrating signal source is included in the equipment for the latter requirement. For the former requirement, a standard noise source is provided; it is usually built into the test equipment. Briefly, it consists of a diode operating at saturation, the shot-effect noise of the tube providing a sufficiently constant source of noise for calibration purposes. To maintain space current in the tube at saturation, a filament control (rheostat) is provided. For calibration procedure the instructional literature that accompanies each test equipment should be followed in detail. In general, it should be noted that the above test equipments require extreme care and careful maintenance to provide reliable service. Realignment and calibration are necessary at frequent intervals, especially if the equipments are often transported from place to place.

11-147. Finally, it should be stressed that equipments which normally do not produce noise may do so when they are defective or in bad condition. This is particularly true of rotating or vibrating machinery. All commutators, slip rings, brushes, and brush holders must be in good condition. All normal ground connections to the frame or housing should be clean and tight. Movable contacts, such as switch points, relay contacts, etc, should be clean and properly adjusted for minimum arcing. All shielded connections and bonding must make clean and tight electrical contact. Therefore, when the source of interference is located, corrective maintenance should be considered as part of the job of eliminating noise and interference. Further information regarding noise interference is contained under paragraph heading 11-449.

11-148. AVC CHARACTERISTIC MEASUREMENTS.

11-149. GENERAL. An automatic-volume-control (avc) circuit reduces the effects of fading by maintaining a constant carrier level at the detector input of an a-m receiver, despite variations of the input signal carrier level. To determine the effectiveness of the avc circuit, its characteristic should be measured at the center frequency of each band covered by the receiver. A curve can then be plotted to compare the change of the receiver output to input signal levels.

11-150. OUTPUT VOLTAGE METHOD. Figure 11-5 shows the test equipment arrangement. The signal generator is connected through a dummy antenna, and modulated 30 percent at 400 cps. It is important to prevent overloading by setting the applied signal to 2 volts and adjusting the volume control for the desired maximum output level. In this test, the overload point is taken as one-half the maximum undistorted power output. Begin with an input signal level of 1 microvolt, and increase the signal in suitable steps to 2 volts. (This instruction should be followed, even in the case of low-sensitivity receivers.) Record each output voltage. Take numerous readings in the region of the bend of the curve, in order to show clearly the effect of the avc as it comes into operation.

11-151. Readings also should be taken at close intervals for inputs greater than 0.1 volt. Here the purpose is to detect "modulation rise," which is likely to occur when the input signal is strong. (Modulation rise is an increase in the modulation percentage caused by non-linearity of any tuned amplifier, primarily the last i-f stage. The rise is attributed to the generation of new frequency components, which appear in the output of the receiver as additional harmonic distortion, especially second harmonic. A 20-percent modulation rise, for example, corresponds to a second harmonic of 5 percent.) If the output voltage should increase unexpectedly as the input to the receiver becomes strong, the possibility of modulation rise can be verified by distortion readings of the receiver output, with particular stress on second harmonic. It is important not to confuse modulation rise with overloading of the audio power output stage. Note that the characteristic curve of a simple avc circuit should be curved throughout, without any predominant region of bend.

11-152. OUTPUT POWER METHOD. Another method of obtaining the avc characteristic, employing an rf signal modulated 30 percent at 400 cps, is described below. The

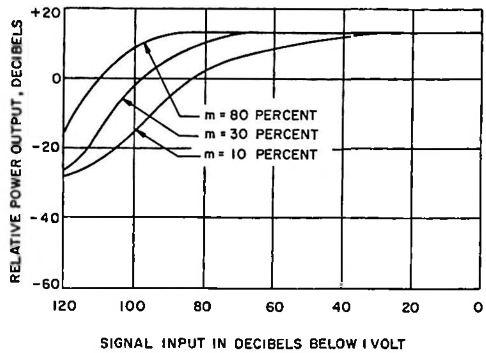


Figure 11-14. Automatic Volume Control, Characteristic Curve

receiver volume control is initially set to provide maximum output power, and the input signal is adjusted to the lowest amplitude of the test. For each succeeding step, the input signal is increased appropriately. Eventually, the output power will be increased to approximately one-quarter of the rated maximum. At this point, the volume control is set back until the power output is reduced by one-tenth. Then, the process of increasing the input signal is resumed. However, the power output to be recorded is 10 times the measured value. Once more, the output reading will eventually be increased to one-quarter of the rated maximum, and again the volume control should be set back until the reading is reduced by a tenth. Now the correct power is 100 times the measured value. The test is continued until a maximum input signal of 1 volt is reached. The curves obtained by this procedure represent the power-output-versus-input-voltage relationship that would exist at the maximum volume control setting if the audio amplifier would not overload.

11-153. Avc characteristics obtained by this method disclose the following information:

a. The sensitivity of the receiver, in microvolts, for any desired output level. To determine the sensitivity at an output of 0.5 watt, for example, refer to the avc curve of interest, and read the input corresponding to an output of 0.5 watt.

b. The residual noise level of the receiver (especially receivers of comparatively low sensitivity). Thus, a constant output power at the lowest input levels prior to the usual increase of output power can be considered as the residual noise level.

c. The power output resulting from any input voltage at any position of the volume control.

11-154. STANDARD METHOD. The standard method for measuring the avc circuit characteristics is described below. Set the signal generator for the standard 30-percent, 400-cps modulation and a carrier amplitude corresponding to a typical signal level, such as 5 millivolts, for an amplitude-modulation communication receiver, or 1 millivolt for a television receiver video signal. The receiver volume control is set for a standard test output. The carrier level is then varied over an extreme range, such as 1 μv to 1 volt, and the relative output plotted in db as a function of the carrier input level, also in db. The curve may be repeated, if desired, for several percentages of modulation. Typical results are shown in figure 11-14; the flatter the curve, the better the avc circuit.

11-155. FM RECEIVER METHOD. In conducting a test of the avc characteristic of an fm receiver, it is necessary to measure the avc voltage directly. One acceptable procedure is to connect a milliammeter in series with the grid leak, or to connect a vacuum-tube voltmeter across the grid leak of the limiter (or in series with the diode load resistor of the ratio detector). If a vacuum-tube voltmeter is to be connected across the grid leak, include an isolating resistor of a few thousand ohms resistance in series with the meter leads. Either a modulated or unmodulated signal can be applied to the receiver. The readings of the meter used are plotted against the corresponding magnitudes of the applied rf signal. The setting of the volume control for this test is, of course, immaterial.

11-156. DELAYED AVC CONSIDERATIONS. Delayed avc circuits are often incorporated in a receiver because even the weakest signal received in conventional avc circuits tends to reduce the gain of the receiver somewhat. The delayed avc adaptation incorporates a separate diode (avc diode), in addition to the detector diode. Part of the signal fed to the detector diode is coupled to the avc diode by a small capacitor. The avc diode is maintained at a suitable bias; this bias keeps the tube from conducting and thus producing the avc voltage until the peak voltage of the amplified signal voltage equals the bias introduced to the diode. For very weak signals, which do not produce enough voltage on the plate of the avc diode to overcome the existing negative potential, no avc voltage is developed. Thus, the sensitivity of the receiver remains constant, just as if the automatic volume control were not being used. When normal-strength signals are being received, which do not need the maximum sensitivity of the receiver, enough signal voltage will be coupled to the avc diode to overcome the bias applied. Avc voltage will be developed normally for these stronger signals. The functions of detection and avc voltage rectification are generally incorporated in a single tube. Measurements on delayed avc circuits are made in the same manner as described for conventional avc circuits; however, particular attention should be given at the low-input portions of the curve.

11-157. AFC CHARACTERISTIC MEASUREMENTS.

11-158. GENERAL. Automatic-frequency-control (afc) circuits are most often found in frequency-modulation receivers and in very-high-frequency and ultra-high-frequency receivers, because of the high-degree oscillator-frequency stability required. Fm receivers incorporate discriminator circuits whose output voltage and polarity are contingent upon the direction and deviation from a center, or mean, value. For purposes of oscillator frequency control, a sampling of this voltage is filtered to remove any ac component, and the resulting variation of dc voltage is applied to a reactance tube. This tube simulates a variable capacitor in the tuned plate circuit of the local oscillator. As the frequency deviation of the receiver oscillator increases, the dc voltage acting upon the reactance tube increases, with corresponding changes in the effective reactance presented at the output of the tube. If the introduction of such changes in the tank circuit of an oscillator tends to restore the desired frequency of oscillation, frequency control of the oscillator is effected. The use of this technique can decrease the amount of frequency deviation by a factor as high as 100 to 200, with respect to an uncontrolled receiver. Because the variation of the reactance is a direct function of oscillator frequency, a measurement of the afc characteristics is best made by performing an oscillator frequency-drift measurement.

11-159. FM RECEIVER CONSIDERATIONS. In measuring frequency drift in an fm receiver, you must remember that normally the detector stage cannot be used as a heterodyne detector. Unless such a detector is improvised, there is no way of setting the second generator for a zero-beat condition. Of the various solutions to this difficulty, only a few simple ones will be mentioned. If a limiter is present, it can be modified to serve as the heterodyne detector in this measurement. The tuned circuit in the output of the stage can be replaced with a resistor of approximately 50,000 ohms. By means of a large capacitor (0.05 or 0.1 μf), a set of headphones is then coupled to the junction of the resistor with the plate. Assuming that the grid circuit of the limiter is driven into saturation, it should be possible to conduct the frequency-drift measurement on the basis of the beat note audible in the headphones. The signals provided by the generators should be unmodulated. On the other hand, there may not be any limiting stage in the receiver. If a ratio detector is employed, short out or remove one of the diodes. This procedure should cause the remaining diode to act as a heterodyne detector, making it possible to conduct the test in the same way as for an amplitude-modulation receiver. If a locked-in detector is used, disable the oscillating section of the stage and proceed in the usual way. Finally, in the case of a gated-beam fm detector, short out the cathode biasing resistor. This should convert the stage into a heterodyne detector, and once again make it possible to proceed as described previously for amplitude-modulation receivers, in paragraph 11-129. More elaborate procedures make use of circuits that indicate deviation from a center frequency.

11-160. ALIGNMENT OF CRYSTAL FILTER CIRCUITS.

11-161. Crystal filters are incorporated in communications receivers which require an extremely high order of selectivity. These filters are usually located between the receiver-mixer and intermediate-amplifier stages. The crystal is the major component of the filter, and is used because of the extremely high Q that can be obtained from its use. Ordinarily, the filter is externally adjustable so that a variation in bandpass can be obtained.

11-162. Crystal filters are often used as wave traps, and are extensively used in single-sideband equipment because of the sharp-frequency-cutoff properties required for this type of equipment. However, these are generally of the crystal-lattice type; they are usually hermetically sealed or potted, and should not be tampered with. Some lattice or half-lattice-type crystal filters are provided with several adjustable trimmers, accessible as screwdriver adjustments. Some of these are labeled as factory adjustments, and should never be disturbed. In any case, the manufacturer's data should always be consulted before any adjustments are attempted. Plug-in filters are readily replaced. Mechanical filters should never be tampered with or repairs attempted under any condition.

11-163. Schematic diagrams of crystal filters usually indicate variable capacitors and often variable inductances. Such diagrams may be misleading to those unfamiliar with filter circuits. Capacitors in parallel with filter crystals are usually of very small value, on the order of one to several picofarads. These often consist of a pair of leads given a slight wrap or twist, or they may be a piece of wire bent near the crystal holder or electrode. Such capacitors are factory-adjusted, and are usually not accessible without dismantling the filter.

11-164. The schematic symbol for a crystal and its equivalent electrical circuit are illustrated in figure 11-15A. A crystal in its holder is actually a combination of both series- and parallel-resonant circuits, and as such it has two resonant frequencies, as indicated in figure 11-15B. The series-resonant frequency occurs at the point where the reactance curve crosses the zero-reactance line, and the parallel-resonant (anti-resonant) frequency occurs at the point where the reactance curve rises to a high inductive reactance and then falls sharply through the zero-reference line to a high capacitive reactance. In most crystals, the two resonant frequency points will occur within a few hundred cycles of each other, but the points can be spread (or narrowed) by shunting them with a lump constant so

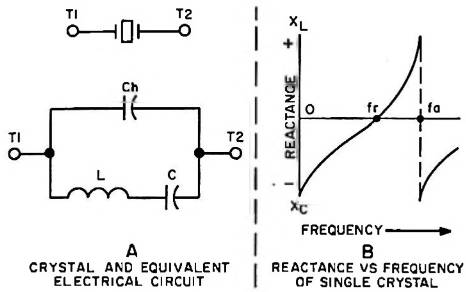


Figure 11-15. Equivalent Electrical Circuit and Reactance Curve of a Quartz Crystal

that a suitable filter network can be designed. Phasing controls on interference filters are examples of capacitance introduced into the filter circuit to shift the crystal rejection slot (parallel-resonant frequency) so that particular unwanted signals can be rejected.

11-165. When aligning filter circuits, you must consider the circuit in which it is integrated. When connected in the intermediate-frequency amplifiers of a communications receiver, either conventional a-m or single-sideband, the alignment will consist principally of properly tuning the resonant input circuit to the filter (usually a mixer plate circuit) and the resonant circuit at the output circuit (usually a grid circuit), as shown in figure 11-16, for maximum output. The points of parallel resonance must be aligned for sharp cutoff (maximum attenuation) at the design frequency. This is an especially important consideration for equipments containing automatic-frequency-control (afc) circuits, because if sufficient response is not allowed, the carrier may be severely attenuated at a slightly too low (or too high) frequency, and thereby cause the afc circuit to drop control. Thus the desired limits of afc operation are also considered in the bandpass of the filter, and vice versa.

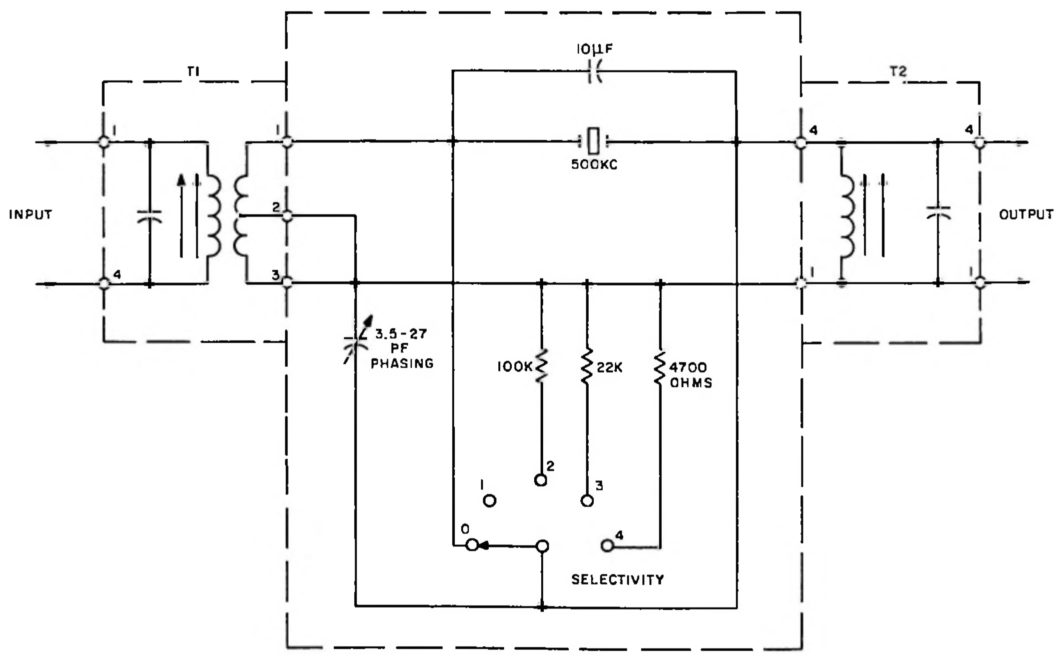


Figure 11-16. Crystal Filter Circuit

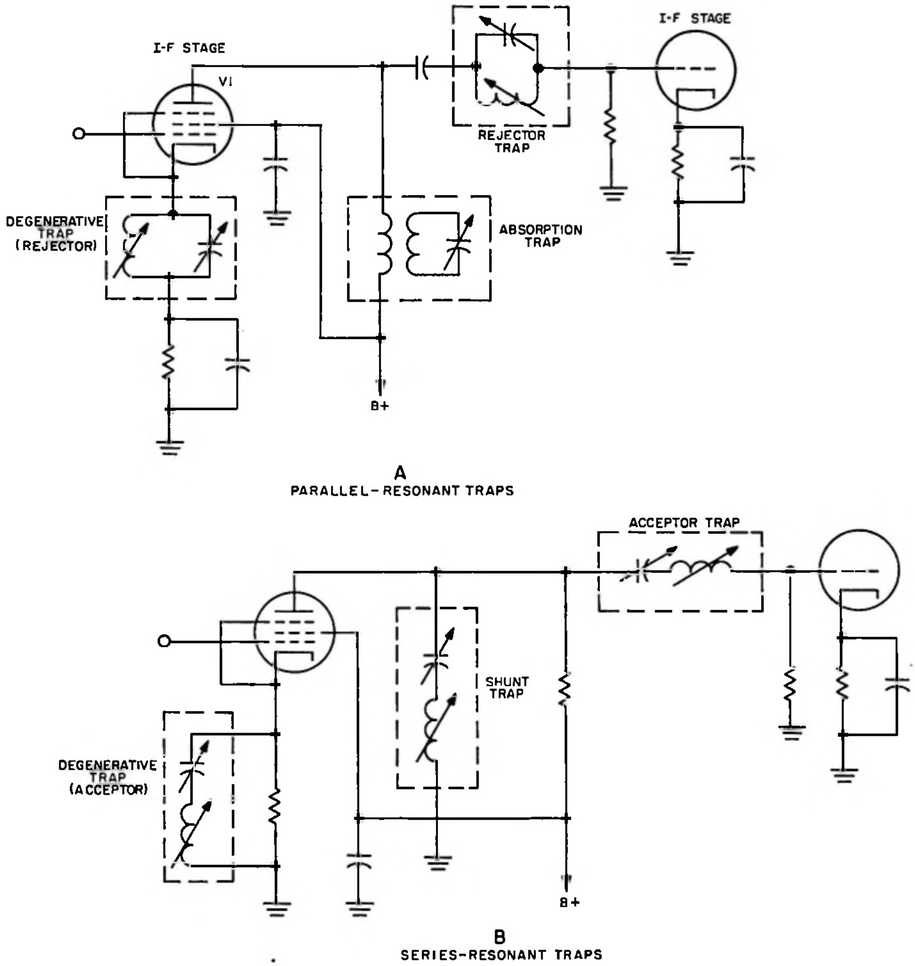


Figure 11-17. Wave Trap Circuits

11-166. ALIGNMENT OF WAVE TRAPS.

11-167. The term wave trap usually refers to a resonant element used as an auxiliary device to provide additional frequency selectivity in a radio circuit at a particular frequency. It may take a distributed form (resonant stub or cavity), or it may consist of a lumped reactor combination (inductor and capacitor), as in figure 11-17. A trap normally provides a means of rejecting (or accepting) signals over only a relatively narrow band of signal frequencies, of a width which depends in part on the effective Q of the trap circuits.

11-168. The trapping desired may result from the "shorting" effect of a series-resonant circuit shunted across the signal path, from the selective opposition to the flow of current afforded by a high value of resonant impedance in series with the path; from selective degeneration in an amplifier, produced by using resonant circuits to provide frequency-dependent feedback, etc.

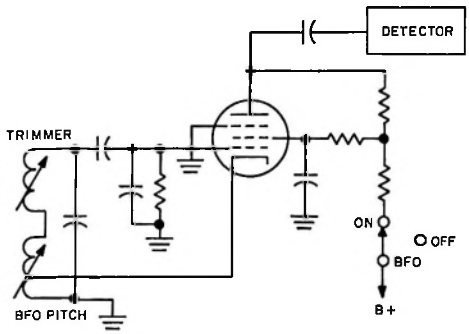


Figure 11-18. Beat Frequency Oscillator Circuit

11-169. Some of the more common lumped-reactor wave-trap applications are shown in figure 11-17. In addition to the wave traps illustrated, many other forms may be found in radio equipment. In some applications, a wave trap is used to suppress response at a frequency not desired in one channel, and the resulting trap resonance at that frequency is used as a means of supplying signal to a second channel. Resistance-capacitance (RC), resistance-inductance (RL), and inductance-capacitance (LC) networks, affording high-pass and low-pass characteristics, are also employed to provide band elimination or bandpass effects for wave-trapping purposes.

11-170. The operating frequencies and apparent effects of wave traps differ from one type of equipment to another. In general, it can be expected that the traps will be left until the last steps in a prescribed alignment procedure, because of their auxiliary or corrective nature. Adjustment of wave-trap trimmers must usually be accomplished at very specific frequencies and under particular conditions, which should be rigidly observed. If adequate instructions for wave-trap alignment are lacking in an equipment technical manual, immediate steps should be taken to obtain further instructions. An incorrectly adjusted trap circuit may produce serious shortcomings in equipment operation, which are not apparent to the operator under ordinary conditions.

11-171. The signal generator and output indicator commonly employed in the alignment of receiver tuned circuits will usually serve for wave-trap alignment in receiving equipment. Other forms of radio equipment employing traps may require special instrumentation, such as field strength meters and oscilloscopes.

11-172. ALIGNMENT OF BEAT-FREQUENCY OSCILLATORS.

11-173. Beat-frequency oscillators (bfo's) are incorporated in communications receivers to provide an audible indication of received continuous-wave transmissions, to receive teletypewriter transmissions by providing the "mark" and "space" audio-frequency signals, and to calibrate dials in receivers which contain internal crystal calibration oscillators.

11-174. Figure 11-18 shows a schematic diagram of a typical manually operated bfo which is heterodyned against the intermediate frequency. The procedure for aligning this type of oscillator is to feed an unmodulated signal into the input of the receiver, or feed the i-f frequency directly to the mixer, with the receiver bfo switch ON. The bfo pitch control should then be adjusted for zero beat. If the bfo control index (or 0) does not match the index on the chassis, the setscrew on the dial should be loosened and the dial adjusted. When these conditions prevail, the index position of the dial represents a bfo frequency which is exactly that of the intermediate frequency. If the bfo dial is calibrated for the frequency of pitch, you should use an accurately calibrated audio generator, a frequency counter, or oscilloscope to measure the actual pitch frequency. The bfo dial is then set at maximum calibrated pitch while you adjust the pitch trimmer for proper calibration. Because the bfo pitch control and the pitch trimmer interact electrically, you may have to repeat the procedures several times until the bfo is properly calibrated. Specific information regarding the alignment of beat-frequency oscillators, which are part of critical circuits, are included in the technical manuals of the receivers involved.

11-175. SQUELCH (SILENCER) CIRCUIT MEASUREMENTS.

11-176. Fm and high-frequency receiver circuits inherently have a high noise level when no signal is being received. During communications, where a receiver is tuned to a specific frequency for long stand-by periods in anticipation of signals that may appear at any time, the continuous roar of noise is highly objectionable to anyone in the vicinity of the receiver. In a communications receiver, therefore, it is desirable that a squelch (or silencer) circuit be incorporated to silence the audio output during these periods when no signal is being received. Similarly, in broadcast receivers, it is undesirable to have a large noise output when tuning from one station to another. A squelch circuit eliminates this undesirable interstation noise. When very low-level signals are being received (nearly always from interfering stations), the receiver output is also shut off to eliminate unwanted signal noise and other disturbances that are annoying and fatiguing to the operator. Most squelch circuits operate on the principle of applying a large negative bias to the grid of the first audio amplifier tube whenever the signal voltage is very low or entirely absent at the detector (in the case of a-m receivers), or at the limiter (in the case of fm receivers). The squelch bias must be sufficiently in excess of cutoff to prevent the noise output from the intermediate amplifiers from causing a plate current to flow in the first audio amplifier tube, even momentarily on the noise peaks.

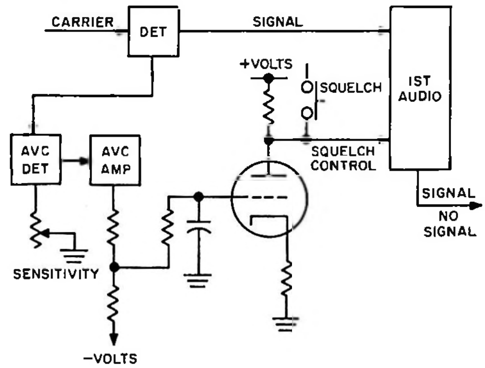


Figure 11-19. Squelch (Silencer) Circuit

11-177. Figure 11-19 shows a simplified schematic diagram of an electronic silencer circuit. In the absence of an incoming signal, the potential of the squelch control-tube grid is positive, as a result of the avc action, with a subsequent flow of heavy plate current. The resulting plate voltage drop cuts off the first audio-frequency amplifier, thereby blocking all noise and weak signals by preventing transmission through the audio-frequency amplifier. When a signal is received, the avc detector output reduces the current through the avc amplifier, the action of which cuts off the squelch tube and removes the squelch bias to the audio amplifier. The point at which the squelch control circuit allows signals to be received is determined by the setting of the avc sensitivity control.

11-178. For the determination of the squelch characteristic of an a-m receiver, the test equipment should be connected to the receiver as shown in figure 11-5. The signal generator should be set to frequency with 400-cps, 30-percent modulation. With the signal generator rf output control set for zero output, the receiver output should be noted;

it should be essentially zero. Gradually, the signal generator rf output should be increased until the squelch circuit operates. Operation of the squelch circuit is indicated by a sudden increase in the radio receiver output. The signal generator rf output required for the operation of the squelch circuit may be recorded as the squelch characteristic.

11-179. RECEIVER LIMITING-LEVEL MEASUREMENTS.

11-180. Limiting-level measurements determine the minimum signal input level (threshold) required to produce a constant output. This "threshold of limiting" can be obtained by a plot of signal input (horizontal) versus power output (vertical). To perform the measurement, connect a signal generator to the receiver input and measure the output power or voltage across the output terminals, as shown in figure 11-5. The generator is first set to frequency, with a modulation of 400 cps and a 30-percent modulation deviation. The generator output attenuator is set for a high level of output, on the order of 10 millivolts, and the receiver volume control is set for maximum undistorted power output. Following these initial adjustments, the generator output is lowered to 1 microvolt, and the output power noted. The generator output is then increased, in convenient steps, until no further change in output power is noted. The knee of the curve on the resulting graph will show the input level at which the limiting action is initiated.

11-181. A-M RECEIVER ALIGNMENT.

11-182. GENERAL. Prior to the alignment of an a-m receiver, the automatic gain control or automatic volume control should be turned off, and the gain adjusted by means of the manual radio-frequency gain control. For communications receivers, the gain level should be set to give the standard 6 milliwatts of audio output with about 100 to 1000 microvolts of signal input at the receiver antenna terminals. This alignment condition is desirable to reduce the detuning effect of receiver gain variations as reflected in changes of over-all selectivity. It insures that the circuits will be resonated under average load conditions, at approximately the middle working value of tube input reactance, and with freedom from serious regeneration.

11-183. With most communications receivers, this condition also reduces receiver noise to a degree which makes it unnecessary to quiet the receiver by removing a tube in the amplifier stage preceding the point of alignment-signal injection. The preselector main tuning control should be adjusted to a low frequency, as already mentioned, retuning slightly, if necessary, to insure that no unwanted signal is present. For preselector alignment, the antenna should be disconnected from the receiver.

11-184. DISABLING AUTOMATIC GAIN CONTROLS. Amplitude-modulation receiving equipment which operates with automatic gain control as a permanent condition, with no built-in provision for the alternative manual control of rf and i-f gain, may present a problem, especially if considerable regeneration is normally present at full gain. It may not be feasible to disable the automatic gain control in order to add a temporary battery-biased manual gain control potentiometer in its place. In such cases, it will be necessary to align each section of the receiver, with suitable signal-input levels at the various points of signal injection, to produce final detector operation below the threshold of agc action. A tube should be removed in a stage preceding the point of alignment-signal injection, to preclude the presence of unwanted signals and noise.

11-185. DISABLING LOCAL OSCILLATORS. It is preferable not to disable the heterodyne oscillator or oscillators when aligning a receiver, except for the beat frequency oscillator which is used to provide tone output from the final detector in code (cw) reception. The heterodyne oscillator injection voltage is ordinarily a major factor controlling the mixer tube operating bias and impedance, with consequent influence on gain and both mixer input and output circuit resonance.

11-186. In some cases, suitable adjustment of heterodyne oscillator tuning may not be feasible as a means of preventing undesired beats or spurious signals that may result from the interaction of the alignment signal and the heterodyne injection voltage. The oscillator must then be disabled, preferably by removing either the oscillator tube or the final multiplier tube (if the heterodyne oscillator employs frequency-multiplier stages). Stopping an oscillator by short-circuiting its grid to ground or by shorting its tank circuit may cause serious damage to the oscillator tube and other electronic parts.

11-187. BFO CONSIDERATIONS. The beat-frequency oscillator injection voltage in a properly designed cw receiver usually produces a large fixed bias at the final detector, which will mask its rectified voltage changes. This masking is very objectionable when the rectifier signal voltage is employed as an output indication for alignment.

11-188. Before starting the actual alignment, you should disable all those auxiliary functions provided in the receiver which may interfere with proper output indication or circuit resonance. This includes silencer (squelch) circuits and noise limiters.

11-189. I-F AMPLIFIER ALIGNMENT. With a few exceptions, such as some trap circuits, i-f resonant circuits are aligned by adjusting their trimmers to produce maximum signal voltage. The i-f trimmers of the typical a-m receiver are thus adjusted to produce maximum final-detector signal input voltage, using the input-signal frequency or frequencies prescribed in the technical manual for the equipment. In many cases, this will be the nominal band-center frequency of the particular i-f amplifier. In other instances, usually involving relatively wide i-f pass bands, "peaking" of some or all trimmers for maximum response at one or more frequencies off the band center will be specified.

11-190. In general, the last i-f transformer preceding the detector should be aligned first, unless some other order is required in the equipment technical manual. The input from the signal generator should be adjusted to produce a signal output level which is well above the noise level at the output indicator, but which is also well below saturation or the overdrive level for the vacuum tubes of the system. The signal input should be progressively reduced as needed, as more circuits are brought into proper alignment, with the progression of circuit adjustment moving toward the mixer stage. After the first round of alignment adjustments of the i-f amplifier stages is completed, an over-all check of the i-f alignment should be made.

11-191. A similar procedure should be used for the alignment of the preceding i-f amplifier(s) in receivers employing more than one frequency conversion. The i-f signal input should, in each case, be injected at the input electrode of the mixer preceding that particular i-f amplifier. This insures inclusion of the i-f transformer located in the output circuit of the mixer. The associated conversion oscillator should be disabled if necessary, as previously discussed.

11-192. RF STAGE ALIGNMENT. In addition to a suitable signal generator, the dummy antenna specified by the receiver instruction book should be used to simulate an ideal antenna for the receiver. The signal generator (modulated or unmodulated, as required) should first be accurately adjusted to the upper alignment frequency specified for the particular receiver tuning band, using an external frequency standard if necessary. If an antenna trimmer control is provided on the front panel of the receiver, it should be set to the middle of its range. The receiver is then carefully tuned to that signal frequency, and the generator output adjusted to produce the desired maximum (or other specified optimum) deflection on the receiver output indicator. The tuning-dial frequency indication should coincide closely with the signal frequency being supplied. If it does not, the tuning dial should be reset to indicate the proper frequency. The high-frequency (shunt capacitance) trimmer of the oscillator tank circuit should then be adjusted to produce optimum output from the test signal. Following these adjustments, the interstage and antenna circuit shunt-capacitance trimmers should be adjusted for optimum output, with the test signal input level reset, as needed, to avoid receiver saturation effects.

11-193. Oscillator shunt trimmers occasionally have an unusually wide range of adjustment. For this and other reasons, it is possible to misalign the circuits so as to place the heterodyne oscillator on the wrong side of the signal frequency. In many instances, this mistake will be revealed in inability to obtain anything resembling good circuit tracking over the tuning band. Sometimes, however, the mistake will not be so clearly apparent. Therefore, it is always wise to ascertain that the oscillator is being trimmed on the proper side of the desired signal frequency. Determination of the proper relationship from the equipment instruction book, together with careful observation of shunt trimmer positioning (whether its capacitance is increasing or decreasing relative to the two positions of heterodyne response which it produces), will help to prevent error.

11-194. Next, it is necessary to check oscillator alignment at some specified frequency near the low-frequency end of the tuning band. In many military receivers, iron-core or eddy-current trimmers are provided in the rf coils to permit tank inductance adjustments for optimum low-frequency tracking of all the rf circuits. The inductance adjustments should be made on all coils except the oscillator coil before the oscillator series padder is checked. Then the series padder should be trimmed to produce optimum output while the receiver tuning control is "rocked" back and forth through the region of best signal response. When this process is completed, the shunt trimmer adjustments should be touched up for optimum response and correct tuning-dial reading (calibration) at the high-frequency alignment point in the band. The low-end padder adjustments should then be touched up for optimum response, and the tuning-dial reading checked against the test-signal frequency.

11-195. If oscillator tracking relative to the other rf circuits is poor over the band, as indicated by abnormal variations of gain (and/or output noise) as the tuning control is operated throughout its range, it may be necessary to adjust the oscillator tank inductance trimmer. The correction needed to produce better tracking may be determined by trial readjustment of the oscillator shunt-capacitance trimmer. If the tracking (as checked by tuning from the high-frequency to the low-frequency alignment points) can be improved by increasing the shunt trimmer capacitance, the oscillator tank inductance is low. If the shunt trimmer capacitance must be decreased to obtain improvement, the tank inductance is high. Correction adjustment of oscillator tank inductance will necessitate some changes in oscillator series-padder and shunt-trimmer adjustments, and the entire preselector alignment procedure must be repeated.

11-196. If the tuning dial is then still in error over part of the band, it may be possible to correct the calibration to some degree by further slight readjustments to the oscillator and other trimmers. This realignment should be undertaken only after careful study of the tracking discrepancies and calibration errors over the entire band, and with a full understanding of the superheterodyne tracking problems if adequate directions are not available. In general, it is inadvisable to sacrifice receiver gain and selectivity for the minor convenience of accurate tuning-dial calibration.

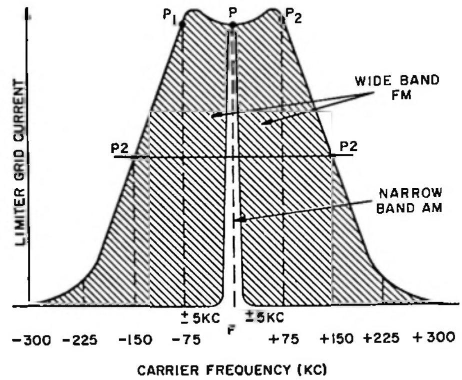


Figure 11-20. FM Versus A-M Resonance Curves

11-197. Receivers that incorporate i-f traps in their rf circuits should be checked by applying a signal, at the intermediate frequency, to the receiver input. The trimmers for such traps are usually adjusted for minimum output at the center frequency of the first i-f amplifier, and may require large input signal amplitude at that frequency.

11-198. FM RECEIVER ALIGNMENT.

11-199. GENERAL. The basic difference between receivers that are used for the reception of frequency- or phase-modulated signals and those that are used for the reception of amplitude-modulated signals are in the types of demodulator and i-f amplifier circuits employed. In an fm receiver a frequency-sensitive demodulator is used, and the i-f amplifiers are designed to cause, rather than avoid, amplitude limiting.

11-200. When testing several amplifier stages that have similar operating functions, such as successive i-f stages, you may be tempted to test immediately for an "over-all" response curve like that shown in figure 11-20. Such a curve may be obtained at the output of the last i-f stage or at the grid of the limiter. When using an fm signal generator for testing wide-band equipment, such as an fm receiver, the response curve can be observed directly on the screen of an oscilloscope. This procedure, used improperly, could consist of varying "at random" the different adjustments in all the stages until the over-all response curve appears to be satisfactory. The apparently good-looking curve results only too often from a compromise. This generally means that a poor alignment in one stage is compensated by overemphasized and shifted alignment in other stages. The reason for this is that one stage may be peaked unsymmetrically, another stage may have a center peak, and the other stages may have two response peaks. This has reference to all i-f stages. Hence, no stage by itself satisfies the condition required for linear networks with respect to amplitude and phase, and distortion is bound to result.

11-201. Regardless of the method of aligning the receiver, it is good practice to first align the discriminator (provided that a sufficient signal source is available, the sensitivity of the indicator is high, and the i-f transformers are not excessively detuned), and then to align the remainder of the set up to the discriminator. The correct i-f alignment should always be made by aligning first the i-f stage ahead of the limiter, then the preceding i-f stage, etc. As far as the rf

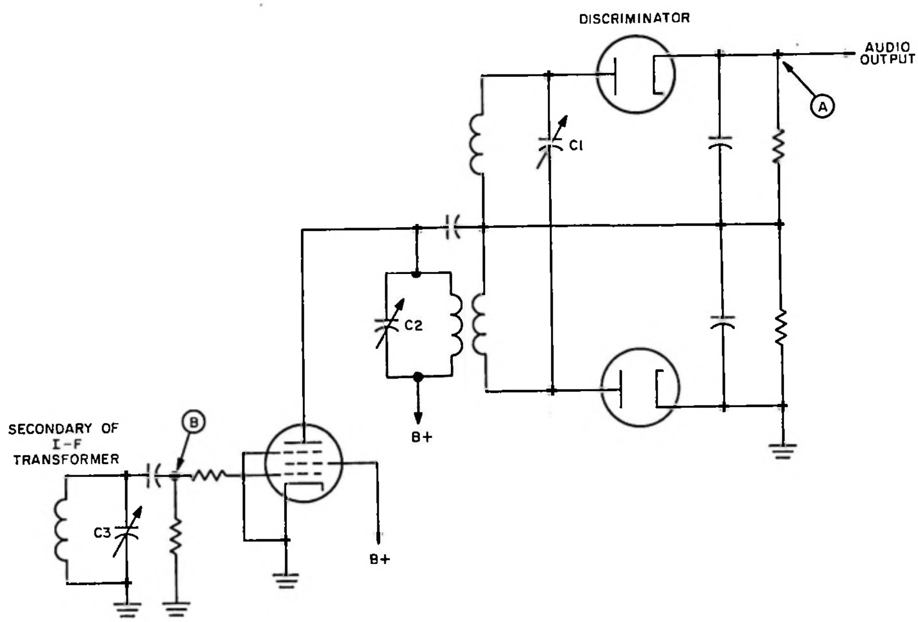


Figure 11-21. Limiter-Type Discriminator Circuit

stage or stages before the mixer stage are concerned, ordinary single peaking is generally practiced, since coil and tube damping nearly always provides the required broadness of the response curve. The receiver can be aligned most conveniently by use of an fm signal generator. However, if an fm generator is not available, an a-m generator can be used. A meter or an oscilloscope may be used as an output indicator. Normal receiver alignment consists of the following procedures:

a. Alignment of the demodulator (discriminator) stage.

b. Alignment of the limiter stage.

c. Alignment of the i-f amplifier stages.

d. Alignment of the rf stages.

11-202. LIMITER-TYPE DISCRIMINATOR ALIGNMENT. Figure 11-21 shows a schematic diagram of a limiter-type detector. In the double-tuned circuit shown, the primary and secondary are tuned to the carrier frequency. At the carrier frequency, the voltages developed across the diodes are equal to each other, and the diode currents are equal; thus the opposing voltages developed across the output diode resistors are equal, and therefore cancel, so that no voltage is developed at point A. At frequencies lower than that of the carrier frequency there is more current flow through the lower diode than through the upper diode, causing a negative voltage to be developed at point A. Conversely, at frequencies higher than that of the carrier there is more current flow through the upper diode, causing a positive voltage to be developed at point A. Within the range of carrier-frequency swing, the dc output changes in proportion to the frequency change.

11-203. There are several ways in which you can measure the linearity of a discriminator, but the most straightforward method seems to be the use of a high-resistance voltmeter to measure the output voltage between point A and ground (figure 11-21), while varying the applied intermediate frequency in known steps. A voltage of the center intermediate frequency is applied to the grid of the limiter (or to the grid of a mixer if the i-f amplifiers are properly aligned and the limiters are properly set), and the setting of the secondary capacitor C1 is varied until zero output is noted. (Note that both terminals of this capacitor are above ground potential; therefore, an insulated screwdriver should be used for this adjustment). The signal generator is then set above and below the intermediate frequency. The voltmeter should indicate equal but opposite direct voltages for equal but opposite frequency deviations. If unequal voltages are obtained, the setting of primary trimmer capacitor C2 is incorrect and must be adjusted until equality is secured. You may have to repeat these operations until the proper indications are obtained. The linearity over the entire range can be determined by plotting the values obtained for steps of frequency deviation, as shown in figure 11-22A. The output of the generator should be constant, or the limiter should be in full operation.

11-204. A visual method of aligning the discriminator, using an fm signal generator and an oscilloscope, has the advantage over the meter method in that the discriminator curve may be observed. Since the effects of the adjustments are visible, no guesswork is involved. There are two methods of setting the discriminator to its proper center frequency. The first method to be described is recommended for its more accurate and more easily observed results. It consists of applying an amplitude-modulated rf signal of the correct center frequency to the discriminator, and adjusting the discriminator secondary for minimum signal output. The output signal will disappear if the discriminator characteristic is symmetrical about the center frequency, since the output will be zero at the center frequency. In the second method, a marker pip is made to appear on the discriminator response curve, and is set for the cross-over point at the center frequency, so that the center frequency is actually determined by noting where the marker disappears and reappears. Because

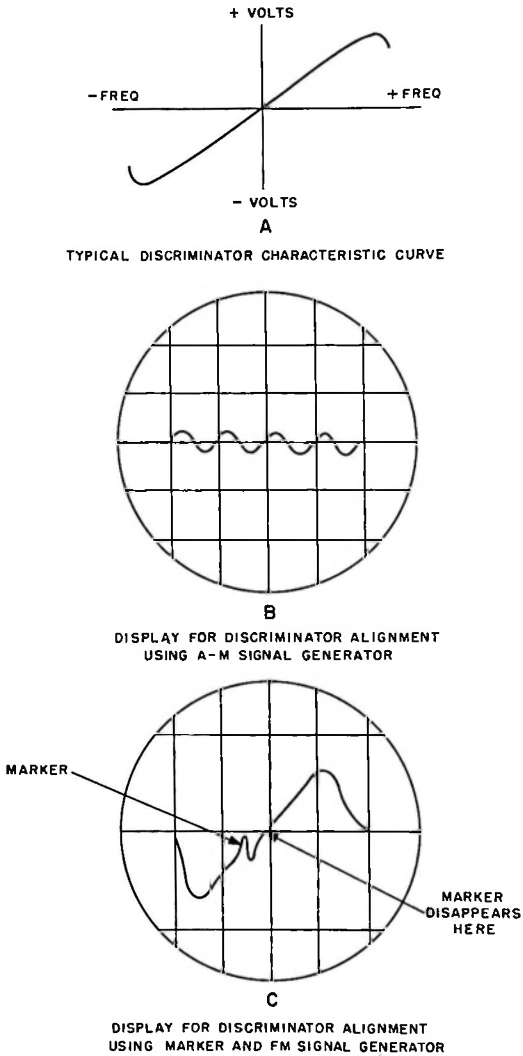


Figure 11-22. Discriminator Characteristic Measurements

of the difficulty in observing the exact points of appearance of this pip, this method sometimes leads to inaccurate results.

11-205. To use the a-m signal method for aligning the discriminator to its center frequency, connect an a-m signal generator to point B (figure 11-21) and the oscilloscope vertical input to point A. Set the signal generator for 400-cps amplitude modulation, and adjust the oscilloscope controls for a convenient pattern size. When the discriminator secondary trimmer capacitor, C1, is not adjusted to the current frequency, but is close to it, a pattern similar to that of figure 11-22B will appear. To align the discriminator to the correct center frequency, adjust the secondary trimmer capacitor slowly in one direction, and then in the other direction, until the 400-cps signal disappears and then reappears with a further movement of the trimmer. Set the capacitor midway between these points. Then connect an fm signal generator to point B (figure 11-21). Leave the oscilloscope vertical input at point A, and connect the horizontal input to the modulation circuit of the signal generator. Adjust the signal generator for full frequency deviation, and set the oscilloscope controls for a convenient pattern size. Then adjust the primary trimmer capacitor, C2, for a symmetrical curve similar to the one shown in figure 11-22A.

11-206. When you use the marker method, connect an fm signal generator and oscilloscope as described above. Then couple a marker generator or wavemeter with the signal generator output to point B (figure 11-21), so as to produce a pip on the discriminator response curve. With the marker signal generator or wavemeter set at the discriminator center frequency, adjust the secondary trimmer capacitor until the marker disappears at the cross-over point in the center of the response curve. Then adjust the primary trimmer capacitor for a symmetrical curve.

11-207. RATIO DETECTOR ALIGNMENT. Another type of fm detector, shown in figure 11-23, is called a ratio detector. This circuit is based on changes in the ratio of the voltages across the two diodes, rather than to differences in voltage. A ratio detector is virtually insensitive to amplitude variations. The tuning and coupling provisions are practically the same as in a limiter-type discriminator. As a result, the rf voltage developed across the diodes at any instant depends upon the amount of frequency deviation from the carrier center frequency. Unlike the arrangement in the limiter-type discriminator, however, the diodes are connected to conduct simultaneously, so that a negative voltage is developed across the load resistor. A filter capacitor connected across the load resistor has a value which is sufficient to hold the voltage constant, even at the lowest audio frequencies to be reproduced. The voltages across the diodes differ according to the instantaneous frequency of the carrier, and the rectified voltages across capacitors C3 and C4 are proportional to the corresponding diode voltages. Although each of these voltages is varying, their sum is held constant by the filter. As a result of this restriction, an audio output is developed across C4. Many modified versions of the ratio detector are in common use. However, the operation and alignment procedures are similar to those described for the ratio detector.

11-208. To align a ratio detector, connect an fm signal generator between point B and ground, and connect the input to the oscilloscope between point A and ground (figure 11-23). Set the generator to the center frequency, with maximum frequency deviation. Adjust discriminator the primary trimmer capacitor, C2, for a curve of maximum amplitude, which will appear somewhat S-shaped if the secondary trimmer capacitor C1 is not excessively detuned. It will be necessary to keep the generator attenuator control set for an output below that to which

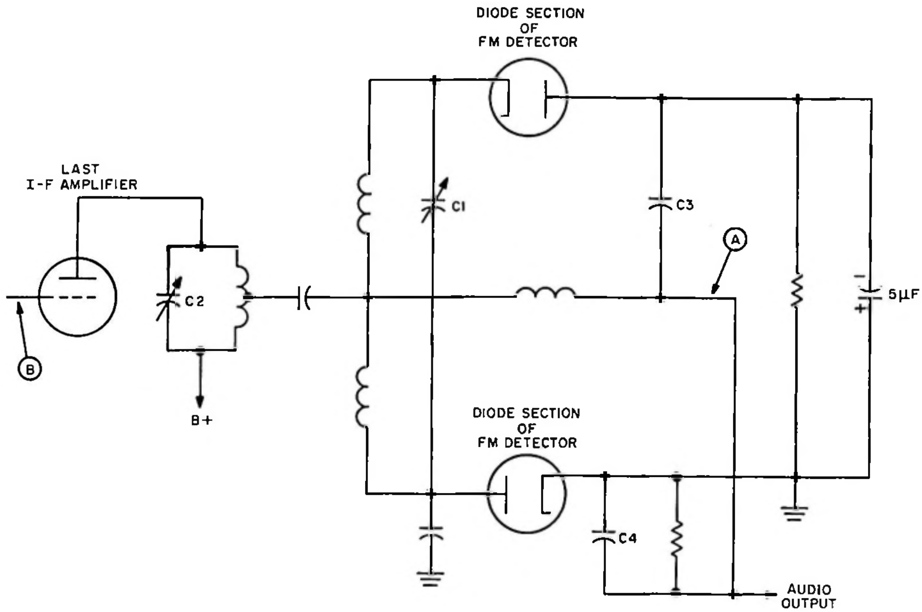


Figure 11-23. Ratio Detector Circuit

limiting occurs. Next adjust the secondary trimmer capacitor until the S-shaped curve is symmetrical (figure 11-22A), and set to exact center frequency as described previously for discriminators. Retune the detector primary trimmer for a symmetrical response of maximum amplitude, if necessary.

11-209. I-F AMPLIFIER ALIGNMENT. Specific alignment procedures are generally included in technical manuals for particular receivers. In these cases, specific response curves for each transformer may be given so that a particular over-all response curve may be obtained. However, in general, the intermediate-frequency amplifiers may be aligned by feeding an fm signal from the generator to the grid of the i-f amplifier just preceding the limiter, while observing the discriminator output. The secondary of the i-f amplifier is then adjusted for a symmetrical S-shaped curve, having a proper frequency response of maximum amplitude. The procedure is repeated to tune the primary. If the output can be kept below the threshold of limiting by reducing the generator output, adjust each i-f secondary and primary in order, proceeding from the last i-f stage back to the first i-f stage, for a symmetrical response curve of maximum amplitude. Should limiting occur, the response curve of the amplifier can be observed at the grid of the limiter, and the i-f amplifiers tuned for proper bandpass (figure 11-20).

11-210. RF AND OSCILLATOR STAGES ALIGNMENT. The rf and oscillator stages may also be aligned by means of an fm signal generator and an oscilloscope. To align these stages, connect the output from the generator to the antenna terminals of the receiver through a suitable matching network and connect the oscilloscope input to the discriminator output. Set the generator to a frequency in the approximate center of the band being tested, and set the frequency deviation greater than the receiver band-pass. Observing the receiver output response, tune the shunt (high-frequency trimmer capacitor) for maximum output. Set the signal generator and receiver to the low end of the band. Use a tuning wand and observe the oscilloscope pattern. If the signal amplitude decreases when either end of the wand is inserted into the oscillator coil, the tracking is satisfactory. If the output increases with the brass end of the wand inserted, spread the turns of the oscillator coil; if the output increases with the iron end of the wand inserted, compress the turns of the coil. Do not bend the coil excessively, as only a slight physical change is necessary at the high frequencies at which this type of equipment generally operates. Repeat these steps until no further change is noted. The last adjustment should be that of the shunt (high-frequency) trimmer capacitor. Return the signal generator and receiver to the center frequency, and adjust the shunt trimmer capacitor of the mixer-grid circuit for maximum output. If an rf stage is employed, also adjust the shunt trimmer capacitor of the rf stage for maximum output. Lower the frequency setting of the receiver and generator, and check the tracking of the mixer and rf grid circuits with the tuning wand. If the output increases with the brass end of the wand inserted in the coil, spread the coil turns; if the output increases with the iron end inserted, compress the coil turns. If the output decreases when either end is inserted, the tracking is correct. Do this for both the mixer and rf coils. Repeat these adjustments until no further improvement is noted.

11-211. You may also use an oscilloscope to check the response of the rf and mixer sections of a receiver. Connect the fm signal generator to the receiver input through a suitable matching network. Then connect the oscilloscope vertical input to the mixer plate decoupling network, or, by means of a high-frequency detector probe, to the mixer plate. You should remove the first i-f amplifier tube during this test to reduce the loading effect of the oscilloscope input. Couple a marker signal generator or wavemeter with the fm signal generator to the receiver input in order to determine the frequency points of the response curve. Set the fm signal generator for the desired frequency deviation; in many communications receivers the front-end response curve is from 150 to 200 kc wide. The bandwidth of the front end is largely fixed by the number and Q of the rf circuits, since all of the circuits are usually tuned to identical frequencies. Usually, the bandwidth of the rf circuits is considerably greater than the i-f bandwidth, so that the latter mainly determines the bandwidth of the entire receiver. When you make any adjustment of the front end, both the rf and over-all response must be considered, since it is important that the rf response be wide enough to pass all of the important frequency components of the signal. This check may be made by the previous procedure for measuring the rf response.

11-212. COMMUNICATIONS TRANSMITTER TESTING.

11-213. Because the operating procedure for most radio transmitters calls for continuous tuning and retuning, tube currents are automatically measured by the indicators provided. If these readings are periodically recorded, the cumulative record bears the same relationship to the transmitter that a sensitivity check does to the receiver. When the emission of a tube falls to 80 percent of normal rating, it should be replaced. Because it is possible for large tubes to become soft (gassy) when stored for a considerable time, they should be rotated with tubes of the same type in spare storage so as to maintain three sets of large tubes in good condition at all times. In some cases the residual gas can be removed by absorption into the filament if the filament voltage is operated at a slightly higher value than normal for a period of several hours, with no voltage applied to the plate and screen.

11-214. Ventilation and dust conditions present a greater problem to transmitters than they do to receivers, principally because there is more heat to be dissipated. Since dust forms a film which absorbs moisture, the insulation resistance is lowered to a point at which flash-over can occur. Therefore, cleaning periods must be initiated and regularly followed. These periods will vary for different locations. Insulators must be wiped, corroded metal parts cleaned, and arc-overs repaired. It is possible to detect poor contacts by inspecting for evidences of local overheating (or arcing). Such contacts must be thoroughly cleaned and tightened. Likewise, it is important that the antenna connections be regularly inspected and cleaned.

11-215. During radio silence the proper tuning of transmitters presents a problem, because an appreciable amount of rf energy must be prevented from radiating. Previous tuning of the transmitter through a great number of operating frequencies, with accurate logging of all the dial readings, eliminates the need for operation during radio silence. Interpolations are used if exact frequencies are not required. Oscillator frequencies which are not crystal-controlled are set by means of an appropriate frequency meter.

11-216. When the procedure mentioned above is not feasible, the following plan may be used if the transmitter is enclosed in a steel shelter with metal doors and window covers. Disconnect and ground all antenna leads entering the shelter. A receiver located outside the shelter should be tuned to ascertain whether there is any rf leakage. Preliminary stages may be tuned and neutralized while the power-amplifier stage is de-energized. To tune the power-amplifier stage, the antenna is ungrounded and connected to the transmitter, which is de-energized, and a receiver that is pretuned to the desired frequency is coupled to the power-amplifier stage. Tuning is accomplished by varying the tuning of the antenna, transmission line, and power-amplifier stage until a maximum random noise is measured on the signal-strength meter of the receiver.

11-217. FREQUENCY MEASUREMENTS.

11-218. GENERAL. Transmitter frequency measurements are necessary for reliable operation and compliance with communications regulations. These regulations concern harmonic frequencies, sidebands, and tolerances for the assigned frequency of transmission. There are a large number of frequency measuring equipments. Some, such as wavemeters, heterodyne frequency meters, Lecher wires, and resonant-coaxial-line frequency meters, are described in Section III of Chapter 8. Frequency measuring methods using calibrated receivers, wavemeters, and heterodyne frequency meters are discussed in Section III of Chapter 10.

11-219. PRIMARY FREQUENCY STANDARD. A primary frequency standard is shown in block diagram form in figure 11-24. A standard signal from a very stable crystal oscillator is fed to a series of four "locked-in" oscillators. These oscillators divide the standard frequency into submultiples of tenths, and also isolate the standard frequency oscillator from the output circuits. Frequency-dividing multivibrators can be used in place of the locked-in oscillators, although the accuracy of the latter is somewhat superior. The output of any of the locked-in oscillators can be selected as excitation for the harmonic generating stage,

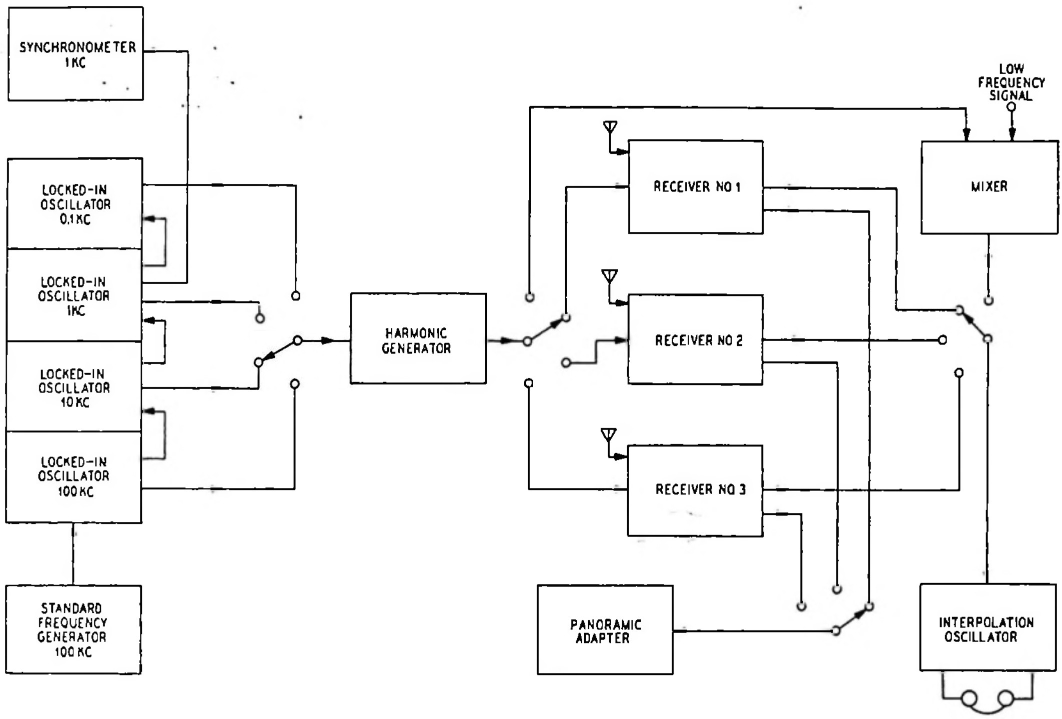


Figure 11-24. Primary Frequency Standard Equipment

but as a rule the output of the 10-kc oscillator is used.

11-220. If the output frequency of a transmitter is to be measured, the harmonic generator is coupled to the receiver or mixer component of the test set which covers the required frequency range. The output signal of the receiver will then contain the difference-frequency components obtained from the signal under test and the two nearest harmonics of the 10-kc comparator signal. The difference-frequency signals are applied to the interpolation oscillator, which is adjusted for a zero beat. Reference to the dial of the receiver is the usual way of deciding which harmonic is closest in frequency to the signal under test. In the event of any uncertainty, the display of the panoramic adapter can be observed. In this display, the various harmonics and the signal under test appear as vertical pips on the indicator; thus you can determine which harmonic is closest to the signal under test. The frequency of the signal under test is then determined by adding (or subtracting) the frequency obtained from the dial of the interpolation oscillator to (or from) the known harmonic frequency.

11-221. In order to check the standard frequency, a 1-kc synchronometer is included in the test set. This device, driven by a 1000-cps synchronous motor, is capable of accurately counting the number of cycles produced by the standard oscillator during a designated time interval. Readings are facilitated by a large, illuminated, 24-hour dial with a long sweep hand. There is also a microdial contractor that operates once each second. Calibrated in hundredths of a second, it makes comparisons with time signals accurate to one part in ten million over a 24-hour interval. The usual procedure is to compare the synchronometer with the time signals transmitted by Station WWV or WWVH. This shows how accurately the standard frequency has been maintained during the period between transmissions. Phasing of the microdial mechanism is accomplished by means of a panel control. A 60-cycle motor is provided in order to start the 1000-cycle synchronous motor when a push button on the panel is operated.

11-222. FREQUENCY MONITOR METHOD. Frequency monitors are secondary frequency standards that are useful for checking the output of transmitters which normally operate on one assigned frequency. Generally, a deviation within 30 cps of the assigned frequency can be read directly from a calibrated meter; in some instruments greater deviations are indicated.

11-223. This type of frequency meter also uses the heterodyne principle to measure frequency. This is done either by employing a crystal oscillator operating at the assigned frequency, and measuring any frequency difference between the transmitter and monitor directly, or by having the crystal oscillator frequency differ by a predetermined amount, usually 1000 cps, from the assigned frequency. In the latter case, the instrument is adjusted so that a half-scale deflection is produced by a 1000-cps signal, and the meter scale is calibrated to indicate zero frequency deviation at this point.

11-224. OSCILLOSCOPE METHOD. An oscilloscope is often employed for comparing frequencies. For such purposes, the oscilloscope method is an alternative to the heterodyne method of determining when two frequencies have an exact relationship. One advantage of the oscilloscope method is that it permits the establishment of a harmonic relationship without the necessity of generating the harmonic. The simplest procedure for using the oscilloscope to determine the harmonic relationship between two sine-wave voltages is to employ one frequency to produce the horizontal deflection, while feeding in another (or the same) frequency to the vertical deflection plates. The configuration of the resulting Lissajous figure depends upon the frequency ratio and upon the relative phase of the two waves. When the ratio of the frequencies is exactly a ratio of integers, the pattern is stationary and the ratio of horizontal to vertical frequency is the number of times the side of the figure is tangent to a vertical line, provided that the return traces do not coincide. Lissajous patterns become increasingly difficult to interpret for increasingly higher ratios.

11-225. An alternate method for measuring a frequency ratio by means of the oscilloscope (one that is suitable for determining higher ratios) is to use a resistance-capacitance phase splitter, as shown in figure 11-25, to produce a circular or elliptical pattern on the cathode-ray tube. Here, the lower frequency is impressed across the horizontal and vertical deflection plates, as shown, and the higher frequencies are used to control the intensity of the electron beam. The resulting pattern consists of a circle of spots in which the ratio of the high frequency to the low frequency is equal to the number of spots divided by a suitable integer, n. The value of n must be known, and it should be less than the number of spots.

11-226. POWER MEASUREMENTS.

11-227. GENERAL. The electrical power delivered to a load at any instant is equal to the product of the voltage across the load and the current passing through it, or P = EI. Under stable dc conditions, this product is also equal to the average power consumed. In ac circuits, on the other hand, the presence of either inductive or capacitive react-

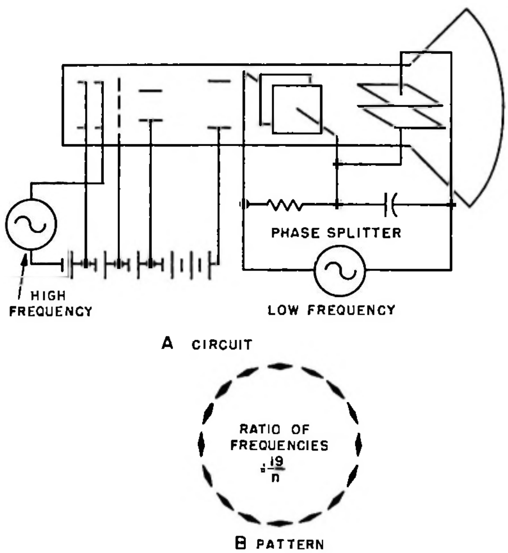


Figure 11-25. Method for Comparing Signals That Have a High Frequency Ratio

ance means that the apparent power, EI, where both voltage and current are rms values, must be multiplied by a number called the power factor to obtain the true power. Briefly, this is necessary because pure inductors and capacitors store energy furnished by the line, and, during a later portion of the cycle, restore the energy to the line. If purely reactive circuits were possible, therefore, none of the power would be dissipated. Naturally, resistance is an unavoidable component of any reactance, so from a strict standpoint nondissipative networks are not attainable. In practical ac circuits, the power dissipated is equal to the apparent power multiplied by the cosine of the phase angle between the voltage and current.

11-228. Calibrated voltmeters and ammeters are used as a direct approach to power measurements in dc circuits. Usually, approximate indications of power are satisfactory in electronic circuit work, so that neither the expense nor the inconvenience of two separate meters is justified. A typical instance is that of determining the power dissipated by a resistor. It is generally acceptable to measure the voltage across the resistor and then to calculate the power by the basic equation $P = E^2/R$. For current rather than voltage measurement, the power is equal to I^2R. If the designated resistance of the part is not sufficiently reliable, it may be measured approximately with an ohmmeter. Conceivably, a more accurate resistance measurement may be desired, in which case a determination by means of a Wheatstone bridge can be made. Similarly, it may be necessary to determine the ac power dissipated by a resistive load, either at audio or radio frequencies. The same method is reliable, provided that the resistance of the device is known at the frequency of operation.

11-229. AUDIO-FREQUENCY POWER MEASUREMENTS. To test the output capacity of modulator amplifiers, you may choose one of several types of power meters. The usual components of such instruments are a variable-ratio transformer, a constant-resistance multiplier, and a voltmeter with a copper-oxide rectifier. The use of the transformer, which is loaded by various resistances, enables the effective load imposed on the output stage to be varied over a number of steps. The constant-resistance multiplier acts as a range multiplier for the voltmeter, while presenting a constant resistance to the secondary of the transformer. The indicating instrument is calibrated directly in watts.

11-230. When reactive components of the dissipative impedance introduce a phase angle, the foregoing methods are no longer applicable. A device proportional to the power factor as well as to the apparent power must be used instead. Accordingly, a large number of instruments, called wattmeters, have been successfully developed for low-frequency power measurements.

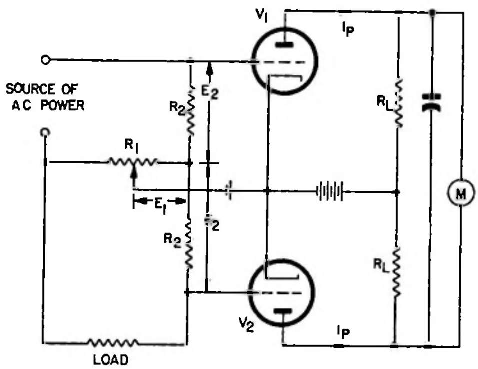


Figure 11-26. Electronic Wattmeter Circuit

11-231. ELECTRONIC WATTMETER METHOD. This method uses an electronic wattmeter circuit based on the balanced modulator principle. As shown in figure 11-26, a source of ac power is connected to the load through the series resistors, R2. These two resistors which are of equal value, are made small enough to keep the drop across them from reducing the load voltage appreciably. In contrast, resistor R1 is made very large so that its power consumption is negligible. These restrictions make the voltage across R1 equal to the load voltage, and the voltage across either series resistor proportional to the load current. Inspection of this circuit shows that the voltage across the grid circuit of the upper tube is E1 plus E2, while the net voltage applied to the lower tube is E1 minus E2. In the output circuit, the resistances in the plate circuits of the tubes are equal. This provision makes the difference of potential between the plates proportional to the difference in the output currents of the tubes. The average value of the difference is indicated by the dc meter connected to the plates. For the circuit to function as a wattmeter, the tubes must be operated over the nonlinear portion of their characteristic curve. This operation causes the difference current to consist of a number of components. The components proportional to either E1 or E2 make no contribution to the reading, as the average value of a sine wave is zero. (E1 and E2 are rms values of sine-wave voltages.) The only other appreciable component is the one proportional to the product of E1 and E2, that is, proportional to the product of the load voltage and current. The average value of such a component is proportional to the product of E1 and E2 multiplied by the cosine of their phase differences. Consequently, the meter reading is proportional to the power consumed by the load, and the scale is therefore calibrated in watts. Nonlinearity of the scale may be minimized by using inputs of low value to the tubes.

11-232. FREQUENCY LIMITATIONS. The preceding methods for the direct measurement of ac power are satisfactory only when the frequency is fairly low. As the frequency increases, a number of serious complexities are introduced to the problem of measurement. The most serious of these is that of stray capacitance and inductance. Other problems are skin-effect resistance, the problem of determining the power factor at high frequencies, and the difficulties encountered in measuring large magnitudes of rf power. To overcome these conditions, other methods of measurement have certain advantages in power measurement at higher frequencies.

11-233. PROGRAM CIRCUIT POWER-LEVEL MEASUREMENTS. Whenever speech and music intelligence must be transferred by means of telephone circuits, it is advantageous to keep the audio power level within prescribed limits. This limitation is made to prevent crosstalk and to minimize the effort required in controlling the transmission. The monitoring of the audio-frequency energy involved is ordinarily accomplished by bridging the transmission line or circuit by a high-impedance rectifier-type ac voltmeter having negligible effect upon the circuit under measurement.

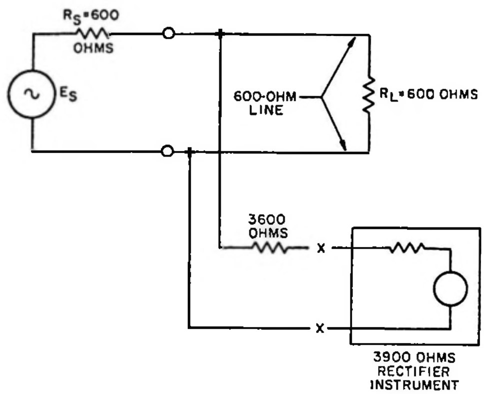


Figure 11-27. Power-Level Indicator for Program Circuits

11-234. Speech and music modulation cause rapid fluctuations in the intensity of the program-circuit currents; for this reason, standards have been established regarding the dynamic characteristics of the power-level indicators used for this type of circuit. Figure 11-27 shows a standard power-level indicator. The rectifier-type voltmeter has an input impedance of 3900 ohms, and is connected in series with a 3600-ohm resistor; thus a total resistance of 7500 ohms is shunted across a 600-ohm transmission line. The ballistic characteristic of the power-level indicator in this circuit is such that a peak application of a single frequency will cause the meter pointer to overswing by 1 to 1.5 percent, and to reach 99 percent of the steady-state deflection in 0.3 second. Because of this damping characteristic, the indicator reads the mean-power level based on short-time average amplitudes, rather than instantaneous power levels based on peak voltages. The scale of the standard power-level meter is calibrated with a steady sine-wave voltage in terms of dbm. The deflections obtained are customarily given in volume units (vu's), which are numerically equal to the number of decibels above a reference level of 1 millwatt in 600 ohms. The level in vu is therefore equivalent to a definite power rating which may be converted into the equivalent power in watts, and represents (for speech) the power level corresponding to the average power level of the speech.

11-235. RF POWER MEASUREMENT. Because even a substantial reduction of power does not necessarily decrease the operating range significantly, the precise measurement of rf power may appear to be unnecessary. This is a false conclusion. One reason is that a change in the power output may result from altered operating conditions that are capable of causing equipment breakdown unless remedied. In addition, power measurements are often the surest way of determining whether the over-all performance of a transmitter is normal and in general consistent with the specifications of the manufacturer. Tests should therefore be made periodically, either with the same equipment or with different instruments of equal accuracy. If this practice is observed, a change in the indication will reliably signify trouble rather than a discrepancy between the instruments. Methods for measuring the rf output power of radio transmitters are described in Section II of Chapter 10.

11-236. MF TO VHF RANGE. Relative power output readings are often acceptable for certain communications transmitters. These checks are simple, and generally employ meters which are integral with the equipment. The dc voltage and current for the final rf amplifier stage is measured. The dc input power to the stage is then calculated from the Ohm's law relationship of P = EI, where:

P (watts) = relative output power level

E = plate voltage of the transmitter tube

I = plate current of the transmitter tube

This method is most useful when the final transmitter stage is a Class C amplifier, since for other types of amplifiers errors become large. You must also make certain that the transmitter is properly tuned before this measurement is made; otherwise, the test is meaningless. This check is particularly suitable for transmitters that are operated over a large range of frequencies, and where the antenna radiation resistance may be unknown at the frequency of operation. Although this is a relative method of measurement, and is not accurate as a power output indication because of the loss of power in the antenna coupling networks, it can be performed while the transmitter is in operation to provide an indication of transmitter performance. For these reasons, this type of test is included in many performance checks for transmitter equipment.

11-237. Another check for power output of some communications transmitters is obtained from direct readings of thermal-coupled ammeters, calibrated to indicate power. This type of meter is connected in series with the antenna output terminals, and will give an accurate indication only when the antenna impedance is at a value for which the meter has been calibrated. An application of this method is to substitute a low-reactance resistor of proper wattage (dummy antenna) for the antenna. If the thermal-coupled meter is calibrated to indicate current or voltage, power can be accurately determined by the relationship

$P = I^2R$ or $P = E^2/R$, where:

P = power output

R = resistance of dummy antenna

I = effective current

E = effective voltage

11-238. UHF AND SHF RANGE. Indirect power measurements are generally employed for equipment operating in the ultra-high-frequency and super-high-frequency range. Invariably, these methods convert rf energy into another form of energy, such as light or heat, which can be evaluated more readily.

11-239. TRANSMISSION LINE POWER TESTS. Since the impedance along a perfectly matched transmission line is equal to the characteristic impedance of the line, the power transmitted to the load can easily be determined. In lines having low loss at radio frequencies, the impedance is resistive, and can be calculated from the dimensions of the line. The power on such a non-resonant line is I^2R or E^2/R, where:

R = the characteristic impedance of the line

I = the rms value of the current flowing along the line

E = the rms value of the voltage across the line

The power measurements made in this manner will be accurate provided that the arrangement employed to measure the voltage or current on the line does not introduce an appreciable standing wave.

11-240. A thermocouple ammeter connected in series with the line can be used for measurement of the current at almost all radio frequencies; also, a diode-type vacuum-tube voltmeter can be employed to measure the voltage across the transmission line, provided that the frequency is low enough that the input capacitance of the voltmeter shunted across the line has a reactance that is very much higher than the characteristic impedance of the line. A method of minimizing the undesirable effect of input capacitance of a diode voltmeter at high frequen-

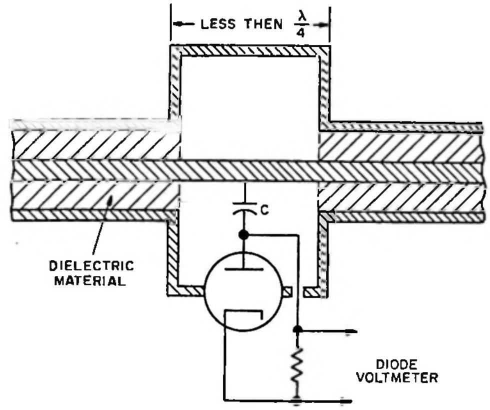


Figure 11-28. Method for Determining Voltage Across Coaxial Transmission Line

cies (in the case of coaxial lines) is illustrated in figure 11-28. The diode is shunted across a specially designed section of the line, which has greater spacing between the inner and outer conductors than the line, and which has a length somewhat less than a quarter of a wavelength at the highest frequency for which the arrangement is to be employed. The enlarged spacing increases for inductance per unit length of line, while reducing the capacitance per unit length; the absence of dielectric material in the enlarged section further reduces the capacitance. The ratio of the capacitance to the inductance of the special section should be the same as for the line, so that the special section will introduce only negligible reflection. The special section will maintain essentially the same characteristics for frequencies appreciably less than that for which the special section is designed, and its behavior is independent of frequency over this range.

11-241. MODULATION MEASUREMENTS.

11-242. GENERAL. Measurements and calculations to determine the percentage of amplitude modulation and the index of frequency modulation are described in Section V of Chapter 10. More complex methods of determining the index of frequency modulation are discussed in this section.

11-243. FREQUENCY MODULATION MEASUREMENTS. As explained in paragraph 10-457 of Volume VIII, the concept of percentage of modulation as discussed in connection with amplitude modulation does not apply to frequency modulation. The amplitude of the fm wave is constant, and the extent of modulation must be described in other terms than those of the amplitude-modulated wave. When referring to a class of stations, a certain maximum frequency swing is established as representing 100 percent modulation. For example, in the case of fm broadcast stations, a frequency swing of $\pm$ 75 kc from the unmodulated center frequency (frequency deviation) is commonly considered as being the equivalent of 100 percent modulation. However, the more widely accepted method of describing the extent of modulation is to state the value of the modulation index. This index (m) is simply the ratio of the amount by which the transmitted frequency swings from its average frequency (frequency deviation) to the frequency of the modulating signal. The relationship of these quantities is shown by the following equation:

$$m = \frac{F_d}{F_m}$$

where:

m = modulation index

F_d = frequency deviation

F_m = frequency of modulating signal

By means of this basic relationship, it is possible to determine the frequency deviation when the modulation index and the modulating frequency are known.

11-244. It should be carefully noted, in describing the extent of frequency modulation, that the modulation percentage and the modulation index are defined in a different manner. The percentage is proportional to the frequency swing. The modulation index is also directly proportional to the frequency swing, but in addition, is inversely proportional to the highest modulating frequency. Thus, in contrast to amplitude modulation, the modulation index of a frequency-modulated wave is not the decimal equivalent of the modulation percentage. The modulation index of a frequency-modulated wave, for example, will exceed 1 by many times when the frequency swing is large and the modulating frequency is low.

11-245. The frequency-modulated output is the sum of a center frequency component and numerous pairs of sideband frequency components. The center frequency component has the same frequency as the unmodulated carrier. The two components of the first sideband pair have frequencies respectively higher and lower than the center frequency by the amount of the modulating frequency, just as in amplitude modulation. In frequency modulation, however, there are additional pairs of sideband components which can have appreciable amplitude. For example, the second pair of sidebands, having frequencies that are higher and lower than the center frequency by twice the amount of the modulating frequency, can also be important. The same can be true of the third pair of sidebands, which are removed from the center frequency by three times the modulating frequency, and of higher orders of sideband pairs, whose frequencies differ from the center frequency by correspondingly greater amounts.

11-246. When the modulation is slight, only the pair of sidebands nearest in frequency to the carrier frequency component will have sufficient amplitude to be important. Under this condition, the bandwidth required is no greater than for an amplitude-modulated wave. As the frequency modulation is increased, however, more pairs of sidebands acquire appreciable amplitude, and the bandwidth requirements are greater than for amplitude modulation.

11-247. The actual amplitudes of the frequency-modulated-wave sidebands and carrier, as compared with an unmodulated carrier amplitude of 1, may be read directly from table 11-1 for modulation indices up to 6. To find the amplitude of any sideband pair, determine the modulation index (m), read the corresponding amplitude factor for the sideband pair, and multiply this factor by the amplitude of the unmodulated carrier. The amplitude of the carrier during modulation is found in the same manner, taking the amplitude factor from the $J_0(m)$ column. Where no value is given in a column, the amplitude factor is less than 0.005, and the sideband pair will not be important for normal considerations.

11-248. The values of $J_0(m)$, $J_1(m)$, and $J_2(m)$ over the range m = 0 to m = 16 are plotted in figure 11-29. A study of these curves reveals some interesting facts about the composition of frequency-modulated waves. $J_0(m)$ is less than 1 for all values of m greater than zero. This indicates that as sideband components appear with modulation, the amplitude of the center frequency component is less than its amplitude in the absence of modulation. This fact is evident when it is remembered that the amplitude of the frequency-modulated wave is constant, so that the average power during each radio frequency cycle is the same as that during any other radio frequency cycle. In order that the power in the wave will not change when frequency modulation causes sideband currents to appear, the amplitude of the center frequency component must decrease sufficiently to keep the total of the I^2R products of all the components equal to the power of the unmodulated wave.

Table 11-1. Bessel Factors for Finding Amplitudes of Center and Sideband Frequency Components

MODULATION INDEX m	$J_0(m)$ F	$J_1(m)$ $F \pm Fm$	$J_2(m)$ $F \pm 2Fm$	$J_3(m)$ $F \pm 3Fm$	$J_4(m)$ $F \pm 4Fm$	$J_5(m)$ $F \pm 5Fm$	$J_6(m)$ $F \pm 6Fm$	$J_7(m)$ $F \pm 7Fm$	$J_8(m)$ $F \pm 8Fm$	$J_9(m)$ $F \pm 9Fm$
0.0	1.000									
0.1	0.9975	0.0499								
0.2	0.99	0.0995								
0.3	0.9776	0.1483	0.0112							
0.4	0.9604	0.196	0.0197							
0.5	0.9385	0.2423	0.0306							
0.6	0.912	0.2867	0.0437							
0.7	0.8812	0.329	0.0589	0.0069						
0.8	0.8463	0.3688	0.0758	0.0102						
0.9	0.8075	0.4059	0.0946	0.0144						
1.0	0.7652	0.4401	0.1149	0.0196						
1.2	0.6711	0.4983	0.1593	0.0329	0.005					
1.4	0.5669	0.5419	0.2073	0.0505	0.0091					
1.6	0.4554	0.5699	0.257	0.0725	0.0150					
1.8	0.3400	0.5815	0.3061	0.0988	0.0232					
2.0	0.2239	0.5767	0.3528	0.1289	0.034	0.007				
3.0	-0.2601	0.3391	0.4861	0.3091	0.1320	0.0430	0.0114			
4.0	-0.3971	-0.066	0.3641	0.4302	0.2811	0.1321	0.0491	0.0152		
5.0	-0.1776	-0.3276	0.0466	0.3648	0.3912	0.2611	0.131	0.0534	0.0184	
6.0	0.1506	-0.2767	-0.2429	0.1148	0.3576	0.3621	0.2458	0.1296	0.0565	0.0212

11-249. Regardless of the differences between amplitude and frequency modulation, it is possible to make an analogy between percentage of amplitude modulation and frequency deviation. Specifically, frequency deviation is proportional to the amplitude of the modulating signal, as is the percentage of amplitude modulation. Because of this analogy, it is convenient to extend the concept of percentage of modulation to frequency modulation by arbitrarily designating the maximum allowable frequency deviation of a class of operation as 100 percent modulation. An important distinction to remember is that no distortion results from modulation percentages greater than 100 in fm transmissions. However, any percentage larger than the figure sanctioned by the proper authorities will produce excessive channel width, making interference with other stations possible. For example, the maximum frequency deviation for commercial fm sta-

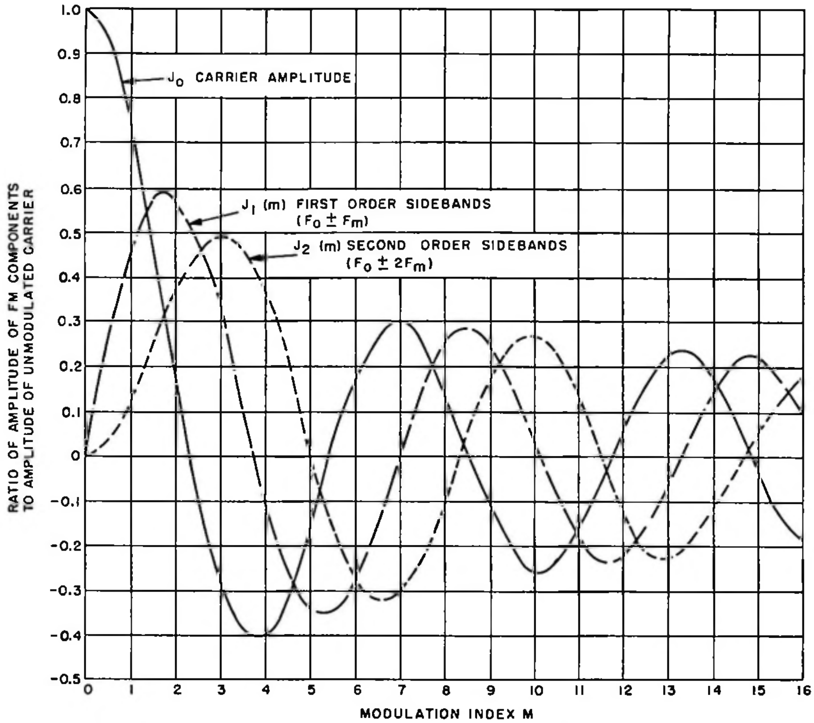


Figure 11-29. Variation of FM Wave Components with Degree of Modulation

tions is limited to 75 kc; for military applications the maximum deviation is limited to 40 kc (and is classed as narrow-band fm transmission), and the sound transmission of television stations is restricted to a deviation of 25 kc. A simple procedure for frequency deviation measurement is discussed under paragraph 10-460 of Volume VIII. The Bessel zero method, which is discussed below, is somewhat more complicated.

11-250. BESSEL ZERO METHOD. If some of the more elaborate test equipment designed especially for monitoring fm transmissions is not available, a technique known as the Bessel zero method enables the determination of frequency deviation by use of the following equipment: a variable-frequency (low-distortion) audio oscillator and a communications receiver, which is tunable to the carrier frequency of the fm transmitter and which includes a beat-frequency oscillator.

11-251. It was stated earlier that the modulation index (m) determines the relative amplitude of the carrier and sideband frequencies emitted by an fm transmitter. The modulation index may be measured by utilizing the fact that the carrier amplitude becomes zero whenever the modulation index is such that $J_0(m) = 0$, where J_0 is a Bessel function of the zero order. The values of modulation index for these conditions are given in table 11-2.

11-252. Specifically, the carrier component disappears completely for certain values of m, ie, m = 2.405, 5.52, 8.654, etc. (Note that $J_0(m) = 0$ in figure 11-29 for these values of m.) For these specific values of m,

Table 11-2. Values of Modulation Index for Which Carrier Wave Has Zero Amplitude

ORDER OF CARRIER ZERO	MODULATION INDEX
1	2.405
2	5.52
3	8.654
4	11.79
5	14.93
6	18.07
m (m > 6)	18.07 + π (m − 6)

all of the transmitter power is contained in the sidebands. This fact allows the measurement of specific values of modulation index by adjusting a receiver to measure the amplitude of the carrier component only. The modulation on the fm carrier is increased from zero to the first point at which the detected carrier disappears. The point at which the carrier first disappears corresponds to m = 2.405. Upon increasing the modulation further, the carrier reappears and then disappears a second time. The second vanishing of the carrier corresponds to m = 5.52. Further increases in modulation will produce the higher carrier zeros (or null points). The frequency deviation at the first null point, for example, is $F_d = 2.405 F_m$.

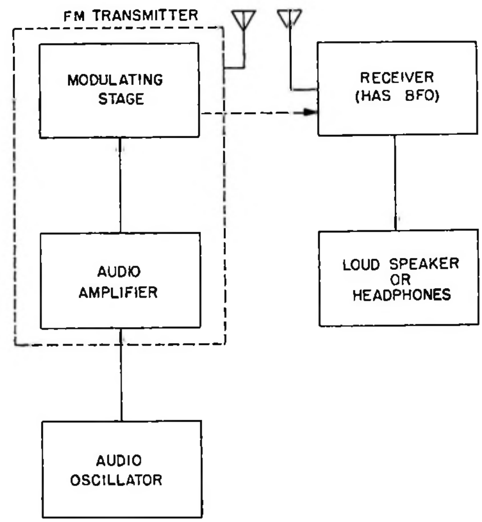


Figure 11-30. Test Arrangement for Measurement of Frequency Deviation by Bessel Zero Method

11-253. TECHNIQUE FOR FINDING ZERO POINTS. Figure 11-30 shows the general test arrangement for measurement of the frequency deviation by the Bessel zero method. The communications receiver used for the Bessel zero method of measuring frequency deviation must incorporate a bfo, and must be tunable to the fm transmission frequency. First, the receiver is tuned to an unmodulated transmission of the equipment under test. In order to produce an audible note, it is necessary to adjust the bfo until a tone, such as 500 cps, is plainly heard in the output of the receiver. Once the pitch of this note is set, the bfo adjustments should be fixed so that the ear can become accustomed to this particular pitch. This is an important requisite. Later, the frequency deviation will be varied by changing the amplitude of a modulating signal, diverting power from the carrier into various sideband frequencies. As a result, many conflicting tones will be introduced in the output of the receiver, but they must be ignored completely at all times. Practice is recommended so that you will become adept at judging when the beat note produced by the carrier vanishes. The use of a low-pass filter is often helpful, although not absolutely necessary.

11-254. An audio frequency signal from a suitable oscillator is now applied to the audio-amplifier section of the transmitter. If the amplitude of the signal is increased, the sideband frequencies and their innumerable beat notes are produced. Note, in particular, the gradual attenuation of the beat note arising from the carrier. At some point this beat note should disappear entirely, only to reappear once more as the modulating signal is increased further. In general, the modulating signal should never be increased to the point where any stage of the transmitter becomes saturated.

11-255. When the receiver beat-frequency-oscillator (bfo) frequency is less than the modulating frequency, as is sometimes required, you should avoid ratios of modulation to bfo frequency such as 1/2, 1/3, 1/4, etc, to avoid possible error. For example, if the modulation frequency is to be 1000 cps, and the bfo is adjusted to give a 500-cps beat note with the unmodulated carrier, the application of modulation will produce another 500-cps beat note between the bfo and the first sideband component on the bfo side of the carrier. Such a condition could be encountered, especially if the receiver contains appreciable distortion. If a relatively high modulating frequency (for example, 10 kc) is used, the likelihood of spurious audible beats is very small.

11-256. ESTABLISHING A MAXIMUM FREQUENCY DEVIATION. Since the frequency deviation is proportional to the amplitude of the modulating signal, it is evident that the modulating signal cannot be greater than a certain amplitude. This amplitude can readily be ascertained by the Bessel zero method.

11-257. NARROW-BAND TEST PROCEDURE. If a transmitter output signal must be limited to a frequency deviation of 25 kc, you should first find the modulating frequency that produces a modulation index of 2.405 when the frequency deviation is 25 kc. An audio signal of this frequency should then be applied to the fm transmitter, and increased in amplitude until the carrier is completely suppressed, as evidenced by the disappearance of the carrier beat note in the receiver. At this point the modulation index is 2.405 and the corresponding frequency deviation is given by the formula: $F_d = 2.405F_m$. This relation, solved for F_m, will indicate the audio frequency to be employed. For example, if the maximum frequency deviation is 25 kc, then $F_m = F_d/m = 25/2.405 = 10.395$ kc.

11-258. After finding the above quantity, proceed as follows:

a. Set the audio oscillator to 10.395 kc.

b. Apply this signal to the audio amplifier of the transmitter. It is assumed that the amplifier is coupled to the modulating stage of the transmitter.

c. While listening attentively to the beat note produced by the carrier component of the fm transmission, steadily increase the amplitude of the audio signal. When the beat note disappears completely, a frequency deviation of 25 kc has been reached. This indicates, then, that the corresponding amplitude of the modulating signal must never exceed this figure during the operation of the transmitter.

11-259. WIDE-BAND TEST PROCEDURE. When a large frequency deviation (such as 75 kc) is to be established, the process is more involved. Repeating the procedure of computation used previously, it is found that $F_m = 75/2.405 = 31.2$ kc. This frequency, however, is too high for the audio stages of the transmitter. There are two ways of overcoming this difficulty. One is to use the second value of the modulation index at which the carrier disappears. This value (5.52) leads to the relationship $F_m = 75/5.52 = 13.586$ kc. Such a frequency is within the audio bandpass of many fm transmitters, but not all.

11-260. Assuming that 13.586 kc is not too high, proceed as follows:

a. Set the audio oscillator to 13.586 kc.

b. Apply the signal to the transmitter in the same way as in paragraph 11-257.

c. Steadily increase the audio signal until the carrier disappears, as indicated by the disappearance of the carrier-produced beat note. The first point of carrier disappearance corresponds to a relatively low frequency deviation.

d. Continue to increase the audio signal. When the carrier disappears for the second time, the frequency deviation of the transmitter is 75 kc.

11-261. If the audio frequency (13.586 kc) is too high, recourse is made to another method. This procedure makes use of the fact that a frequency-multiplying stage increases both the center frequency and the frequency deviation by the same factor. For example, in a transmitter incorporating a reactance tube, the frequency generated by the oscillator might be increased eight times before application to the power-output stage. In such a transmitter, a frequency deviation of 9.375 kc at the oscillator corresponds to a frequency deviation of 75 kc. In order to make use of this property, a portion of the oscillator signal should be coupled to the receiver, as indicated by the dotted lines shown in figure 11-30. The receiver, of course, must be tuned to the oscillator frequency. With this setup, it is necessary to find the amplitude of audio signal that produces a frequency deviation of 9.375 kc. The frequency of the modulating signal is determined by the relationship $F_m = 9.375/2.405 = 3.9$ kc. After determining the above quantity, proceed as follows:

a. Couple the receiver by means of a small pickup coil to the modulated-oscillator stage of the transmitter.

b. Tune the receiver to the oscillator frequency.

c. Set the audio oscillator to 3.9 kc.

d. Increase the amplitude of the signal until the carrier-produced beat note in the receiver disappears. This amplitude corresponds to a frequency deviation of 75 kc.

11-262. MONITORING METHODS. Although of great help when special monitoring equipment is unavailable, the Bessel zero method cannot compete with the convenience afforded by the many frequency and modulation monitors used with transmitting sets. In addition to determining the percentage of modulation, these elaborately designed instruments perform a number of other functions. Depending upon the instrument, these other functions may include monitoring the center frequency of transmission, monitoring the modulating signal, measuring distortion and noise, and determining the a-m noise content. In contrast to the lengthy procedure of the Bessel zero method, monitoring equipments have meters and lamps that provide direct indications.

11-263. AUDIO DISTORTION MEASUREMENTS.

11-264. GENERAL. The faithful reproduction of the audio components of a transmitted wave is achieved only when the over-all distortion is of a relatively low percentage. Although the FCC allows the harmonic distortion of a commercial a-m transmission to be as high as 10 percent when the modulation level is 85 percent, the distortion produced by many transmitters is as low as 2 percent. Although very few receivers at present are in the extreme high-fidelity class, the majority of them will readily disclose poor quality in a transmission. Modern transmitters use carefully balanced circuits extensively. In these balanced circuits, small changes in tube characteristics or the values of critical parts can produce an intolerable increase in distortion. There are numerous devices available for measuring distortion, and, through proper testing and maintenance, it is possible to keep distortion to a very low value.

11-265. ANALYSIS OF HARMONIC DISTORTION. The distortion represented by a particular harmonic is simply the ratio of the harmonic to the fundamental frequency expressed as a percentage. Total harmonic distortion is expressed as the root-mean-square sum of all the harmonics present in the signal. If, for example, the second harmonic contributed 4 percent distortion, the third harmonic 6 percent, and the fourth harmonic 5.5 percent, the total distortion (D_T) would be as follows:

$$D_T = \sqrt{16 + 36 + 30.25}$$

$$= \sqrt{82.25} = 9.077$$

The value of 9.07 percent represents the total rms harmonic distortion.

11-266. There are four methods of harmonic analysis in use: numerical method, schedule method, mechanical method, and direct method. The direct method is used in electrical systems. In the direct method, instead of first getting an oscillograph record of the current or voltage and then analyzing it mathematically, the analysis is performed directly by the apparatus. The following paragraphs are concerned with the various electronic instruments which are used as harmonic wave analyzers.

11-267. One of the most important uses of harmonic wave analyzers is in the harmonic analysis of the output of an electronic amplifier. In an ideal amplifier, the amplified output is exactly proportional to the input. For example, if the input is a pure sine wave, the output is a pure sine wave. The nonlinear characteristics of tubes, transistors, and transformers cause the output of an amplifier to depart somewhat from the ideal. The principal effect of nonlinear distortion is to cause the amplified output to contain frequency components not present in the signal applied to the amplifier. Thus, when the applied signal is a sine wave, nonlinear distortion will cause the output to contain harmonics of the input frequency, particularly the second and third harmonics. When two sine waves are fed to an amplifier, sum and difference frequencies are also produced. Figure 11-31 illustrates the distorted output together with its components, the fundamental and the first two harmonics, for an amplifier that has a single sine-wave input. The fundamental is an amplified reproduction of the input; however, the amplifier has also produced harmonics which, together with the fundamental, add up to an irregular curve. Since distortion usually increases with an increase in amplifier output, the tolerable distortion sets a practical limit to the output of a given amplifier.

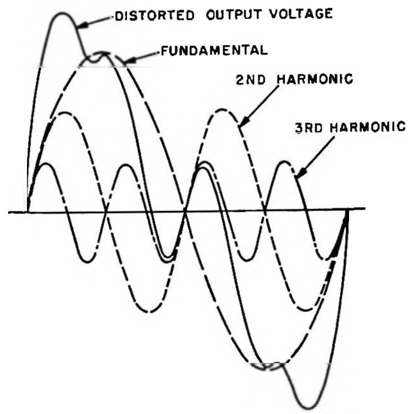


Figure 11-31. Distorted Sine Wave, Showing Fundamental Plus Second and Third Harmonic Components

11-268. The two most common methods of measuring nonlinear distortion directly are the harmonic analysis method and the intermodulation method. The harmonic analysis method consists of applying a sine wave to the input of an amplifier and examining the output for harmonics. In the intermodulation method, two sine wave voltages of different frequency are applied to the input of the amplifier, and the output examined for sum, difference, and various combination frequencies. It is of interest, however, to note that the results obtained by one method may be converted to the results of the other.

11-269. Four general types of test equipment are used for the measurement of the harmonic distortion contained in a signal. They are the tuned circuit analyzer, the heterodyne analyzer, the dynamometer analyzer, and the fundamental-suppression analyzer. The discussion to follow provides a brief description of these instruments and sets forth some of the advantages and disadvantages involved in their applications.

11-270. TUNED-CIRCUIT HARMONIC ANALYZER. The harmonic content of a wave can be determined by a method which makes use of a tuned circuit, as shown in figure 11-32. The complex wave to be analyzed is applied to a selective network, the essential feature of which is a sharply tuned resistance-inductance-capacitance circuit whose frequency is controlled by varying the capacitance. Theoretically, a tuned circuit will pass one frequency and reject all others. Since this is not possible in actual practice, these circuits are designed to tune as sharply as possible and pass a very narrow band of frequencies. The narrowness of this band constitutes the selectivity of the circuit. A maximum response of the circuit (greatest current flow) occurs at each frequency of tuning which coincides with a component of the complex wave. By tuning the circuit successively to the frequencies of

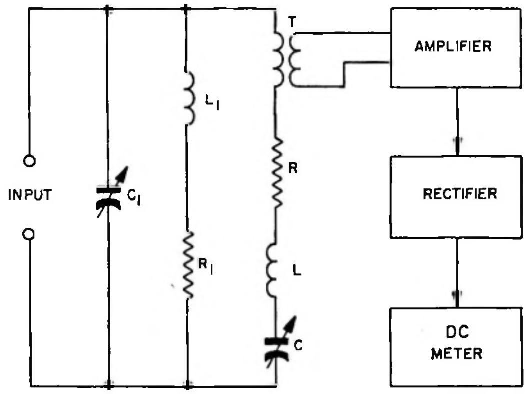


Figure 11-32. Tuned-Circuit-Type Harmonic Analyzer

the various harmonics in the complex wave and observing their amplitudes, the wave can be analyzed. The instrument is calibrated by introducing known frequencies of known amplitudes. The series-resonant circuit consisting of inductor L and capacitor C is tuned to a specific harmonic frequency, and, by means of transformer T, this harmonic component is applied to an amplifier. The output of the amplifier is rectified and used to actuate a dc meter. After a reading is obtained, L and C are tuned to the next harmonic, and another reading is taken. The parallel-resonant circuit composed of L1, R1, and C1 provides compensation for the variation in the ac resistance of the series circuit, and for the variation of the amplifier gain over the frequency range of the instrument. Consequently, the sensitivity of the equipment is nearly the same at all frequencies.

11-271. There are numerous modifications of the basic instrument just described. For example, equalizing networks are generally used in place of the compensating parallel-resonant circuit. Frequently, the transformer is omitted, in which case the amplifier is excited by the voltage across either

L, C, or R. The usual choice is the inductor L, since the increase in its impedance with frequency offsets the usual decline in the amplitude of the higher harmonics. Types of harmonic analyzers involving only tuned circuits have been generally abandoned because of the elaborate switching arrangements necessary to obtain complete audio-frequency coverage.

11-272. There are other major disadvantages of the tuned-circuit method. One is the difficulty encountered at the lower frequencies, necessitating the use of large component values in the tuned circuit. The harmonics of the signal, moreover, are often so close in frequency that they cannot be distinctly separated. In many instances, it is inconvenient to measure each component harmonic individually instead of taking a single reading for the total harmonic distortion. The variation of the tuned-circuit impedance with changing frequency is often another troublesome problem. Lastly, this type of circuit is susceptible to magnetic pick-up and interference.

11-273. HETERODYNE HARMONIC ANALYZER. The principle of the heterodyne analyzer is well known, and is used in the superheterodyne receiver. The heterodyne method of wave analysis is by far the most common one in use today. In the heterodyne type of analyzer, the difficulties of the tuned-circuit method are avoided by the use of a highly selective, fixed-frequency filter. The output of a variable-frequency oscillator is heterodyned successively with each harmonic of the input signal, and either the sum or difference frequency is made equal to the frequency of the filter. As a result of converting each harmonic to a constant frequency, it is possible to use extremely selective filters, often of the quartz-crystal type. By using these filters, only the constant-frequency signal corresponding to the particular harmonic under test is passed to the amplifier. The most critical part of the analyzer is the amplifier. Since the analyzer is used to examine the properties of other amplifiers, it must be much better than any of the amplifiers to be tested. This problem is handled by fixing the amplifier at one frequency and then filtering the input of the amplifier so that only that frequency reaches the amplifier.

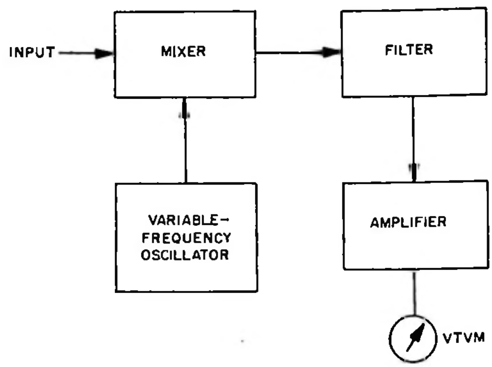


Figure 11-33. Heterodyne-Type Harmonic Analyzer

11-274. Figure 11-33 shows, in block diagram form, the essentials of a heterodyne-type harmonic analyzer. A balanced modulator is commonly employed as the mixing device, since it offers a simple means of lessening the amplitude of undesired components which would otherwise cause errors in the measurement. Another advantage is the low harmonic distortion generated by a balanced modulator as compared with other types of mixers. In addition to quartz-crystal filters, inverse feedback filters also achieve excellent selectivity. A balanced electronic voltmeter generally serves as the indicating device. Some heterodyne analyzers are calibrated to provide direct readings, and in others the harmonics of the impressed signal are compared with a reference voltage, usually by making the latter equal to the amplitude of the harmonic.

11-275. DYNAMOMETER-TYPE ANALYZER. Another method of harmonic analysis, which is well adapted to the analysis of power frequencies and the lower audio frequencies, is the dynamometer-type analyzer method. Briefly, a dynamometer is a device which compares the magnetic force between one moving coil and at least two stationary coils.

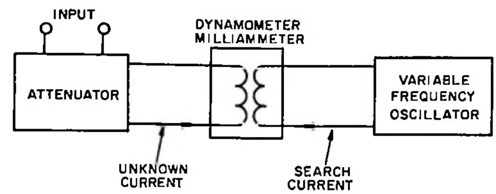


Figure 11-34. Dynamometer Harmonic Analyzer

11-276. The current to be analyzed is passed through one coil of a dynamometer-type milliammeter, while a search current of controllable frequency is passed through the other coils. The operation of the device depends upon the fact that the instrument pointer will not be deflected unless the frequency of the search current is very close to the frequency of a component contained in the wave being analyzed. When the difference between the two frequencies is a fraction of a cycle per second, the pointer will pulsate at the difference frequency. The amplitude of the pulsation is equal to the product of the effective values of the search current and the component of the wave being analyzed. To analyze the wave, the complex wave, after suitable amplification, is applied to one fixed coil of the dynamometer (figure 11-34), and the output of a variable-frequency (search) oscillator is applied to the other. When the frequency of the search oscillator is extremely close to a harmonic of the waveform under test, the moving coil and the indicator attached to it oscillate at the difference frequency. The maximum deflection is proportional to the product of the currents in the two coils (that is, to the oscillator current and to the harmonic component). By holding the oscillator current constant with the aid of a meter, the deflection of the pointer is made proportional to the harmonic alone. A wave is analyzed, therefore, by varying the frequency of the search oscillator and noting both the frequency at which the beats occur and the amplitude of the deflection. Dynamometer analyzers can also be designed to give steady deflection. Instruments of this type, however, are extremely limited in usefulness. One restriction is the inability to analyze frequencies above approximately 3 kc. Another problem is the difficulty in keeping the oscillator signal exactly in phase with the waveform under test.

11-277. The dynamometer-type harmonic analyzer has the following desirable features: direct reading, easily calibrated, simple and rapid in operation, apparatus easily portable, and relatively inexpensive. This type of analyzer is limited to the lower audio frequencies because dynamometer instruments will not operate satisfactorily at higher frequencies, in view of the allowable dissipation in the coils. For this reason, the dynamometer method has not been widely used. Often, in harmonic analysis, the fundamental will be as much as two hundred times as large as the second harmonic. Removal of the fundamental and its separate measurement greatly facilitate accurate measurement.

11-278. FUNDAMENTAL-SUPPRESSION ANALYZER. If the principal consideration is the total harmonic distortion, rather than a knowledge of individual components, the fundamental-suppression method of measuring distortion is most generally preferred. In most of the other methods of harmonic analysis, the higher harmonics are measured in the presence of the fundamental. Although

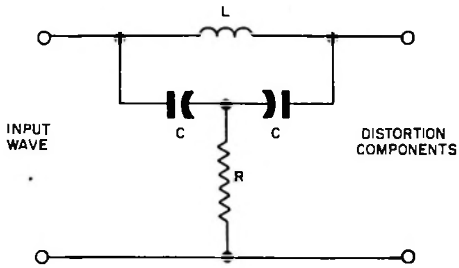


Figure 11-35. Balanced-T-Type Suppression Network

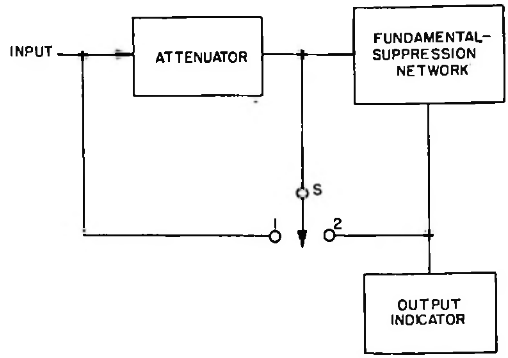


Figure 11-36. Fundamental Suppression Analyzer

this may be done satisfactorily with some analyzers, it should be remembered that the fundamental is often hundreds of times as large as the second or third harmonic. In making measurements of a distorted wave in the presence of the fundamental, the measuring instrument must be able to make accurate measurements of millivolts in the presence of volts. In the fundamental-suppression analyzer method, the input waveform is applied to a network that suppresses the fundamental component and passes the harmonic frequencies with negligible attenuation. If a thermocouple or a square-law electronic voltmeter serves as the indicating device, the rms value of all the harmonic components will be indicated.

11-279. A number of methods are used for removing the fundamental frequency. One method utilizes a high-pass filter, which is so designed that the harmonics lie in the pass band and the fundamental is severely attenuated. Another method is to employ a bridge circuit. If the bridge is adjusted to give a null at the fundamental frequency, a meter placed across the null points will indicate the rms value of the harmonics. Another alternative is a balanced-T circuit similar to the one shown in figure 11-35. In this network the circuit is balanced for the fundamental frequency, and is out of balance for the harmonics. The resonant circuit consists of inductor L and capacitors C1 and C2. The capacitors are tuned to the fundamental frequency, and resistor R is adjusted until the fundamental frequency is suppressed. The figure of merit (Q) of the resonant circuit must be at least 3 to 5, or excessive attenuation of the harmonics will result.

11-280. Figure 11-36 shows, in block diagram form, a basic fundamental-suppression distortion meter. The first reading is obtained by placing the switch in position 1 and adjusting the network for minimum output. At this setting, the fundamental frequency is suppressed. Next, the switch is placed in position 2 and the attenuator is adjusted until the output indication is the same as before. The attenuator reading in db then gives the amount of rms distortion below the amplitude of the fundamental.

11-281. Distortion meters which operate on the principle of fundamental suppression are simpler and less expensive than instruments of the heterodyne type, but cannot be used when the amplitudes of the individual distortion components must be known. However, a network that suppresses the fundamental can be used to advantage in combination with other types of analyzers, such as tuned-circuit and heterodyne meters. Two benefits

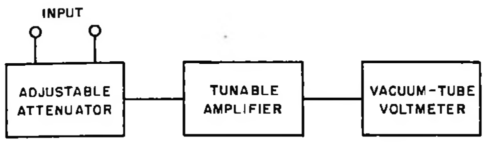


Figure 11-37. Tunable Amplifier Harmonic Analyzer

are gained; one is a reduction in the amount of harmonic distortion generated in the analyzer itself, and the other is less stringent selectivity requirements. This is important when there is frequency drift of the waveform being analyzed. Finally, wave analysis can be accomplished quite accurately by means of an oscilloscope in conjunction with graphical methods, if the fundamental is removed first.

11-282. TUNABLE AMPLIFIER ANALYZER. Another form of tunable selective-circuit analyzer is the tunable-amplifier analyzer. Of the analyzers described, the tunable amplifier is perhaps the most difficult to use, since considerable skill is required to operate the device and obtain consistent results. However, tunable-amplifier analyzers have frequency ranges that are not inherent in the other types of analyzers. In this analyzer audio-frequency coils are eliminated by the use of a tunable amplifier. The essential features of the circuit are shown in the block diagram of figure 11-37.

11-283. The complex wave is coupled to a selective amplifier after it is passed through an adjustable attenuator; the output is indicated by a vacuum-tube voltmeter. The instrument is calibrated by introducing voltages of known amplitude and frequency. A desirable characteristic of this type of amplifier is its extended frequency range. The bandwidth of the amplifier increases with frequency; therefore, the high-frequency selectivity is not good.

11-284. INTERMODULATION DISTORTION MEASUREMENT. The quality of an audio reproduction may prove unsatisfactory, even when the harmonic distortion of a single frequency is low. This condition is frequently attributable to a beating effect between the various frequency components of the waveform, so that appreciable sum- and difference-frequency components are produced. Since the new frequencies are not harmonically related to the original frequencies, their presence is especially objectionable.

11-285. Intermodulation can be determined by simultaneously applying, to the amplifier under test, two sine waves of different frequencies (for example, 400 cps and 1500 cps), and then measuring the amplitude of the difference-frequency component (1100 cps) appearing in the output. By means of a suitable filter, all signals but the 1100-cps component must be suppressed. The remaining component may then be rectified and applied to a dc meter. Intermodulation percentages under 8 or 10 percent are considered good.

11-286. NEUTRALIZATION PROCEDURES.

11-287. GENERAL. Neutralization is the process of balancing out the voltage that feeds through an rf amplifier by means of the interelectrode capacity of the electron tube. Since the usual amplifier operates with its input and output circuits tuned to the same frequency, the amplifier will break into oscillation as a tuned-plate, tuned-grid oscillator if the rf feedback between the grid and plate is not brought to the necessary minimum. Triodes have the greatest interelectrode capacitance, and, therefore, require maximum stabilization. Although the plate-grid capacitance of the screen-grid tubes usually used is reduced to only a fraction of a picofarad by the screen grid, the power sensitivity of these tubes is so great that only a small amount of feedback is necessary to induce oscillation.

Therefore, to ensure stabilization of the amplifier, it is usually necessary to either load the grid circuit or to use a neutralizing circuit external to the tube in order to balance the voltage feedback through the grid-plate capacitance by another voltage of opposite phase. For proper neutralization, the neutralizing voltage must be opposite in phase and equal in amplitude with respect to the feedback voltage between the grid and plate. When the plate circuit is divided so that the neutralization voltage is developed across part of it, the amplifier is said to have plate neutralization; when the neutralization is developed in the grid circuit, the amplifier is said to have grid neutralization. There are many variations in the methods employed to provide neutralization.

11-288. As the operating frequency of an rf amplifier is raised, the output energy coupled back to the input circuit by the grid-plate interelectrode capacitance eventually becomes large enough to cause sustained oscillations of the tuned-plate tuned-grid amplifier. This kind of oscillation differs from parasitic oscillations in that the oscillations occur at the resonant frequency of the tank circuits. In the neutralization process, the regenerative feedback is either cancelled out (neutralized) by a voltage of opposite polarity, or prevented from reaching the input circuit by a high series impedance in the feedback path. The first method utilizes various bridge circuits to provide the neutralizing signal, while the second method makes use of a suitable inductor shunted across the feedback capacitance, to develop a high series impedance at the oscillating frequency.

11-289. Generally speaking, when triode rf stages are employed, regeneration occurs above 100 kc, but when screen-grid tubes such as tetrodes, pentodes, and beam tetrodes are used, regeneration is seldom troublesome below 30 mc. The superiority of screen-grid tubes in this regard is attributable to the shielding action of the screen and suppressor grids, which are placed at rf ground potential. Because of this design the grid-plate interelectrode capacitance is greatly reduced in magnitude. The low value of grid-plate capacitance precludes oscillation at all but extremely high frequencies. No neutralization is required in frequency-multiplier circuits, because the grid and plate circuits are tuned to different frequencies. Another method of reducing feedback is the use of a grounded-grid amplifier, which is particularly effective in vhf and uhf receivers. With the grid of the tube connected to ground so that the input signal is applied to the cathode, any energy coupled from the plate to the grid by the grid-plate capacitance is returned to ground directly. The amplifier is characterized by low input impedance, which loads the exciting circuit, and by high output impedance.

11-290. NEUTRALIZATION CIRCUITS. Bridge methods are shown in parts (A) through (D) of figure 11-38. In part (A), a tapped inductor is used in the plate tank circuit of a single-ended triode rf amplifier. Since the rf voltages at the ends of the tank are 180 degrees out of phase, proper adjustment of the neutralizing capacitor will result in a null across the grid circuit at the frequency of oscillation. This method is successful only at frequencies below 7 mc.

11-291. The split-stator method shown in part (B) of figure 11-38 is used more widely. This arrangement makes the electrical balance virtually independent of the mutual coupling within the coil, and also of the point where the coil is tapped. If adjustment is made at a relatively high frequency, such as 15 mc, the stage can usually be operated at lower frequencies without requiring further adjustment. Sometimes an additional balancing capacitor is inserted between ground and the junction of the split-stator and balancing capacitors. Its purpose is merely to main-

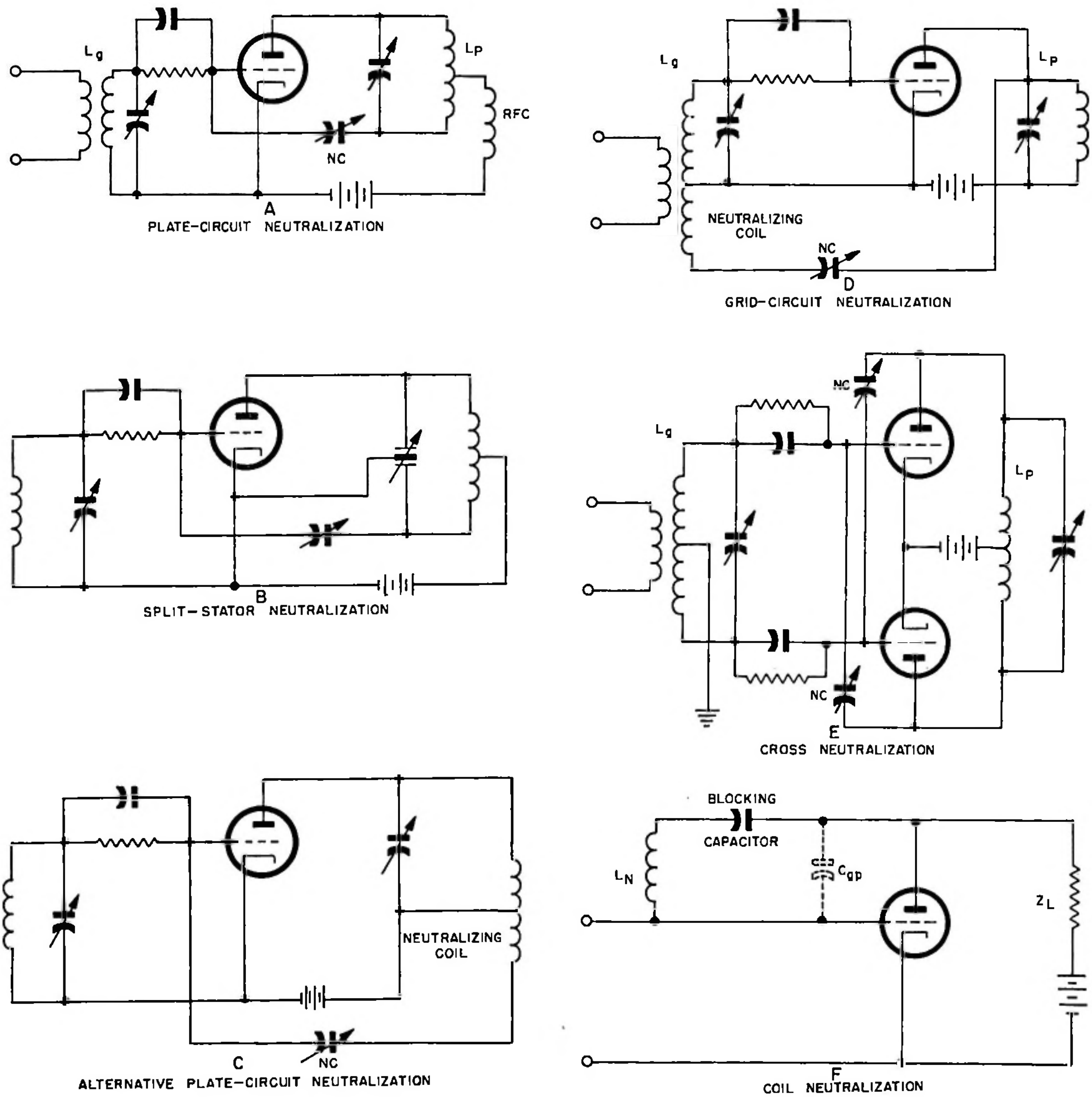


Figure 11-38. Neutralization Circuits

tain equal capacitance to ground on each side of the balanced plate tank circuit.

11-292. In part (C) of figure 11-38 is shown another plate-neutralizing circuit similar to the one in part (A), but lacking its limitations. A separate neutralizing coil is inductively coupled to the plate tank inductor. Part (D) shows a similar arrangement in which a coil is inductively coupled to the grid tank. Note that there is no flow of tank current through the neutralizing coil in either case. The size of the neutralizing capacitor, NC, depends on the coefficient of coupling between the tank and the neutralizing coils, and upon the relative values of the inductances. By proper proportioning of the neutralizing coil used on each band of operation, it is possible for one value of neutralizing capacitor to be used on all bands. In another form of grid-circuit neutralization, the tank coil is center-tapped in the same way as the plate tank coil shown in part (A).

11-293. In push-pull stages the symmetry of the circuit makes neutralization especially simple, as shown in part (E) of figure 11-38. Push-pull neutralization has the advantage of making balance possible more readily than in single-ended amplifiers; this is an especially desirable property at very high frequencies. In addition, neutralization is usually preserved throughout the various bands of operation.

11-294. The coil-neutralizing method is shown in part (F) of figure 11-38. Inductor L_N is resonated with the grid-plate capacitance of the tube at the frequency of oscillation. This causes the grid-plate impedance at that frequency to be high enough that regeneration due to interelectrode capacitance is unable to occur. The principal advantage of this method is that single-ended tank circuits can be used with a single-ended amplifier. However, there is also the disadvantage of restricting the neutralization to a limited range of frequencies. This limitation can be offset somewhat by shunting a trimmer capacitor across the neutralizing coil. The stage can then be neutralized at any desired frequency within a band of operation, provided that the trimmer is tuned appropriately. Coil neutralization is used extensively in broadcast transmitters.

11-295. NEUTRALIZING TECHNIQUES. The technique used in neutralizing an rf amplifier is essentially the same regardless of the type of tube or circuit employed. First, the plate voltage (and screen voltage, if applicable) is removed from the amplifier; the filament voltage of the tube is then turned on, and the rf grid excitation from the driver or preceding stage is applied. A fairly sensitive indicator should be loosely coupled to the plate tank coil.

11-296. USING FIELD STRENGTH METER. A field strength meter or a grid-dip meter makes a sensitive neutralizing indicator. In this application of a grid-dip meter, no plate potential is applied and the grid-cathode portion of the tube acts as a diode. Consistent with the behavior of a conventional field strength meter, the meter reading will increase perceptibly when the grid-dip meter is closely coupled and tuned to a source of rf energy. For this measurement the plug-in coil of the grid-dip meter should be coupled to the output tank coil at the low-potential or ground point. Care should be taken to make sure that the coupling is loose enough at all times to prevent burning out of the meter or rectifier of the field strength meter.

11-297. The plate tank circuit of the amplifier should be tuned to resonance, which is indicated by a maximum reading on the rf indicator. The neutralizing capacitor is then adjusted until the rf indicator shows a minimum reading. This operation may detune the plate tank of the driver stage slightly; if so, the plate tank of the driver stage should be carefully retuned to resonance, and then the plate tank of the amplifier again tuned to resonance. The rf indicator will usually show another maximum reading, but one of considerably less magnitude than the original reading. The neutralizing capacitor is again adjusted for minimum (or zero) rf indication. After this procedure has been repeated several times, a setting of the neutralizing capacitor should have been found which shows no rf voltage in the plate tank circuit of the amplifier. As the point of correct neutralization is more closely approached, the coupling of the rf indicator will usually have to be tightened, because there will be less rf voltage available to operate the indicator. After each adjustment of the neutralizing capacitor, the driver tank and the amplifier tank should be retuned to resonance. When the rf indicator shows zero rf voltage in the amplifier tank, the stage is properly neutralized. If a push-pull stage is to be neutralized, both neutralizing capacitors should be adjusted simultaneously. They will not, however, always have exactly the same setting when neutralization is reached, because of slight differences in stray capacitances and because the tuned tank circuit may not be electrically symmetrical. During the neutralization adjustment, you must make certain that the shielding, including metal covers, for the amplifier stage is in its normal location. If this precaution is not observed, the neutralization adjustment will not be correct when the metal covers are replaced or the other shielding returned to its normal location.

11-298. USING OSCILLOSCOPE AND OTHER RF INDICATORS. The above method of neutralization may be used with other suitable indicators, such as a neon bulb, flashlight bulb, thermogalvanometer, or standing-wave meter. An oscilloscope can also be used, provided that the rf output is applied directly to the deflection plates of the oscilloscope, and that the frequency involved is not high enough to be shorted out by the capacitance of the leads. An oscilloscope is sometimes the most convenient voltage indicator, since it may already be connected for observing the rf envelope, or for other tests.

11-299. USING GRID METER. A very sensitive neutralizing indicator is a dc milliammeter connected in the grid-return circuit of the amplifier which is being neutralized so as to measure rectified grid current. This meter is a part of many transmitters. With the plate-voltage lead disconnected as before, the driver tank circuit is tuned until the dc meter in the amplifier grid circuit shows a maximum reading. If the amplifier is not properly neutralized initially, tuning its plate tank circuit through resonance will cause the dc grid current to vary. The neutralizing capacitor should be adjusted slowly while the plate tank circuit of the amplifier is tuned gradually back and forth through resonance. As the point of correct neutralization is approached, the flicking of the needle of the dc grid meter will gradually decrease in amplitude. If the amplifier is perfectly neutralized, tuning the plate circuit through resonance will not change the meter reading, even slightly. During these adjustments, the driver plate circuit should occasionally be retuned to resonance, as indicated by a dip in its dc plate current or by a maximum in the dc grid current of the amplifier.

11-300. Because the rectified dc grid current is a measure of the rf excitation applied to the amplifier, the use of a dc meter is usually advisable. The grid meter is not only useful for neutralizing adjustments; it also provides a continuous check on the operation of the amplifier and the driver stage as well.

11-301. In some cases it may be found that, while a setting of the neutralizing capacitor can be made which will give a definite minimum rf indication, no adjustment will entirely eliminate rf voltage from the tank circuit. This effect is sometimes due to stray coupling between the amplifier and driver plate tanks or to stray capacitance between various parts of the amplifier, which tend to unbalance the neutralizing circuit. Adequate shielding between grid and plate circuits and between stages will often eliminate neutralizing difficulties. Shielding may actually cause trouble, however, if it is placed too close to the tuned circuits or to the neutralizing capacitor. It is important that the ground lead from the rotor of a split-stator capacitor be made direct (and as short as possible) to the filament circuit.

11-302. DATA LINK TESTS.

11-303. Radio transmitters that are used for the transmission of data are generally part of terminal or repeating relay stations that are used with multiplexing equipments; therefore, they generally operate in the very-high-frequency to super-high-frequency region, and are modulated over very wide bandwidths. Measurements of frequency, modulation percentage, standing-wave tests, etc, necessary for other type transmitters are also required for this type of equipment. Since the signal-to-noise ratio of data-link systems must be high to insure reliable operation, both the transmitting and receiving equipment should operate at optimum performance. Most of the newer multiplexing relay equipments operate in the microwave band, and incorporate various methods of time-division or frequency-division modulation; consequently, much of the test equipment is designed especially for the particular radio set.

11-304. FACSIMILE COMMUNICATIONS TESTING.

11-305. Facsimile is a process of transmitting a picture, map, or other graphic material (page copy) from one terminal to another. At the transmitting equipment, the picture is mounted on a circular drum or cylinder. The picture is then scanned by an electro-mechanical device. Picture variations along the scanning lines are converted to light variations, and then into electrical impulses which are used to modulate a transmitter. At the receiving end, the demodulated electrical impulses are fed to a facsimile recorder which reproduces the picture, using an electro-mechanical scanning device similar to that used at the transmitting end except that it reverses the process. Received copy can be recorded either directly on chemically coated paper, using electrical impulses, or photographically in positive or negative form by converting the electrical impulses back into light variations.

11-306. Facsimile is used with either wire or radio communication circuits. When used with wire line services, the facsimile transceiver can be connected directly to the line or through a coupling coil; a conventional telephone headset may be coupled to the line for signal monitoring purposes. When used with radio communication circuits the facsimile transceiver can be connected in several ways, by use of auxiliary equipment, to produce different types of radio signals.

11-307. FACSIMILE SIGNALS.

11-308. SUBCARRIER AMPLITUDE MODULATION (SCAM). When using the subcarrier amplitude modulation method of signal transmission, the facsimile equipment output is connected directly to the microphone circuit of an amplitude modulation transmitter. This type of circuit is used to duplicate the range of half-tones detected in the original copy, but is very subject to copy degeneration from fading and environmental noise.

11-309. SUBCARRIER FREQUENCY MODULATION (SCFM). The subcarrier frequency modulation method of transmission is superior to the scam method in that noise is greatly reduced. For this type of transmission, the facsimile transceiver employs a converter unit for the purpose of frequency-modulating the amplitude-modulated 1800-cps signal from the facsimile equipment. The converter output is then fed into a conventional radio-telephone transmitter; the transmitter emission is a narrow-band subcarrier frequency modulation signal. The facsimile intelligence is then transmitted as a constant amplitude signal, the frequency of which is varied by the facsimile copy. An a-m receiver, converter, and the same type of facsimile equipment used for transmission are used for reception. The converter changes the subcarrier fm signals back to a-m signals, which are required for operation of the facsimile transceiver.

11-310. FREQUENCY-SHIFT MODULATION. A large volume of Air Force facsimile communications, such as weather maps, are transmitted in black and white tones only. Frequency-shift modulation provides an excellent method of transmitting black and white line drawings or weather maps. When the facsimile transceiver is used with both a converter and a transmitter-exciter unit, the facsimile intelligence can be sent by fm transmission. When receiving, a stable a-m communications receiver with a crystal-controlled, high-frequency oscillator and a converter are used to convert the frequency-shift signals to the necessary a-m signals. Receivers that do not have stable beat-frequency oscillators can be used by employing a frequency meter as an oscillator.

11-311. OPERATING CONSIDERATIONS.

11-312. Usually, the facsimile circuit is just one of many circuits incorporated in a communications network. In this case, the facsimile circuit is fed to one or more channels of multiplexing equipments. Since many facsimile equipments are remotely located in weather centers, the facsimile signals originating in the weather center are delivered to the communications relay center, where they are normally transmitted to a transmitter site and then retransmitted to a distant station, where the procedure is essentially repeated during reception. Facsimile checks are fundamentally equipment checks; however, at the various links, a facsimile converter and recorder are provided to facilitate operating checks of radio facsimile transmissions. The equipment may be used as a monitor to record the facsimile signal received from the communications center, or, when used in conjunction with a receiver at the transmitter site, to detect and record the facsimile signal radiated as an rf signal from the transmitter site. The prime purpose of the monitor equipment is to determine the source of trouble in the event that a distorted or otherwise unsatisfactory signal is reported.

11-313. Most defects resulting from poor installation, faulty adjustments of the equipment, and unfavorable radio propagation can be determined by analyzing the transmitted or received copy. The facsimile signal is generally an a-m, 1800-cps carrier, and the circuit used is capable of passing both the upper and lower sideband frequencies, 900 to 2700 cps, as flat as possible for satisfactory reproduction. If the impedances throughout the entire circuit are not properly matched, reflections will result and will show on the copy. However, the matching of impedances is only important as far as visible effects on the picture are concerned in telephone line transmission. When the circuit distance is long, the received signal is reflected back to the sending station, and is reflected again over the line to the receiving station. This causes two or more images to appear on the received copy. The double image could also occur in radio transmission because of multiple paths from the transmitter to the receiver with different time delays. Echo suppressors are generally employed to overcome this type of defect.

11-314. It is important that the signal level at the receiving station be constant, especially when receiving photographic copy, since instantaneous changes as small as 0.25 db in signal level will cause a noticeable change in the shading of the recorded picture. However, an increase or decrease of 1 to 2 db is permissible if the change occurs gradually during the transmission of the entire picture.

Delays in the line cause a distorted picture, but may be corrected with delay-correcting networks which equalize the frequency transmission rates. A high signal-to-noise ratio is desirable to eliminate dark streaks or spots on the recording which are a result of noises. Other defects, like jag or jitter, irregular skew, etc, are generally caused by improper operation of the facsimile equipment.

11-315. Transmitting and receiving equipments at the ends of the circuit should have the same index of cooperation, the same speed, and a recording system which is responsive to the modulated signal of the transmitter. The index of cooperation is equal to the scanning drum diameter times the pitch (lines per inch scanned). Satisfactory operation may be obtained, even though the indexes of cooperation are not exactly the same. Differences of as much as 5 or 10 percent are generally not noticeable in photographic or message copy. The indexes of cooperation may differ by as much as 50 percent for weather map work or other line illustrations without affecting the usability of the received copy. The maximum speed tolerance between the transmitting and receiving scanning drums of many facsimile transceivers is 0.00033 percent. Table 11-3 below lists the standards applicable in present-day military and commercial wire and radio-photo facsimile circuits.

11-316. Adjustment of the facsimile transmitting transceiver will depend on the type of transmission being made and the type of receiving process in use. Coordination is necessary between the transmitting and receiving operators. Transmission can be either positive or negative. In positive transmission, the black portions of the copy are transmitted as high-level signals which print at the receiving end as black. In negative transmission, the white portions of the copy are transmitted as high-level signals which print as black, producing a negative. The relative signal levels are indicated by the db meter on the facsimile transceiver. The kind of copy being sent and the use to be made of the copy at the receiving end determine whether the transmission should be negative or positive.

11-317. TAPE FACSIMILE.

11-318. Tape facsimile machines scan copy in the form of tape. The copy height is 1/4 inch and the total width of the tape is 3/4 inch. Text is hand-printed with pencil by the operator on a writing stand provided with guides and a means for holding a roll of blank tape. The copy is threaded through the sending machine, which scans it along parallel lines running crosswise of the tape. The sending and receiving machines are mounted in a single housing and the receiver can be used to monitor outgoing signals, if desired. The receiver uses direct electromechanical recording, in which the tape is pressed against an inked printing element to make the marks. The machine handles the tape at the rate of 50 inches per minute. About four words can be hand-printed every 5 inches; however, it is impossible to keep up the handprinting on the tape at more than about 15 words per minute. The scanning lines run 72 to the inch, and are 3/8-inch high. The amplitude-modulated signals from the photo-cell pickup are converted to signals which are one of two frequencies, 1650 cps for black and 1150 cps for white. Either one or the other of these frequencies is transmitted, depending on whether the copy being scanned is black or white. At the receiving station, reception of these two frequencies operates an appropriate mechanical system for producing black and white. Synchronization between the sending and receiving ends is effected by a vibrator of approximately constant frequency at each end. The limitation on the speed difference between the two ends is made considerably more lenient than in the case of the page facsimile by printing two

Table 11-3. Standards for Military and Commercial Facsimile Circuits

STANDARD	INDEX	SPEED (RPM)
Facsimile Transceiver TT-1( )/TXC-1	576	60
Facsimile Transceiver FX-1-( )	264	90
Acme	290	100
Associated Press	380	100
CCIT and CCIR (Consultive Committee on International Telegraph and Consultive Committee on International Radio)	264	90
CCIT and CCIR	352	60
International News Service	264	90
Western Union	254	180

lines of received copy on the tape. Imperfect synchronization causes one line to approach an edge of the tape, and eventually to run off and come back from the other edge. With two lines of record copy, however, one line remains continuous and completely legible.

11-319. Tape facsimile equipment is generally applicable only to situations where speed is not necessary, and where audible reception (such as with telephone or manual telegraph) is under an unusual handicap, as in areas where ambient noise (room noise) is high. Other situations may occur, such as where slightly more privacy is desirable than can be obtained with ordinary teletypewriter circuits, and where weight is an important factor. It is thought that the required signal-to-noise ratio is intermediate between that for teletypewriter and that for manual, so that somewhat more transmission range than for teletypewriter can be obtained at relatively slow message speeds, without the necessity for skilled operators. Tape facsimile is rarely used in a communications network. It is employed mostly for a particular tactical situation.

11-320. TELETYPEWRITER COMMUNICATIONS TESTING.

11-321. Signals can be produced by the operation of a teletypewriter keyboard similar to that of a typewriter, and the corresponding characters types in page form or on tape by printer mechanisms at both the sending and receiving stations. Signals may also be sent from a punched tape, prepared by another device called a perforator, which has a keyboard which is also similar to that of the typewriter. Signals received from a transmitting station do not necessarily have to be printed in page form; they may also be printed on a perforated paper tape by means of a typing reperforator. Some typing reperforators are furnished with a keyboard so that it may be used for the transmission of messages, especially under tactical conditions. The perforated tape is shown in figure 11-39. The small holes are perforations used to feed the tape through a semi-automatic device known as a transmitter-distributor.

11-322. TELETYPEWRITER SIGNALS.

11-323. GENERAL. Signals originated within the teletypewriter are transmitted to associated circuits in the form of direct current pulses called marking and spacing impulses. A marking impulse is produced by a closed teletypewriter circuit, as shown in figure 11-40, where a direct current is flowing in the series circuit. A spacing impulse occurs when the teletypewriter circuit is in an open-circuit condition, where there is no flow of direct current.

11-324. CODE ARRANGEMENT. All of the functions, letters and figures are composed of five intelligence impulses, each of equal time length, plus an additional two impulses used for synchronizing the receiving teletypewriter with the transmitting teletypewriter. The sequence of intelligence impulses, whether mark or space impulses, are mechanically selected when the operator depresses a teletypewriter key. Refer to figure 11-39, and compare the "perforated tape" with the "line signals," noting that the holes in the perforated tape and the heavy black lines correspond to the marking impulse.

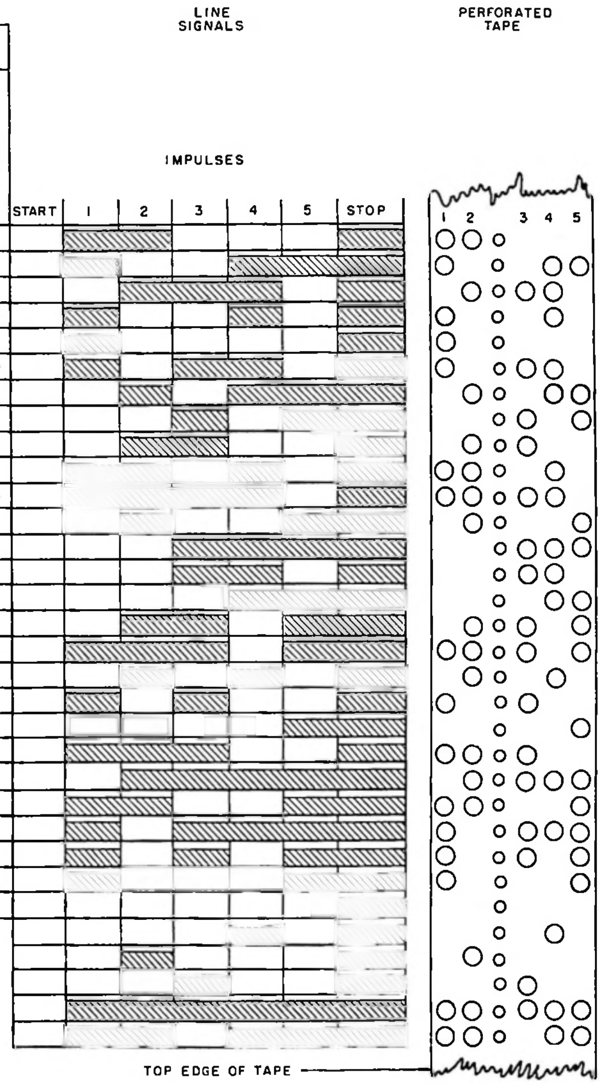
11-325. The total possible number of mark and space combinations in the 5-unit permutation code is 32. It is necessary, since a total of 58 characters would be required to include functions, numerals, punctuation, etc., that each mark and space combination be represented by two teletypewriter characters. Each letter, therefore, is paired-off with a numeral, special symbol, or punctuation mark, as shown in figure 11-39. When the letters of the alphabet are desired, the operator depresses the "letters" keybutton and the printing mechanism will record letters until the figures keybutton is operated. When the "figures" keybutton is depressed, the printer will record numerals and punctuation, or special weather symbols, as the case may be, until the "letters" keybutton is again depressed.

11-326. A further inspection of figure 11-39 reveals that there are two additional impulses transmitted with each character. These are the two synchronizing impluses previously mentioned, and are called start and stop impulses. Notice that the start impulse is always spacing, to permit the relays in the receiving printing mechanism to start translating the coded impulses at the same time that the transmitting mechanism sends the selected character. In other words, both transmitting and receiving teletypewriters act upon the same impulse at the same time. The stop impulse, on the other hand, is always marking, to insure that all receiving mechanisms stop their mechanical operations at the same time the transmitting teletypewriter has finished sending its coded signal.

11-326A. Air Force communications has used teletypewriter message transmission equipments for a number of years and is expected to continue using it for some time in the future. The five-unit teletypewriter coded character set shown in figure 11-39 has been used throughout these equipments as a standard. Because its future use is problematical, this set is actually included in this publication as an interim standard, and nothing more. This approach stems from the great advances made by the automatic electronic information data processing industry. Numerous codes and media practices have been developed recently for input and output use with an ever-increasing variety of processing equipments.

PLATEN POSITION

LTRS: ALL MACHINES	FIGS: COMMUNICATION MACHINES	FIGS: WEATHER MACHINES
A	–	↑
B	?	⊕
C	:	○
D	$	↗
E	3	3
F	!	→
G	&	↘
① H	STOP	↓
I	8	8
J	'	↙
K	(	←
L	)	↖
M	.	.
N	,	⦶
O	9	9
P	0	0
Q	1	1
R	4	4
S	BELL	BELL
T	5	5
U	7	7
V	;	⌽
W	2	2
X	/	/
Y	6	6
Z	"	+
BLANK		–
② CARRIAGE RETURN		
③ LINE FEED		
SPACE		
LETTERS SHIFT		
FIGURES SHIFT		



SIGNAL LENGTHS IN MILLISECONDS, STANDARD SPEED 60 WORDS PER MINUTE

22 — 22 — 22 — 22 — 22 — 22 — 31

SPACING IMPULSES

MARKING IMPULSES

① MOTOR STOP OPTIONAL ON WEATHER MACHINES

② < IS PRINTED ON TYPING REPERFORATORS

③ ≡ IS PRINTED ON TYPING REPERFORATORS

Figure 11-39. Five-Unit, Start-Stop Teletypewriter Code

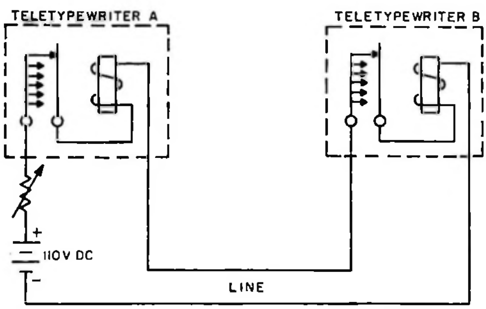


Figure 11-40. Basic Two-Station Teletypewriter Circuit

11-326B. The five-unit teletypewriter coded character set has been little used for either input or output purposes in automatic information data processing for a number of reasons. One reason is the requirement for a case shift to represent the characters in the set; another reason is the lack of an order arrangement in code assignments from the standpoint of information processed by machine methods. Through a cooperative effort made by representatives of the data processing industry, the communications industry, and the Federal Government, a coded character set has been developed and approved during the past three years as an American Standard Code for Information Interchange X3.4-1963 (ASCII) which is referred to in figure 11-40A. The ASCII is intended to serve as an universal code for input/output and interchange purposes in automatic data processing, data transmission, and data capture where coded characters are used. The general use of a standard code will minimize requirements for code conversion and related types of intermediate processing operations during the exchange of information in machine code form.

11-326C. AMERICAN STANDARD CODE. This coded character set is to be used for the general interchange of information among information processing equipments, communication equipments, and associated components as is shown in figure 11-40A. Standard 7-bit set code positional order and notation are shown as follows with b_7 the high-order, and b_1 the low-order bit position. For example, the code for the alpha character R is as follows:

b_7	b_6	b_5	b_4	b_3	b_2	b_1	general bit positions
1	0	1	0	0	1	0	specific code

The legend for the abbreviations shown in the figure is as follows:

NULL	Null/Idle	CR	Carriage return
SOM	Start of message	SO	Shift out
EOA	End of address	SI	Shift in
DC_1-DC_3	Device control	DC_0	Device control reserved for data
DC_4 (Stop)	Device control (stop)		link escape
ERR	Error	SYNC	Synchronous idle

EOM	End of message	LEM	Logical end of media
EOT	End of transmission	S_0-S_7	Separator (information)
WRU	"Who are you ... ?"	t_o	Word separator (space, normally nonprinting)
RU	"Are you ... ?"		
BELL	Audible signal		Less than
FE_o	Format effector		Greater than
HT	Horizontal tabulation		Up arrow (Exponentiation)
SK	Skip (punched card)		Left arrow (Implies/Replaced by)
LF	Line feed		Reverse slant
Vtab	Vertical tabulation	ACK	Acknowledge
FF	Form feed		Unassigned control
DEL	Delete/Idle	ESC	Escape

11-326D. The American standard does not define the means by which the coded set is to be recorded in any physical medium. Moreover, the standard code does not include any redundancy nor does it define techniques for error control. Further, it does not specify a standard collating sequence. For military operations the standard code for teletypewriter equipment shall be the same as the code for digital data communication equipment. The adoption of this code shall not inhibit the use of random binary bit streams. The military code is composed of eight units, that is, the seven bits indicated in figure 11-40A and an eighth bit called a parity bit. For transmission, the eighth bit is used for odd parity check. This means that the bit is set so that the sum of the binary ones of the eight-bit code character is always an odd number, and so that an even sum would indicate an error. For punched tape, all parity bits are automatically reversed to provide even parity. The coded character is set to be transmitted with the low order first. Thus bit b_1 will appear on-line first and parity bit b_8 will appear last, hence the military code for the alpha character in the figure is as follows:

b_8	b_7	b_6	b_5	b_4	b_3	b_2	b_1
0	1	0	1	0	0	1	0

11-326E. Note that the unassigned codes are reserved for future standardization. Their use in information interchange prior to such standardization is a deviation from the standard. Deviations from the standard may create serious difficulties in information interchange and they should be used only with the full cognizance of the agencies involved. The substitution of symbols in the basic code for special purposes shall be permissible provided that in any such coding arrangement, the characters retained shall retain the defined code designations.

b_4	b_3	b_2	b_1	b_7 0 b_6 0 b_5 0	b_7 0 b_6 0 b_5 1	b_7 0 b_6 1 b_5 0	b_7 0 b_6 1 b_5 1	b_7 1 b_6 0 b_5 0	b_7 1 b_6 0 b_5 1	b_7 1 b_6 1 b_5 0	b_7 1 b_6 1 b_5 1
0	0	0	0	NULL	DC_0		0	@	P		
0	0	0	1	SOM	DC_1	!	1	A	Q		
0	0	1	0	EOA	DC_2	"	2	B	R		
0	0	1	1	EOM	DC_3	#	3	C	S		
0	1	0	0	EOT	DC_4 STOP	$	4	D	T		UNASSIGNED
0	1	0	1	WRU	ERR	%	5	E	U	UNASSIGNED	
0	1	1	0	RU	SYNC	&	6	F	V		
0	1	1	1	BELL	LEM	'	7	G	W		
1	0	0	0	FE_0	S_0	(	8	H	X		
1	0	0	1	HT_{SK}	S_1	)	9	I	Y		
1	0	1	0	LF	S_2	*	:	J	Z		
1	0	1	1	V_{TAB}	S_3	+	;	K	[		
1	1	0	0	FF	S_4	,	<	L	\		ACK
1	1	0	1	CR	S_5	-	=	M	]		1
1	1	1	0	SO	S_6	.	>	N	↑		ESC
1	1	1	1	SI	S_7	/	?	O	←		DEL

Figure 11-40A. American Standard Code for Information Interchange

11-326F. The individual standards outlined in the previous paragraphs have not come into general usage at the present time, and due to the widespread use of existing teletypewriter equipments, the existing interim standard will be operational in its entirety for an indeterminate period. This interim standard is the 5-unit code, start-stop, International No. 2 Alphabet (American Variation) as was shown in figure 11-39.

11-326G. INTERIM STANDARD CODE. In order to derive a signalling code, two independent conditions must be satisfied. The first condition is the derivation of a systematic arrangement of elements to form a definite, recognizable, and repeatable pattern. Where the code is to be processed by mechanical methods, the arrangement of elements is usually positional in nature. If the code is to be processed by electrical means, a variety of element arrangements are possible, such as those involving position, size, density, and so forth. The second condition is the arbitrary assignment of intelligence information to each code element devised. Assignment of intelligence information to a code element need not be unique; that is, a single code element can be assigned a dual intelligence role provided that some means is also assigned to eliminate ambiguity for purposes of translation. The recognition of a code pattern in association with the intelligence that it represents is termed translation.

11-326H. INFORMATION BITS. The Baudot code can be mathematically treated using binary notation since the keyed electrical circuit is either in a conducting status or in a non-conducting status depending upon whether the keying device is open or closed. In order to maintain the same time interval for each character, the character interval has to be composed of a specified number of unit time elements sufficient to encode the 26 letters of the alphabet, the 10 basic numerals, and the 8 basic punctuation marks used in the English language. These considerations require the use of 44 characters which yield the binary relationship 2^6 44, thus indicating that each character could be divided into 6 equal time units. On the other hand, if typewriter technology were adopted, the 10 numerals and the 8 punctuation marks could be doubled up with the alphabet by using a special character, thereby reducing the binary relationship to 2^5 26 18. Further reductions are insufficient since the relationship 2^4 26 and would not permit the encoding of the full alphabet. Therefore, each character time interval has been divided into 5 equal units of time, with the code being called the 5-unit code.

11-326I. In general, n different integers can be arranged in n different ways. Clearly, since n = 5, there are 120 different combinations. The most efficient combination arrangement is by a natural permutation, that is, there are no inversions included. Such a permutation would follow the form: 1; 2; 1,2; 3; 1,3; 2,3; 1,2,3; .... 1,2,3,4,5. Table 11-3A shows the permutation code in its natural progression and shows the recorded teletypewriter character obtained for that particular combination. Observation of this table does not reveal any definite planning in assigning the characters to the permutation code. Studies involving the frequency of English letter usage had already been conducted showing that certain letters were used more frequently than others. Table 11-3B shows the first 18 English alphabet letters in the order of their usage versus the code combination; selected for each of the letters. Notice that the code selection has not been as random as might be supposed; the pattern in table 11-3C shows that the letters used most were assigned the fewest number of combinations.

Table 11-3A. Signalling Code Arrangement According to National Permutation Code

COMBI-NATION	LETTER	COMBI-NATION	LETTER	COMBI-NATION	LETTER	COMBI-NATION	LETTER
1	E	1, 4	D	1, 5	Z	1, 4, 5	B
2		2, 4	R	2, 5	L	2, 4, 5	G
1, 2	A	1, 2, 4	J	1, 2, 5	W	1, 2, 4, 5	
3		3, 4	N	3, 5	H	3, 4, 5	M
1, 3	S	1, 3, 4	F	1, 3, 5	Y	1, 3, 4, 5	X
2, 3	E	2, 3, 4	C	2, 3, 5	P	2, 3, 4, 5	V
1, 2, 3	U	1, 2, 3, 4	K	1, 2, 3, 5	Q	1, 2, 3, 4, 5	
4		5	T	4, 5	O		

Table 11-3B. English Alphabet Letter Usage Versus Signalling Code Combination

LETTER	CODE	LETTER	CODE	LETTER	CODE
E	1	S	1, 3	C	2, 3, 4
T	5	H	3, 5	M	3, 4, 5
A	1, 2	R	2, 4	F	1, 3, 4
O	4, 5	D	1, 4	W	1, 2, 5
I	2, 3	L	2, 5	Y	1, 3, 5
N	3, 4	U	1, 2, 3	P	2, 3, 5

Table 11-3C. Signalling Code Arranged According to the English Alphabet

LETTER	CODE*	LETTER	CODE*	LETTER	CODE*	LETTER	CODE
A	1, 2	H	3, 5	O	4, 5	V	2, 3, 4, 5
B	1, 4, 5	I	2, 3	P	2, 3, 5	W	1, 2, 5
C	2, 3, 4	J	1, 2, 4	Q	1, 2, 3, 5	X	1, 3, 4, 5
D	1, 4	K	1, 2, 3, 4	R	2, 4	Y	1, 3, 5
E	1	L	2, 5	S	1, 3	Z	1, 5
F	1, 3, 4	M	3, 4, 5	T	5		
G	2, 4, 5	N	3, 4	U	1, 2, 3		

*Marking Elements

11-326J. Once the coding has been determined, there still remains the problem of adapting the code to practical usage. Marking and spacing terminology, currently used in the telegraph field, was applied to the code so that a particular letter could be identified by stating which of the code elements represented the marking element; then by association, those code elements not specified were assumed to be spacing elements. The signalling code, which is now termed a 5-unit permutation code, is used in its elemental form for paper tape records and retransmission of teletypewriter messages, as is shown in figure 11-39. The punched holes in the paper tape produce marking signals in an associated electrical circuit, while the absence of holes in the tape produces spacing signals. The location of the holes in the paper tape corresponds to the permutation code shown in table 11-3C.

11-326K. SYNCHRONIZATION BITS. Although the perforations in the paper tape contain all of the information bits necessary to identify an English character, certain problems had to be resolved, such as the question of distinguishing the beginning and the end of a character. The movement of paper tape from one character to another required a certain amount of time, and this time had to be allotted before the character could be transmitted. A second very obvious problem was the electrical signal itself, since a particular code combination could leave the electrical circuit in either a conducting or a non-conducting status for any one character.

11-326L. The standard signalling sense was adopted whereby the electrical circuit would normally be conducting, and the marking condition was accordingly defined as the quiescent state of a telegraph circuit. When a teletypewriter signal is in progress, the signal corresponding to the marking element of the permutation code also corresponds with the hole in the perforated paper tape in addition to corresponding with the quiescent state of the telegraph circuit. Conversely, when the telegraph circuit is in a spacing condition, the electrical circuit is in a nonconducting status, corresponding with the unperforated portions of the perforated tape and the unspecified elements in the permutation code.

11-326M. Two additional time elements are combined with the information bits to form the final teletypewriter signalling code. One of the newly added time elements is prefixed to the information bits and is called the start interval. The start interval is always a spacing impulse; all information bits are interpreted with respect to the beginning of this impulse. The second of the newly added time elements suffixed to the information bits is designated as the stop interval. The stop interval is always a marking impulse and it serves to return the telegraph circuit to its quiescent condition prior to the reception of the next start interval.

11-326N. The start and stop bits of the signalling code can be termed synchronization bits. The start interval permits all selector mechanisms in the electrical teletypewriter circuit to initiate the mechanical selection process. The stop interval permits all selector mechanisms to be restored to their unselecting positions prior to the reception of the next start interval. Thus the transmitting and receiving equipments on a signal circuit both will come to a stop at some point between the completion of one code sequence and at the beginning of the next sequence. With the next start interval, these equipments will begin their sequences together, for all practical purposes, and if their motor speeds are correct, will complete their sequences without losing relative synchronism.

11-236O. CHARACTER REPRESENTATION. The standard teletypewriter signalling code consists of 5 intelligence elements and 2 synchronization elements arranged for transmission in the following sequence.

START	FIRST	SECOND	THIRD	FOURTH	FIFTH	STOP
SYNC ELEMENT		INTELLIGENCE ELEMENTS				SYNC ELEMENT

All elements except the stop element have the same unit time interval; the stop element time interval is fixed per given end instrument, but different length intervals are available. The time interval for all elements of the signalling code depends upon the desired modulation rate.

11-326P. The standard modulation rates for printing telegraph equipment below 150.0 bauds shall be 50.0 bauds and 75.0 bauds, respectively. For practical purposes, the term baud means the speed (modulation rate) of the shortest signal element (unit interval) is X units of time, that the signal is a square-wave in nature, and that the signal is probably a non-repetitive waveform in its information bearing condition. The modulation rate, in bauds, is the time length of the shortest mark or space element (unit interval) that is present in the signal, divided into one second. The standard unit interval of each intelligence element and of the start element is 20 milliseconds for a 50-baud modulation rate or 13.33 milliseconds for a 75-baud modulation rate.

11-326Q. The length of the stop element encountered depends upon the telegraph printing equipment. The length of the stop element is referred to as a percentage of the unit signal element, that is, 1.0 or 100 percent, 1.2 or 120 percent, 1.42 or 142 percent, 1.5 or 150 percent, 1.94 or 194 per-

START ELEMENT	FIRST ELEMENT	SECOND ELEMENT	THIRD ELEMENT	FOURTH ELEMENT	FIFTH ELEMENT	STOP ELEMENT
20 ms	20 ms	20 ms	20 ms	20 ms	20 ms	20 ms

1.0 STOP ELEMENT INTERVAL

START ELEMENT	FIRST ELEMENT	SECOND ELEMENT	THIRD ELEMENT	FOURTH ELEMENT	FIFTH ELEMENT	STOP ELEMENT
20 ms	20 ms	20 ms	20 ms	20 ms	20 ms	28.4 ms

1.42 STOP ELEMENT INTERVAL

Figure 11-40B. Baud Character Time Elements cent, 2.0 or 200 percent, etc. Two very common stop lengths are the 1.0 and 1.42 units. For a modulation rate of 50 bauds, this yields a stop length of 20 milliseconds for the 1.0 value, and 28.4 milliseconds for the 1.42 value, as is shown in figure 11-40B. For a modulation rate of 75 bauds, this yields a stop length of 13.33 milliseconds for the 1.0 value, and 19.928 milliseconds for the 1.42 value.

11-326R. The words per minute (WPM) rating is to be discouraged, since this is an operational terminology which has no place in technical terminology. The term "words per minute" informs operational personnel of the approximate number of words that may be expected for a given class of service. It does not indicate any information as to modulation rate, coding, or synchronization type. An example of this is that commercial equipment obtains 65 wpm from the same, unmodified receiving start-stop equipment over the same quality channel from which the military agencies obtain 61.3 wpm because the stop element used in commercial equipment is normally shorter than the stop element of military equipment. It is possible to have at least five different words-per-minute rates at 45.5 bauds. The lengths of the stop element normally encountered in this code is 1, 1.27, 1.42, 1.96, or 2. This means that while the stop element is either the same length or longer than the start or intelligence elements, the number of words per minute that can be transmitted is different for each stop element length in use. The one common meeting ground, however, is that the signals are all transmitted at a modulation rate of 45.5 bauds.

11-326S. Figure 11-40C shows a simple electrical circuit representing the basic teletypewriter line circuit. The transmitting portion of the circuit is represented by the switch, and the receiving portion of the circuit is represented by the electromagnetic coil and armature. When the switch is opened and closed by mechanical motion, an electrical signal is generated which is transmitted over the circuit. The electrical signal is received by the electromagnet coil, and the coil in turn operates the armature; the electrical signal is therefore converted back into mechanical motion. The teletypewriter signal is binary in nature; it is amenable to incorporation within digital data terminology, as is shown in table 11-3D.

11-326T. TRANSMITTED SIGNAL. The transmitter contacts shown in figure 11-40C are a standard representation of a teletypewriter sending mechanism. The open switch segments represent the switches that generate the five intelligence bits, while the closed switch segment represents the switch that generates both synchronization bits. Assuming the transmission of the alphabetical character Y at a 50-baud modulation rate

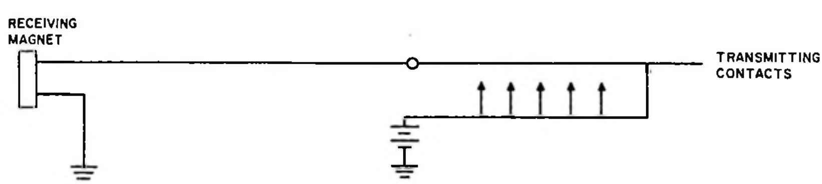


Figure 11-40C. Basic Teletypewriter Circuit

Table 11-3D. Digital Data Representations

FUNCTION	DIGIT 0	DIGIT 1
Signal Interval	Space	Mark
Perforations	No Perforation	Perforation
Single Current Modulation (Neutral Signals)	No Current	Positive Current
Double Current Modulation (Polar Signals)	Negative Current	Positive Current
Amplitude Modulation	Tone-off	Tone-on
Frequency Modulation	High-Frequency	Low-Frequency

using a 1.42 stop element, the operation of the generator contacts illustrated in figure 11-40D are as follows:

a. At time t_0, the S contact opens, circuit current drops to zero, and a mark-to-space transition marks the beginning of the start pulse.

b. At time t_1, 20 milliseconds later than t_0, the S contact remains open while contact 1 closes and the circuit current rises to 60 milliamperes during this space-to-mark transition to begin the first intelligence bit of the character Y.

c. At time t_2, 40 milliseconds later than t_0, contact 1 opens to mark the end of the first intelligence pulse. Contact S remains open. Contact 2 also remains open in accordance with the signalling code for the character Y. The line current drops to zero and the mark-to-space transition marks the beginning of the second intelligence pulse.

d. At time t_3, 60 milliseconds later than t_0, contact 3 closes, causing a space-to-mark transition as the line current rises to 60 milliamperes; the transition marks the end of the second intelligence bit and the beginning of the third bit. Contact S has remained open during this time and continues to do so.

e. At time t_4, 80 milliseconds later than t_0, contact 3 opens to mark the end of the third intelligence bit. Contact 4, in accordance with the signalling code, remains open; hence the line current drops to zero. The mark-to-space transition marks the beginning of the fourth information bit. Contact S is still open.

f. At time t_5, 100 milliseconds later than t_0, contact 5 closes in accordance with the signalling code, marking the beginning of the fifth information bit. The line current rises to 60 milliamperes causing a space-to-mark transition. Contact S remains open.

g. At time t_6, 120 milliseconds later than t_0, contact 5 opens at the same time that contact S closes. The opening of contact 5 indicates the end of the fifth information bit; closure of contact S indicates the beginning of the stop interval. The 60 milliampere line current is not interrupted by this action.

h. At time t_7, 148.4 milliseconds later than t_0, the end of the stop interval is reached, and the send mechanism is now ready for the transmission of the next character sequence.

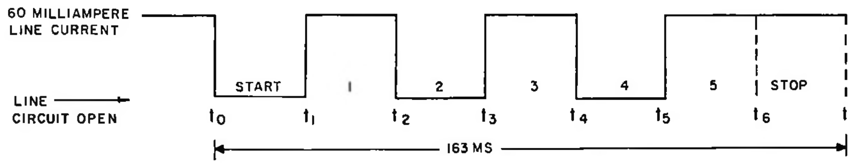


Figure 11-40D. Theoretical Alphabetical Character Diagram of Y

11-326U. RECEIVED SIGNAL. The signal at the receiving teletypewriter is the same as that sent from the transmitting equipment if ideal conditions are assumed. That is, the signal pulses operating the electromagnet of the receiving equipment are 60-milliampere pulses having instantaneous rise and fall times during operation.

a. The selector mechanism contains a series of cams mounted on a rotatable cam sleeve assembly as shown in figure 11-40E. The actual time of selection requires 4 to 5 milliseconds as the selector levers ride into the peak of the cam. Hence the mechanical selection process requires approximately 20 percent of the total transmitted pulse width of each information bit. Figure 11-40F shows a representation of the waveform for the character Y with the mechanical selecting intervals superimposed. The mechanical selecting intervals can be varied with respect to the beginning of the start pulse by a process of orientation.

b. The receiver of the teletypewriter shall correctly accept the next character interval by at least 6.6-unit intervals from the start mark-to-space transition of the previous character interval. As an example, a receiving teletypewriter must be prepared to correctly accept a stop to start transition which is 6.6, 6.7, 6.8, etc., unit intervals from the previous stop to start transition.

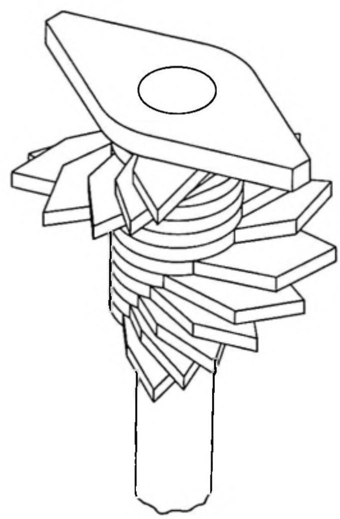


Figure 11-40E. Selector Cams

11-326V. ORIENTATION AND RANGE. The receiving mechanism, when oriented correctly is arranged so that it normally operates only during the central portion of

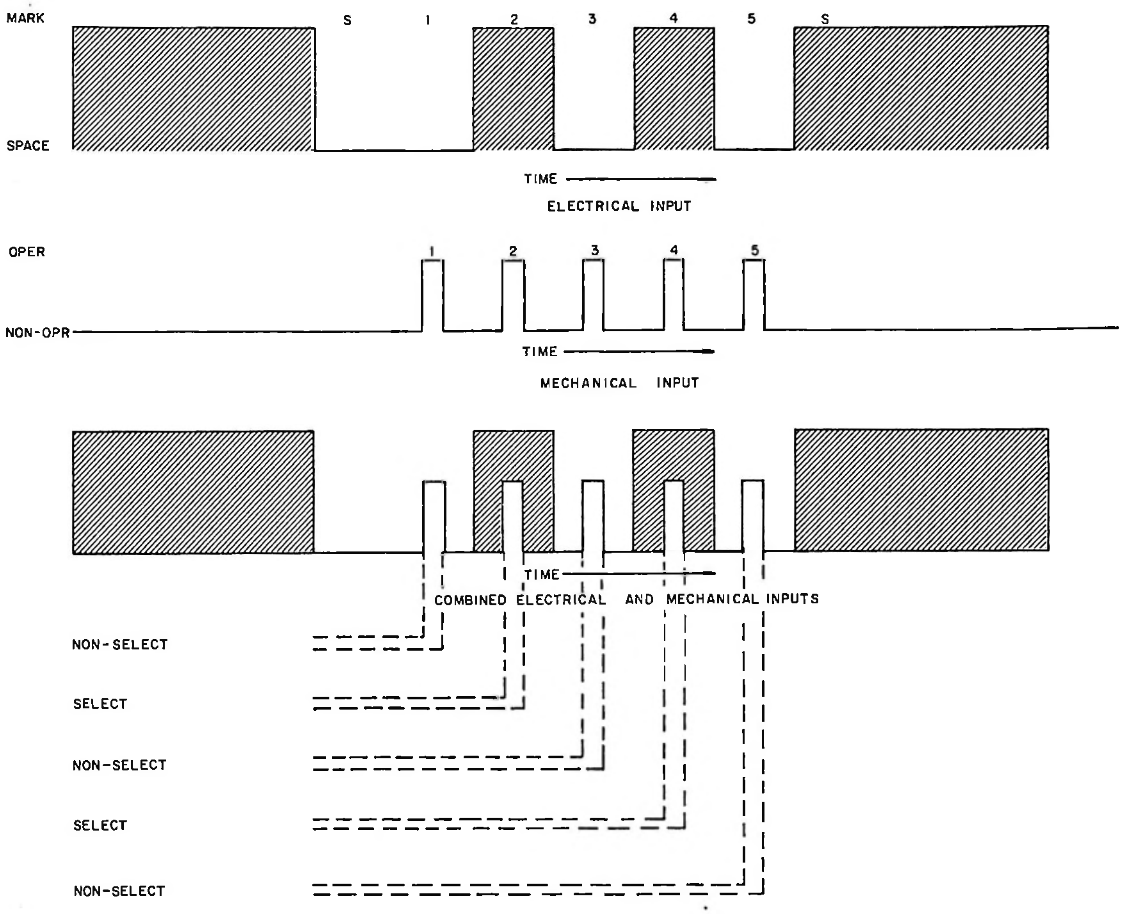


Figure 11-40F. Ideal Relationship of Electrical Signals to the Mechanical Selection

the received pulse. The time length from the edge of each selecting interval to the adjacent transition is 2/5 of the length of the impulse which indicates that the transitions may be shifted toward the selecting intervals as much as 40 percent of the length of the impulse before an error is recorded. The ideal situation is for the selecting intervals of the receiving unit to be at midposition with respect to the sending impulses.

11-326W. The receiving unit of the teletypewriter is equipped with a mechanism whereby a latch assembly assembly may be moved mechanically through an arc corresponding to the length of a perfect pulse unit. By this means all of the selecting intervals may be shifted with respect to the beginning of the start impulse over a range equal to a perfect impulse (22 milliseconds for a 45.5 baud modulation rate). This mechanism is known as the range finder and it is equipped with a scale from 0 to 120. One hundred divisions on this scale represent an arc equal to 100 percent of an impulse. Figure 11-40G shows a rangefinder stop arm and

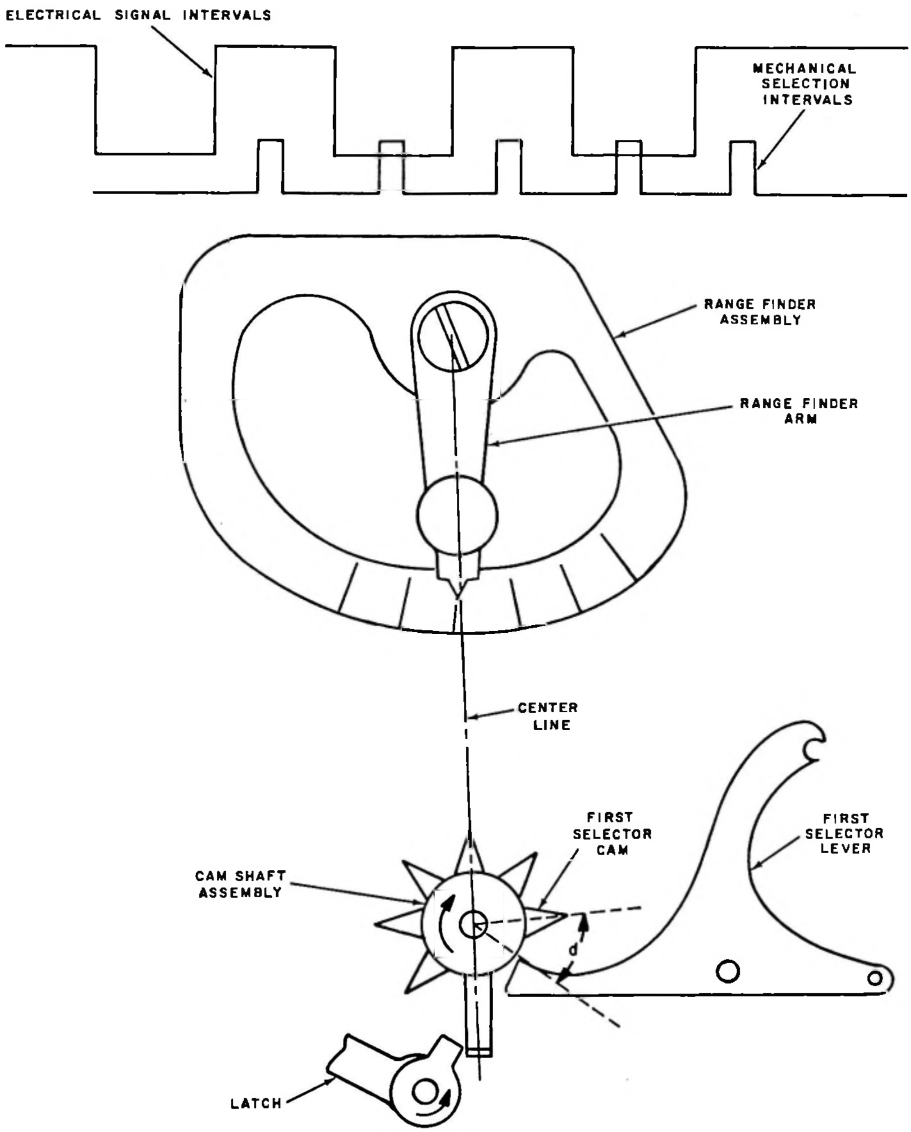


Figure 11-40G. Orientation Range Versus Selection Intervals--Mid Scale

selector cam-shaft assembly in relation to the rangefinder arm and scale. Of special importance, notice the distance between the first selector cam and the number 1 selector level. When a mark-to-space transition is received, the trip mechanism permits the latch to rotate in the direction of the arrow by means of force applied by the camshaft assembly which in turn begins to rotate in the direction of its arrow. Since the rangefinder arm is set to 50 on the scale, this is equal to 50 percent of a unit interval; assuming a 45.5 baud, unit stop, start-stop operation, it will require 33 milliseconds for the number one selector cam to reach the tip of the #1 selector lever, that is, a 22-millisecond start interval plus 11 milliseconds (50 percent unit interval) of the first intelligence impulse. The selection intervals are shown superimposed upon the selector armature mechanical motion pattern; this was shown more precisely in figure 11-40F. When the rangefinder arm is moved toward the lower numbers, it is in effect moving all selecting intervals closer to the mark-to-space transition of the start interval, that is, mechanical selection occurs at an earlier point in time with respect to the mark-to-space transition of the start impulse than it did previously. Figure 11-40H shows the condition where the rangefinder arm has been readjusted to 25 on the scale, the latch shown is mounted (effectively) on the rangefinder and therefore has moved to a new location with respect to its former position. The camshaft assembly continuously bears against the latch in its effort to rotate in the direction of the arrow. Notice that the distance between the #1 selector cam in relation to the #1 selector lever; this distance is much shorter than in the previous figure. Since the rangefinder arm is set to 25 on the scale, this is equal to 25 percent of a unit interval under the same conditions as before, and it will require 27.5 milliseconds for the #1 selector cam to reach the tip of the #1 selector lever. The 27.5-millisecond interval is made up of a 22-millisecond start pulse plus 5.5 milliseconds of the first intelligence impulse; this is an advance in time from the previous conditions. When the rangefinder arm is moved to the higher numbers, it is in effect moving all selecting intervals farther away from the mark-to-space transition; mechanical selection occurs at a later point in time than when the rangefinder setting of 50 or 25 was used. Figure 11-40I shows the condition where the rangefinder arm has been adjusted to 75 on the scale. The latch, mounted so it moves when the rangefinder arm moves, has also moved to a new position with respect to either of its previous locations. Notice the distance between the number one selector lever; this distance is much greater than the distance in either of the two previous figures. Since the rangefinder arm is set to 75 on the scale, this is equal 75 percent of a unit interval, using the same modulation rate, it will require 38.5 milliseconds for the number one selector lever to reach the tip of the number one selector lever. This 38.5-millisecond interval is composed of the 22-millisecond start interval plus 16.5 milliseconds of the first intelligence impulse; this condition represents a retardation from the previous conditions. When an impulse or signal is perfect, the range of the impulse for 100 percent, equals 0 to 100 on the rangefinder scale. Since the ideal position for the selecting intervals is the center of the impulse, the rangefinder arm should be set on 50. The selecting mechanism must use 20 percent of the impulse, and since the rangefinder is set to 50 on its scale, the 20 percent used to interpret the impulse will occur at the exact center of the impulse and leave a total of 80 percent, 40 percent on either side of the selecting intervals. This was shown in figure 11-40F. Therefore, the rangefinder arm can be moved along its scale 40 points each way from center without receiving a wrong impulse, resulting in a range of from 10 to 90, or 80 points total.

11-327. TELETYPEWRITER OPERATIONAL SPEEDS. There are four common teletypewriter operational speeds, which are ex expressed in words per minute or operations per minute. The operations per minute are exact ratings, since they are based upon the time interval required to send a spacing or marking intelligence pulse; these are 368, 404, 470, and 600 operations per minute. The words per minute are, therefore, only ap-

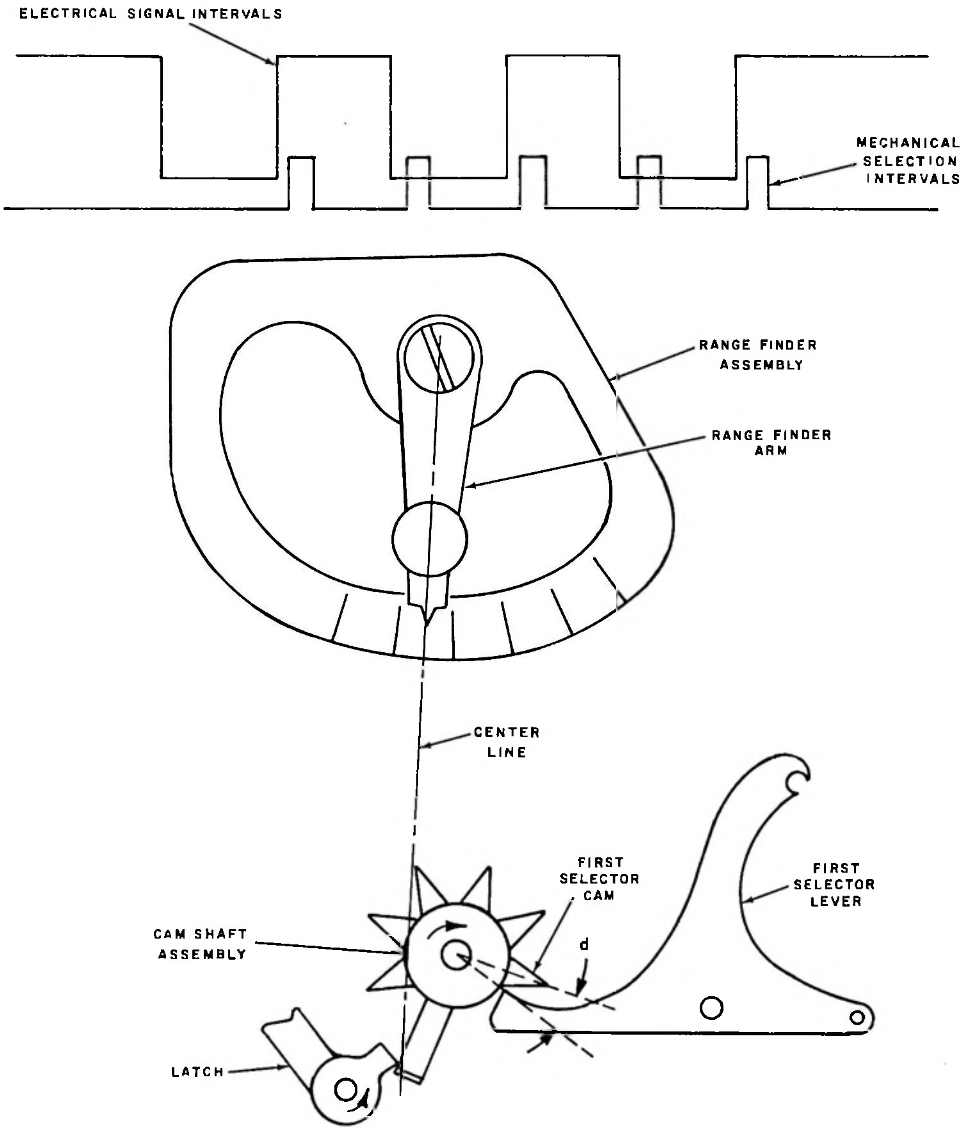


Figure 11-40H. Orientation Range Versus Selection Intervals--Low Scale

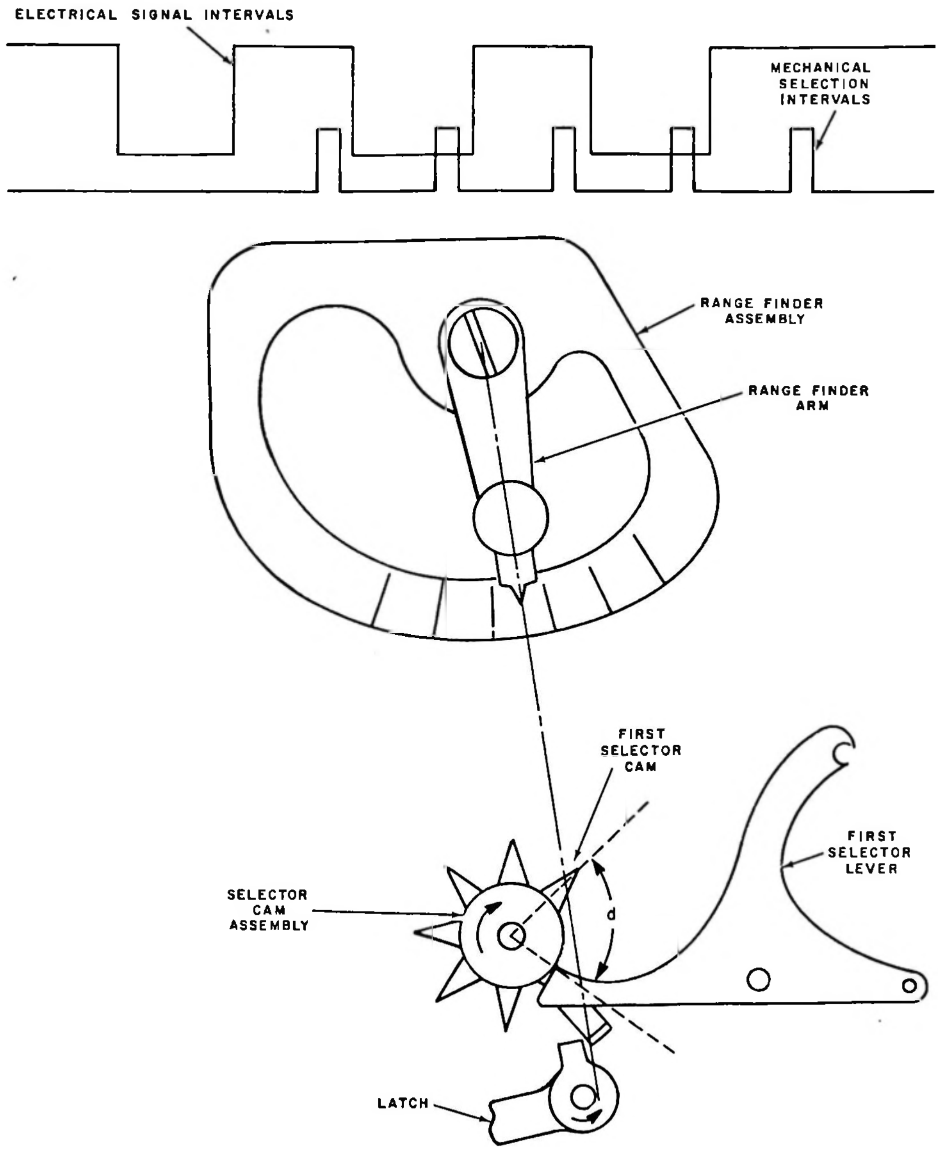


Figure 11-40I. Orientation Range Versus Selection Intervals--High Scale

proximate; they are 60, 66, 75, and 100 words per minute, corresponding to the respective operations per minute listed above.

11-328. At 60 words per minute, the length of the intelligence impulses and also the start impulse is 22 milliseconds. The stop impules at this speed is 9 milliseconds longer than the other pulses, to allow a longer time at the end of a character for mechanical speed of rotation variations inherent in any mechanical device.

11-329. As a general rule, Air Force teletypewriter equipment operates at a speed of 60 words per minute. This speed may be increased to a speed of 66 words per minute to work with British teletypewriters. Newer Air Force equipment is being used which operates at speeds of 75 or 100 words per minute; however, there are other important factors involved in the inter-operation of teletypewriter equipment. When sending directly to a circuit from a keyboard, the average operator will type at a nominal rate of 25 to 30 words per minute; because of the transmission of headings, non-typing functions, etc, the corresponding net message (text) speed is about 23 words per minute. When sending from tape at a normal machine speed of 60 words per minute, the average message (text) speed is about 50 words per minute.

11-329A. Receiving incoming messages from other signal centers or from tributary offices are received on reperforators in the receiving tables. After each message has cleared the reperforator, you tear the tape evenly, preferably at a letters signal, and mark off the corresponding number on the number sheet.

11-329B. A locking push-button release key is associated with the receiving reperforator of each circuit that is operated duplex (simultaneous sending and receiving). When no messages are being received, you can pull out this release-key button, if it is necessary to feed out the tape so that the last message can be readily torn off. Pushing in the button stops the tape feed action. The tape feed action also stops automatically if transmission is resumed while the tape is being fed out. Message tapes received from other semiautomatic signal centers over circuits equipped for automatic numbering feed out automatically whenever the circuit becomes idle. On such circuits, the release key need not be used for feeding out tape.

11-329C. Nonlocking push-button release keys are used for reperforators associated with circuits operated single (alternate sending and receiving), including those operating through a line finder. Each of these circuits has a guard signal lamp, which lights while messages are being received to indicate that the bank transmitter for sending to the same circuit is locked. You depress the nonlocking release key button to release the circuit and extinguish the guard lamp whenever the circuit becomes clear after the recepit of a message or group of messages. This unlocks the band transmitter and allows messages to be sent over the same circuit. In the case of line finder circuits, the release key also releases the reperforator so that it can receive messages from another calling line.

11-329D. A relay box on the line finder has four toggle switches, which act as make-busy (MB) keys for the four associated reperforators. These switches are normally turned on, but when turned off any switch acts to prevent incoming calls from being switched through to the reperforator associated with it. This has no effect on calls that are already connected through. These switches can be used to switch off any line finder reperforator for the replacement of a roll of tape or for any other reason. A make-busy (MB) key on the switchboard can be used to switch off four reperforators in a similar manner.

11-329E. The line finder connects a group of tributary or branch offices to a smaller group of receiving reperforators at the signal center. Each branch line extending to a tributary office is equipped with a band transmitter unit at the signal center and a teletypewriter set with a special motor-control circuit at the branch office. Each line finder reperforator at the signal center is associated with a rotary switch which can connect it automatically to any calling branch.

11-329F. Before sending a message the branch office operator calls by depressing a and releasing a break key. This causes the line finder to lock the band transmitter unit associated with this line and to connect the first idle receiving reperforator to the calling line. If all of the reperforators are busy, the call is held temporarily, and then is connected automatically to the first reperforator that becomes available. The teletypewriter motor at the branch office starts when this connection is made. You then send a message or group of messages, followed by spaces or blank signals to feed the tape out of the reperforator at the signal center. When the receiving operator releases the connection the teletypewriter motor stops.

11-329G. Receiving at a line finder branch office is entirely automatic. The teletypewriter motor starts automatically before the message and stops automatically at the end of the message. You never attempt to break in or send while a message is being received. If a message is being received, and you break in or send a message, then it is possible that the message that is being received will be interrupted or completely disconnected from the circuit.

11-330. CIRCUIT BANDWIDTH. An increasing number of teletypewriter circuits, wire and radio, are being operated on a 75 and 100 words-per-minute standards. Operation at each speed requires properly synchronized teletypewriter equipment and associated testing arrangements. In teletypewriter operation, the line speed in dot-cycles per second is set by the machine independently of the rate at which the keyboard is operated. The speeds are approximately 23 and 25 cps for nominal word speeds of 60 and 66 words per minute (368 and 404 operations per minute), respectively. Although teletypewriter signals are made up of pulses during which the operating current substantially reaches a steady value, they also contain alternating-current components of different frequencies; the range of essential frequencies from the lowest to the highest is called the signal frequency band. It is necessary to provide a circuit having the proper characteristics to transmit this band. In the case of a dc teletypewriter circuit, the band ordinarily extends up to about three times the maximum speed in dot-cycles per second. Thus, in teletypewriter operation (60 word-per-minute speed), a band about 75 cps wide is desirable; a band several hundred cps in width is required for the highest speeds. In carrier and radio telegraph, when the current of a single tone or frequency is keyed on and off, the required band is twice as wide, extending in each direction from the carrier frequency by the amount required for the dc case. When two frequencies, one for marking and another for spacing, and two channels are used, the bandwidth is again doubled. With tone-modulated radio teletypewriter operation, the whole voice band is generally used.

11-331. TYPES OF TRANSMISSION. The two general types of wire teletypewriter circuits are dc and carrier. Carrier teletypewriter circuits operating in the voice range between 300 and 2400 cps generally form the main transmission circuits for long- and medium-distance communication on land lines. Dc circuits are used especially for the shorter distances, including extensions within stations and branch circuits to outlying offices. Wire teletypewriter circuits will, in general, furnish dependable and accurate service if reasonable standards of circuit layout and maintenance are adhered to. Although radio transmission provides a high degree of flexibility for point-to-point communication from a station to any one of a large number of stations, and for broadcasting to several stations simultaneously, transmission is, on the whole, less dependable than that over wires. This is because of variable conditions in the transmission medium and the possibility of accidental or intentional jamming. The transmission impairment is not only in the form of displacements of transitions between marks and spaces (time distortion), but also in variations in strength of received signals and the occasional obliteration of the received signals by interference. Single-channel and multichannel teletypewriter circuits are operated satisfactorily over radio links within the limits of distances from a few thousand yards to many thousands of miles. Most of these applications make use of radio equipment operating within either the hf or vhf range, but in polar latitudes radio frequencies on the order of 50 to 200 kc are generally used. For operation in the hf range over radio circuits using sky-wave transmission, it is generally desirable to use diversity operation. Thus, signals may be received simultaneously at two or more locations separated by several wavelengths (space diversity), or each signal may be sent simultaneously at two or more frequencies, to be combined at the receiving station (frequency diversity). Where space is limited, two different adjacent antennas, such as vertical and horizontal, or differently oriented horizontals, may give some improvement by providing polarization-diversity reception.

11-332. DISTANCE LIMITS. The limiting length of a teletypewriter circuit, either wire or radio, is generally reached when a signal becomes so weak that it is incapable of actuating the receiving apparatus properly, when the waveshape is modified so that the time distortion of the teletypewriter impulses is excessive, or when the received signal strength is too low to override the interference. Usually the circuit can be extended by inserting a repeater before the limiting length is reached. In teletypewriter operation, a regenerative repeater may be used at an intermediate point. This repeater will automatically retransmit the received signals in practically perfect form if they have not suffered an amount of distortion which would cause errors in the copy in a teletypewriter at that point. The punching of a tape by a reperforator and retransmission from a transmitter-distributor are also a means of regenerating signals. This method is applicable to both wire and radio circuits.

11-332A. ELECTROMECHANICAL CONVERSION.

11-332B. There are principally two devices which perform the electromechanical conversion; (a) transmitter contacts, and (b) selector magnets. In the case of transmitter contacts, the conversion is from mechanical positions into corresponding electrical signals. In the case of selector magnets, the conversion is from electrical signals into corresponding mechanical positions.

11-332C. Transmitter contacts take several forms. The basic transmitter contacts of a teletypewriter can be of the form shown in figure 11-40J. The front contact assembly consists of six contact tongues stamped from a single sheet of nickel silver. The rear contact assembly is similar, except that the contact tongues are tempered and fitted with insulating tabs at their upper extremities. The two assemblies are insulated from each other, and from their mounting lugs, by insulation strips and

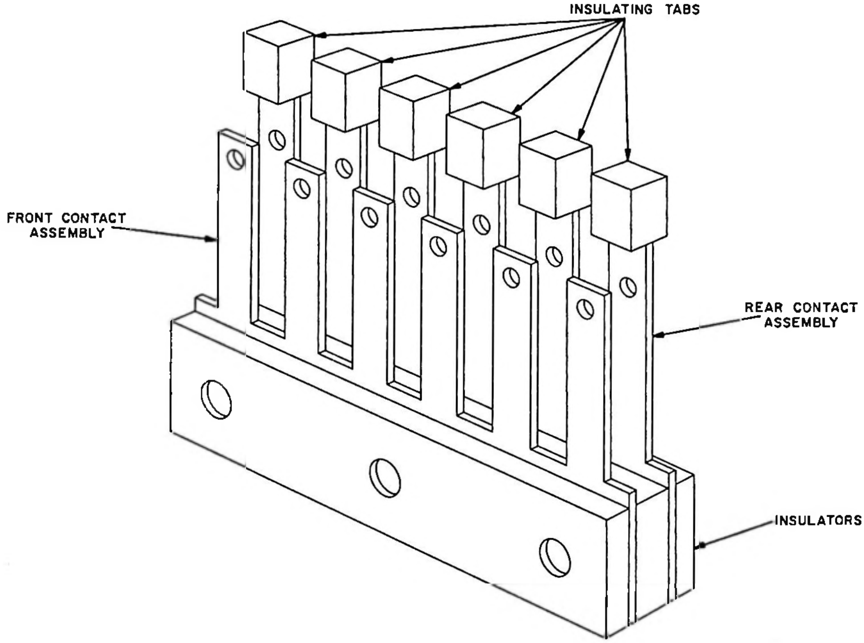


Figure 11-40J. Typical Teletypewriter Transmitting Contacts

bushings. The contacts in the figure are shown in their open positions in order to expose the contact surfaces; this is contrary to their normal positions under these same conditions of isolation from the control mechanism. The actual positions of the contacts are controlled by the transmitter mechanism bearing against the upper insulating tabs of the rear contact assembly. There is one contact tongue for each of the intelligence intervals plus an additional contact tongue to generate both synchronization pulses. Sequential commutation of the contact tongues is under the control of a camshaft. Contacts in this form are limited to neutral signalling modes; where polar signalling is desired, an external conversion device must be used.

11-332D. Another form of transmitting contact, called a pivotal generator, is shown in figure 11-40K. The pivotal generator is capable of polar, as well as neutral, signalling modes. One switching circuit is completed through the marking contact, the contact toggle, the toggle link, and the associated spring. Another switching circuit is completed through the spacing contact, the contact toggle, the toggle link, and the associated spring. The contact toggle is electrically isolated from the toggle extension (and the remaining mechanical parts) by means of an insulating bushing. Both the marking and the spacing contacts are stationary; electrical switching is accomplished by mechanical motion imparted to the contact toggle via the toggle extension. This pivotal generator performs the total switching requirements for intelligence and synchronization intervals; thus the contact toggle must open or close the signal line for each impulse transmitted. Sequential

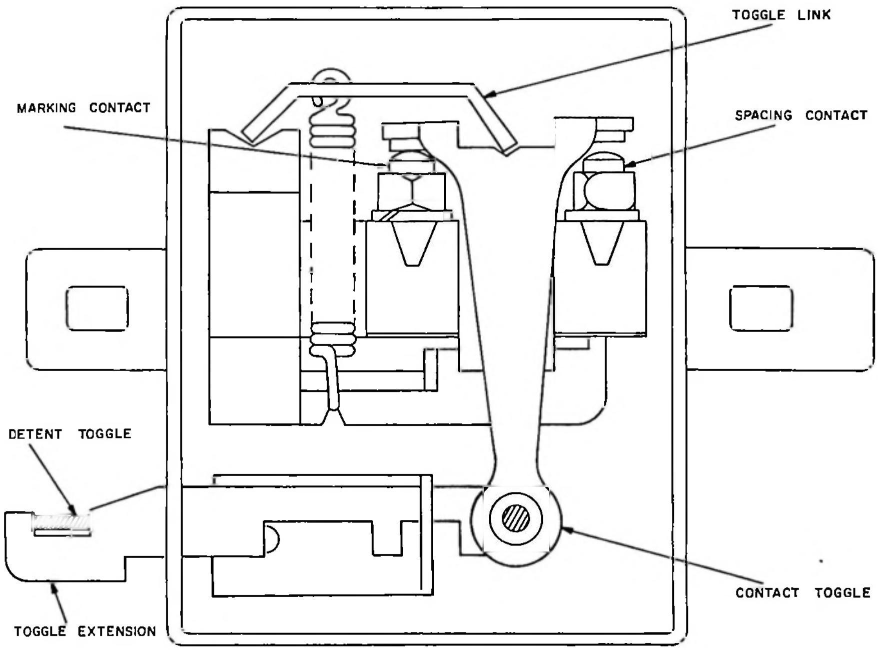


Figure 11-40K. Pivotal Generator

commutation of the contact toggle is under the control of a camshaft.

11-332E. The transmitter contacts shown in figure 11-40L are also capable of either polar or neutral transmission modes. One switching circuit is completed through the upper contact screw, the contact tongue, the contact lever spring, and the terminal. The remaining circuit is the same, except that the beginning point is the lower contact screw. The contact tongue is insulated from the contact lever. Commutation is controlled by a perforated paper tape; sequencing is controlled by a distributor assembly.

11-332F. The distributor consists of the distributor disk and the distributor brush assembly (figure 11-40M). The distributor disk consists of two concentric, conducting rings mounted on a disk of insulating material. The outer ring is divided into seven segments; the inner ring is a continuous conductor. Segments of the outer ring correspond to the impulses of the teletypewriter code; that is, they include a start segment, five code segments, and a stop segment. The start segment precedes the no. 1 segment, and the stop segment follows the no. 5 segment (figure 11-40N). The distributor brush assembly consists of a pair of brushes clamped in a metal brush holder. The brush holder is attached to the brush holder arm, which is mounted on the upper end of the main shaft. The brushes are spaced in the brush holder the same distance apart as are the concentric rings on the distributor disk, and they revolve with the main shaft. The distributor brush assembly makes one complete revolution to transmit the code combination for each character or function. When the distributor brush passes over the start segment, a

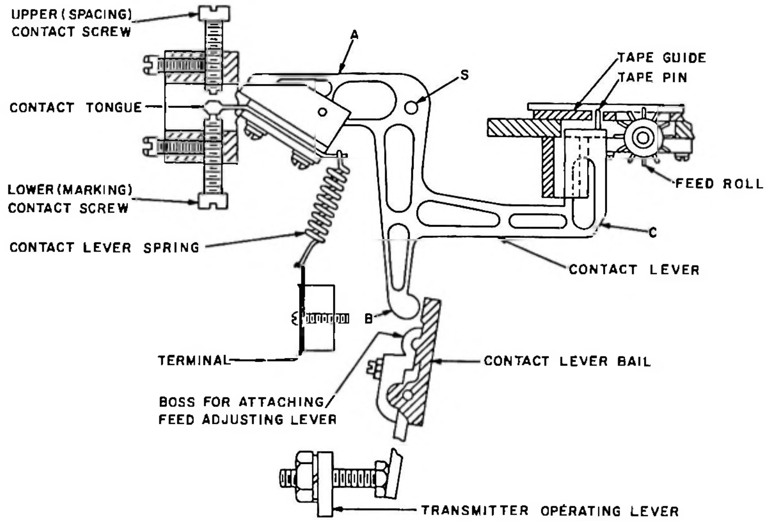


Figure 11-40L. Tape-Sensing Mechanism

spacing impulse is always transmitted; when the brush passes over the stop segment, a marking impulse is always transmitted. These two impulses keep the transmitter distributor and the receiving teletypewriter equipment in step or in synchronism through control of the starting and the stopping of the receiving teletypewriter mechanisms. The receiving mechanisms start to revolve when they receive the start impulse from the transmitter distributor. If all teletypewriter motors in the circuit are operating at the proper speed, the receiving mechanisms select the no. 1 impulse while the distributor brush is passing over the no. 1 segment. The no. 2, no. 3, no. 4, and no. 5 impulses are selected as the distributor brush passes over the no. 2, no. 3, no. 4, and no. 5 segments in a like manner. The receiving mechanism stops after receiving the stop impulse from the transmitter distributor. Because the receiving mechanisms revolve slightly faster than the distributor brush assembly, the receiving mechanisms come to a complete stop before the brush assembly completes the revolution and sends the next start impulse. Therefore, slight differences in speed are not accumulated because the distributor brush assembly and receiving teletypewriter mechanisms start together at the

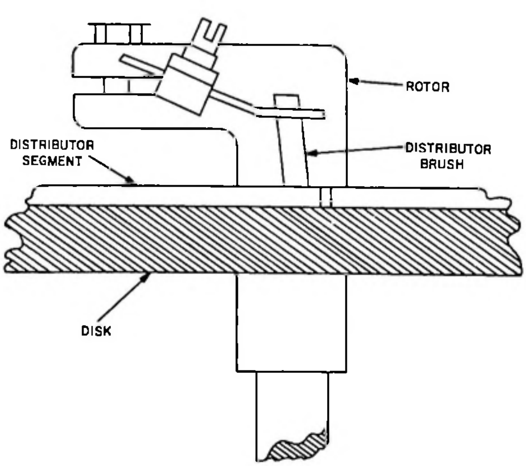


Figure 11-40M. Distributor Disc Assembly (Side View)

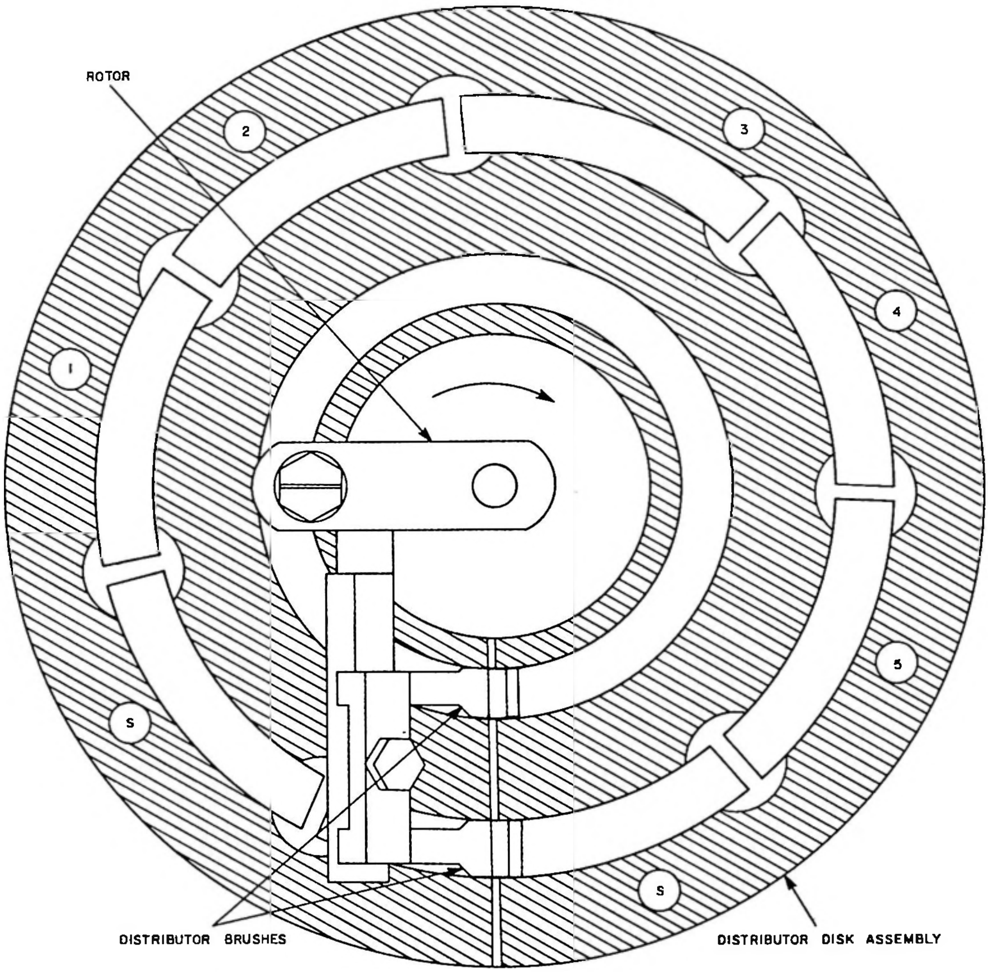


Figure 11-40N. Distributor Disc Assembly (Top View)

beginning of the transmission of each character of function.

11-332G. The selector magnets of a teletypewriter convert the electrical signal variations into mechanical position variations, as is shown in figure 11-40O. When the electromagnets are energized, the armature is pulled toward the electromagnetic polefaces against the tension of a spring; the armature travel is limited by an adjustable stop to provide one specific mechanical position. This position corresponds to an electrical marking impulse and is called the marking position. When the electromagnets are de-energized, the spring pulls the armature away from the magnetic polefaces to provide another specific mechanical position; since this position corresponds to an electrical spacing impulse, it is called the spacing position. The armature travel to the spacing position is also limited by an

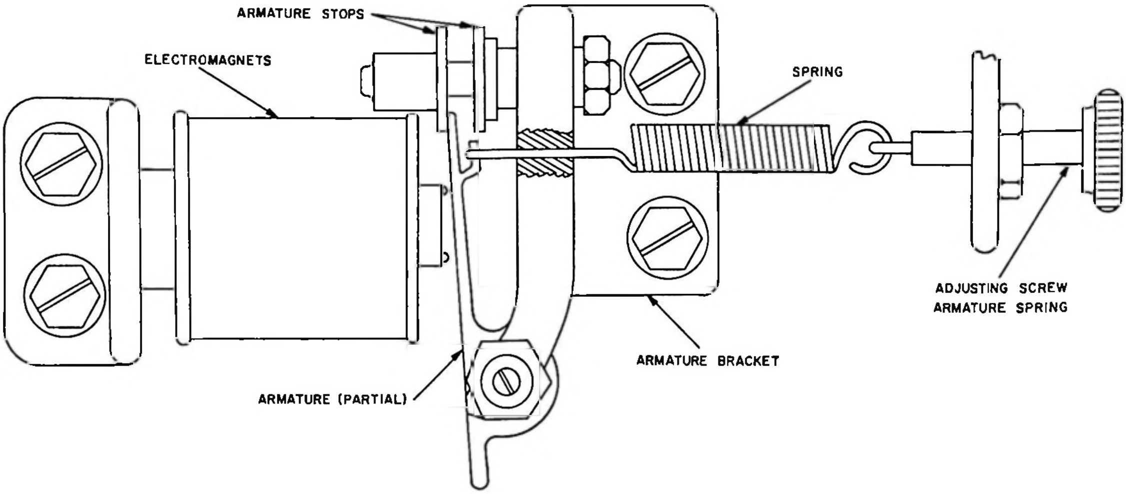


Figure 11-40O. Typical Teletypewriter Selector Magnet Unit

adjustable stop. Selector magnets of this type are suitable only for use in neutral signalling modes.

11-332H. In some teletypewriter equipments, the selector magnets are, in themselves, polar relays. The selector magnet, shown in part A of figure 11-40P, consists of a permanent bar magnet, an armature and two windings which are wound around each arm of a U-shaped silicon-steel core. The magnet is magnetically balanced with respect to either arm of the core. The line windings are in series with the signal line and the code impulses pass through these windings. The bias windings are supplied with a local battery. When a marking impulse is received, current flows in the line windings and the armature is attracted to the right core. When a spacing impulse is received, there is no current flow in the line windings and the armature is moved to the left core under the control of the bias windings.

11-332I. The selector magnet is constructed so that when current is not present in any of the windings, there is a balanced magnetic field, as is shown in part B of figure 11-40P. Under this condition, the permanent magnet is the only source of magnetism. Note that the magnetic flux at the lower end of the magnet (north pole) divides equally and returns to the south pole through both arms of the core and both sides of the armature. The magnetic pull of the cores on the armature is therefore equal, when all windings are not energized.

11-332J. When operating in polar circuits, the bias windings are not used since the only purpose of the bias winding in the neutral circuit is to move the armature to a spacing condition when a space signal is received. During a marking impulse, current flows through the line windings. As is shown in figure 11-40P, terminal 8 is positive and terminal 4 is negative. The magnetic field set up by the line winding around the left arm opposes the field of the permanent magnet, and a minor pull is present on the left side of the armature. The magnetic field, set up by the line winding around the right arm of the core, is poled to aid the field of

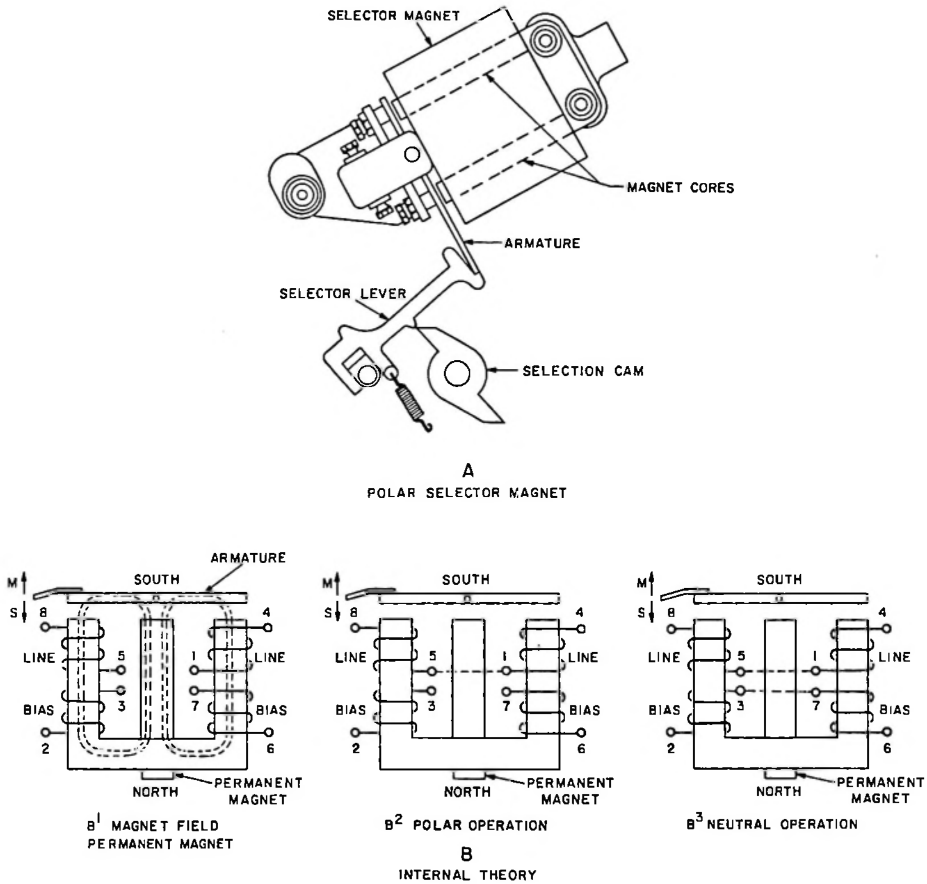


Figure 11-40P. Polar Selector Magnet

the permanent magnet and, therefore, the right end of the armature is pulled toward the right arm of the U-shaped core. The left end of the armature (which includes the armature blade), moves to the marking position and into the path of the selector levers. During reception of a polar-spacing impulse, current flow in the line windings is reversed, and the opposite of the marking impulse process occurs. The magnetic pull on the right end of the armature is weakened and the pull on the left end is increased, causing the armature to be drawn toward the left arm of the core.

11-332K. When operating in neutral circuits, the bias windings are wired in series and are constantly energized. The magnetic field set up by the left bias winding is poled in order to aid the magnetic field of the permanent magnet. The field of the right bias winding opposes the field of the permanent magnet. During reception of a neutral spacing impulse, current is not present in the line windings. The combined magnetic fields of the energized bias windings and the permanent magnet cause the blade end of the armature to be pulled toward the left core, and away from the selector levers. Current flows in both line windings during reception of a neutral marking impulse. The current value in a line winding normally is twice the bias current value. The field set up by the energized line winding around the left core opposes the combined magnetic fields of the left bias winding and the permanent magnet. The energized line winding around the right core aids the field of the permanent magnet and opposes the field of the right bias winding. Therefore, a marking impulse in the line windings will cause the right end of the armature to be pulled toward the right core arm, and will cause the left (blade) end to move into the path of the selector levers (marking position). Adjustment of the bias current value to obtain the most advantageous machine range is accomplished with a potentiometer that is connected in series in the bias circuit.

11-332L. Figure 11-40Q shows a conversion device which uses electromechanical principles to accomplish electrical-to-electrical conversion, and the device is known as the differential wound polar relay. The fundamentals of the differential relay depend upon winding a magnetic core with two equal but opposing windings so that if equal currents flow in the same direction through both windings, the magnetic field produced by one winding will be exactly neutralized

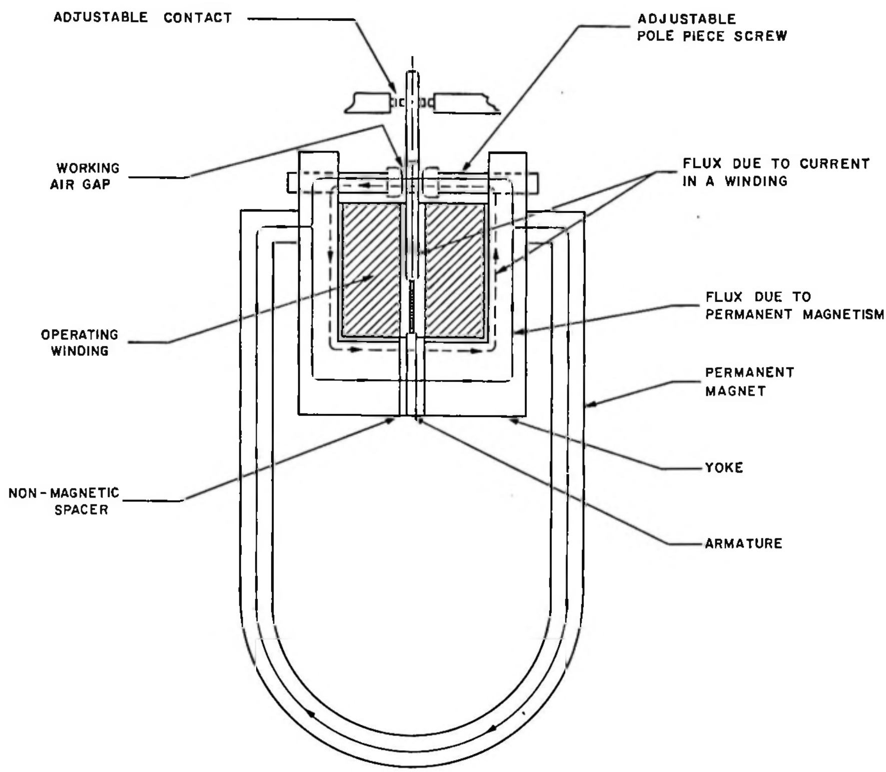


Figure 11-40Q. Magnetic Circuit of Differential Polar Relay by that produced in the other winding. Furthermore, by the use of a permanent magnet and a split magnetic circuit, such a relay may be polarized.

11-332M. Examination of figure 11-40Q shows that the adjustable pole piece screws are of opposite polarity due to the presence of the magnetic flux of the permanent magnet (solid lines in the figure). Current in either of the parallel operating windings will magnetize the armature (dotted lines in the figure) and cause it to move toward the adjustable pole piece screw of opposite polarity. A reversal of current in both operating windings, or a reversed current of appreciably greater magnitude in one of them, will reverse the polarity of the armature and cause it to move to the other pole piece screw. One winding is termed the line winding and the other winding is termed the bias winding. A constant current of 30 milliamperes (ma.) is passed through the bias winding, and a signalling current varying in amplitude from 0 ma to 60 ma (in the opposite direction) flows in the line winding. Thus with line current equal to zero in the line winding corresponding to a spacing condition, the armature is held in one position; this is conveniently called the spacing contact. When a 60 ma line current, corresponding to a marking condition, is sent through the line winding of the relay, the armature moves to the opposite position; this is called the marking contact.

11-332N. The receiving electromechanical devices are sensitive to operating values. A polar relay operating in the neutral mode on a non-rectangular waveform has the problem of deciding (in a logical sense) what its operating point shall be, as is shown in part A of figure 11-40R. While the exaggerated conditions of the figure are not indicative of typical waveforms, they show the logical decision accomplished by a relay, vacuum tube, or transistor device. As accepted practice, the operating point of a polar relay is set approximately to the half-power point of the waveform. The selector magnet of printing telegraph equipment, on the other hand, tends to operate in many cases with as little as 10 percent decrease below its operating current, as is shown in part B of the figure. The line AA_1 indicates the approximate operate-release points of a selector magnet while line BB_1 indicates the approximate time of relay armature change from space to mark positions or from mark to space positions, whichever case applies.

11-332O. SIGNAL DISTORTION.

11-332P. Telegraph transmission is concerned with the reproduction of the transmitted telegraph messages at the receiving station with satisfactory speed, and without error, and without interfering with other signal activities. Up to this point, it has been assumed that these objectives would be satisfactorily accomplished by the various circuits and apparatus of which telegraph circuits are composed. However, the characteristics of these circuits and apparatus

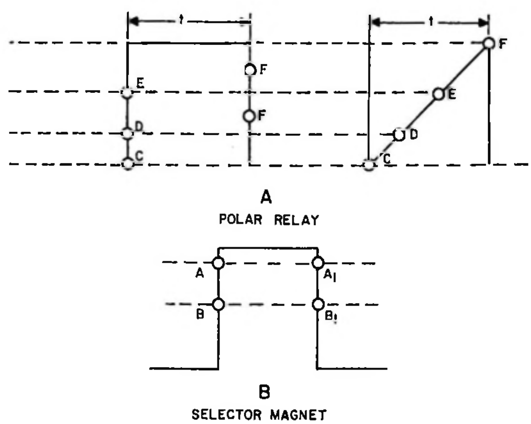


Figure 11-40R. Receiving Device Operating Points are such that the signals transmitted sometimes fail to reproduce accurately at the receiving station the same character that was transmitted. Failure to reproduce the desired character can be due to either mechanical or electrical difficulties; in either case, the cause of failure is attributed to distortion. There are several types of telegraph signal distortion; these are defined as follows:

a. Bias distortion is distortion which affects a two-condition (or binary) modulation in which all the significant intervals corresponding to one of two significant conditions are uniformly of longer or shorter duration than the corresponding theoretical durations. This type of distortion can vary from one hour to the next, or with one adjustment of the equipment to another. However, only a few equipments in a telegraph transmission system, such as maladjusted relays or detuned radio receivers, are capable of producing bias distortion.

b. Characteristic distortion is distortion caused by transients, resulting in consistent malformation of a specific element or elements. Such transients result from modulation in the transmission system, since the elements of the signals are affected by the presence of changing currents due to the capacitance, the inductance, and the resistance which are present. The degree of distortion depends on the magnitude of the condition. The time required for the current to change from one condition to the other may at times exceed the minimum time interval between transitions in the signal. Some transitions must then occur while the current is still in the process of changing from the previous transition.

c. Fortuitous distortion is distortion resulting from causes generally subject to random laws, as for example, accidental irregularities in the operating of the apparatus and moving parts, disturbances affecting the transmission channel, and so forth.

d. End distortion is the shifting of the end of all marking intervals from their proper positions in relation to the beginning of the start pulse.

e. Cyclic distortion is distortion which is neither characteristic, bias, nor fortuitous by type and which, in general, has a periodic character. Its causes are, for example, irregularities in the duration of brush contact time in transmitter distributors or interference by alternating currents, and so forth.

11-332Q. Special consideration must be given to distortion in synchronous operational modes. The transmitting and receiving telegraph equipments must be accurately timed in any synchronous system. Timing is usually controlled by very accurate oscillators, so that once synchronized, the telegraph equipments will remain closely synchronized for several hours, even after the connecting circuit has been removed. In addition, there must be some method of maintaining the receiving device in exact synchronization with the transmitting device. The two devices must operate at the modulation rate and they must maintain the same relative phase, since the phase relationship determines the positioning of the selection interval. To maintain relative phase, the receiving device must be capable of sensing the modulation rate of the transmitting device, and of automatically adjusting to that rate. In synchronous signalling, measurements of the maximum early and late deviation must be added, since synchronous equipment samples each element without regard to the position of the proceeding element.

11-332R. NEUTRAL OPERATION. One

method of telegraph transmission is called neutral, make-break, single-current, or on-off signalling. Neutral signalling employs a signal of some magnitude to represent one binary state and the absence of a signal (noise) to represent the second binary state. Thus in neutral operation, the magnitude of the signal determines whether a marking or a spacing impulse is being transmitted.

11-332S. In field installations, and in most fixed-station applications operating over short, well-insulated lines, the neutral method is used. This method requires a minimum of terminal equipments, and it is easy to install and operate; its greatest advantages are its simplicity and its low maintenance requirements. The chief disadvantage of the neutral system is viewed from the standpoint of signal transmission, because the distortion of signals over the line restricts the length of the circuit. The parameters of neutral signalling require that the tolerances with regard to voltage and current be held quite close, generally within 5 percent for reasonably reliable performance. In the neutral signalling mode, it is essential that all lines be adjusted to the same current value so that reliable patching services can be accomplished.

11-332T. Figure 11-40S shows a two-station teletypewriter circuit; teletypewriter signals when transmitted over a long line are affected by line characteristics that tend to distort the signals. Capacity and inductance are the most important of these distortion characteristics. The capacitance of the entire circuit is lumped together in this figure. When a transmitting contact closes, the first rush of current flows only in part to the receiving station since the current is diverted to the capacitor until the capacitor is charged to full value. As the capacitor is charging, it also tends to

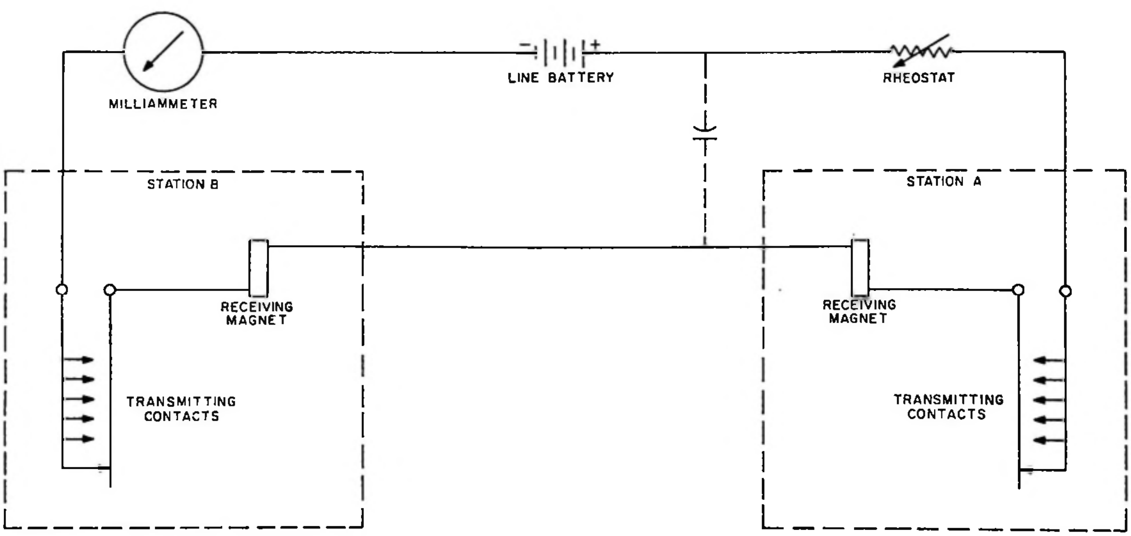


Figure 11-40S. Simple Two-Station Teletypewriter Circuit discharge along the same path as the originally applied voltage, thus allowing an increasing amount of current to flow through the relay. When the capacitor is fully charged, it will be of the same potential as was originally applied, and all the current will flow through the receiving relay. This delaying action causes the waveshapes of the teletypewriter signal to be rounded at the beginning and results in a delayed space-to-mark transition. When current begins to flow through the coil of the relay, a counter electromotive force is set up in opposition to the applied voltage. This action tends to delay current flow through the coil and helps to divert the initial current flow to the capacitor. When the current is increased to its maximum value, the magnetic flux remains constant and the counter electromotive force no longer opposes coil current. These developments indicate that line signal current does not rise at once to its final value, but increases along a curved path as is illustrated in figure 11-40T. The combined effects of capacitance and inductance cause a time delay in the space-to-mark transition resulting in curved leading edges.

11-332U. When the transmitting contact opens, circuit inductance and circuit capacitance are once more brought into play. The inductance of the relay tends to maintain current flowing until the capacitor cannot discharge through the open contact, and therefore, the capacitor will discharge through the relay during the mark-to-space transition; the sloped trailing edge is shown in

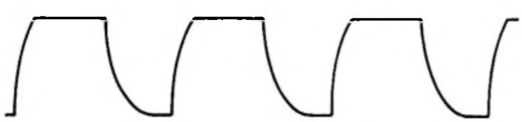


Figure 11-40T. Typical Neutral Signal

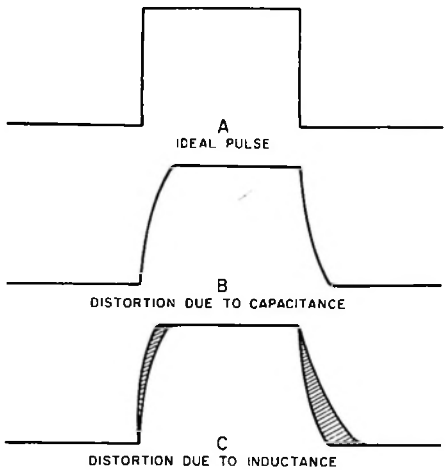
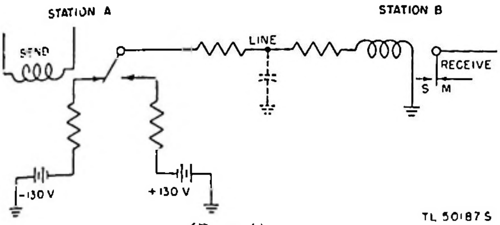


Figure 11-40U. Comparison of Waveshapes-Neutral Signals

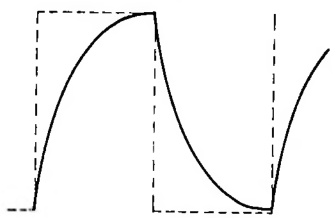
part C of figure 11-40U. Figure 11-40U is a comparison of the waveshape transmitted (part A), the waveshape distorted by capacitance, and the waveshape distorted by combined inductance and capacitance. The shaded portion of part C of the figure represents the effect of inductance.

11-332V. Figure 11-40T shows a typical neutral telegraph signal. Note the lack of symmetry and the fact that the rate of change at the half-power point on the mark-space transition is significantly slower than on the space-mark transition.

11-332W. POLAR OPERATION. Part A of Figure 11-40V shows a two-station teletypewriter circuit. Notice that two electrical sources are employed; the value of the line current is equal for both sources, but the current flow will be changed in direction by the operation of the transmitting contacts. Polar operation is two-current signalling in which current flowing in the positive direction is one binary state, and current flowing in the negative direction is the other binary state. The standard signalling sense for polar signalling shall be a positive voltage or current referenced to signal ground causing the re-



(Part A)



(Part B)

Figure 11-40V. Comparison of Waveshapes-Polar Signals

ceiving device to remain in a quiescent condition. In polar signalling, noise represents a third state, and as a general rule, so long as the signal information states are slightly greater than the noise, the ability to discriminate between the two binary states is adequate. It is important to note that with polar signalling, so long as there is some indication of directional current, the information contained in the signal is generally received correctly. This is not the case in neutral signalling, because there is a wide marginal area around which the approximate magnitude of the signal is known, but due to impulse noise and the sensitivity adjustment of the receiving equipment, the magnitude of the signal remains in doubt.

11-332X. In polar signalling, there are no close tolerance requirements with respect to the magnitude of the signal. Polar circuits will perform quite adequately from 3 ma to 70 ma so long as the marking and the spacing currents are within about 10 percent of each other. Polar circuits are not exempt from line characteristics; space-to-mark and mark-to-space transitions are delayed by capacitance to ground plus inductance of the relays. Since the action of the two properties working together has already been explained under neutral operation, it is unnecessary to explain it again. For the sake of simplicity, only the capacity will therefore be considered. When the line current is marking, the voltage on the capacitor representing the capacity to ground is negative, and while the line current is spacing, this voltage is positive. When a marking impulse is transmitted at station A, the current flowing in the line must build up a charge on the capacitor just as in the neutral system, and thus delays the buildup to a full marking-current value in the receiving relay of station B. When the line current is changed from marking to spacing, the voltage on the capacitor must make a complete change from negative to zero to positive. The line current from the discharging capacitor flows in the same direction as the originally applied line current and tends to keep the line current flowing in the relay. Then, when positive spacing current begins to flow, it builds up to a charge on the capacitor and thus delays the buildup of spacing current to its full value. These two actions of the line capacitance combine to make the transition of the line current from marking to spacing a gradual change as shown in part B of figure 11-40V. The fact that the space-to-mark and the mark-to-space waveshapes are identical is the most valuable feature of polar operation because the space and mark impulses repeated by the relay are of the same length even though the original signals to the relay are distorted.

11-332Y. EXTERNAL DISTORTION. Distortion of the electrical signal that is not the result of a mechanical malfunction is external distortion and may occur anywhere in the line circuit. If additional equipment is added to a telegraph circuit, the additional inductance introduced by the relay in series with the line can change the slope of the signal transitions and new equipment adjustments may be required.

11-332Z. Another factor that will change the length of the signal is a change in the applied signal level or a change in series resistance. Consider the case of increasing the current by using a higher voltage or by removing series resistance from the circuit. Naturally, the operating and release current values of the relay before and after the change remain the same. Since the increase in current with constant inductance steepens the sides of the curve, the net result is an increase in the length of the signal.

11-332AA. In practice, the fluctuation in line currents due to leakage can affect circuit adjustments. To a degree, leakage can be compensated for by your using grounded battery connections at both ends of the circuit, with the connection at one end being positive and the connection at the other end being negative, instead of using a single battery at one end and only a ground at the other.

11-332AB. The change from the spacing to the marking condition is the space-to-mark transition, abbreviated as s-m; the change from the marking to the spacing condition is the mark-to-space transition, abbreviated as m-s. At the instant that the key at the transmitting station is closed, the line current starts to increase in the receiving relay as indicated by the waveshape, but does not reach the operating point of the polar relay until a few milliseconds later. The delay between the closure of the transmitting key and the operation of the receiving relay is called the space-to-mark transition delay and is abbreviated s-mtd. In a similar manner, when the sending key is opened, the line current in the receiving relay does not become zero instantaneously. The receiving relay will be held on its marking contact for an interval of time after the circuit is opened at the transmitting end. This time delay is a mark-to-space transition delay and is abbreviated as m-std. The magnitude of either of these delays ranges from a fraction of a millisecond to several milliseconds. The s-mtd and m-std are determined entirely by the characteristics of the circuit, and though the two delays may not be equal, each transition delay will always be constant for any given circuit, under any given set of adjustments. Each mark, regardless of length, must start with an s-m transition and end with an m-s transition. The s-mtd cuts off the beginning of each mark and the m-std adds to the end of each mark. If the two delays are equal, the length of each mark will be unchanged by transmission over the circuit. Each space interval, regardless of length, begins each space with an m-s transition and ends with an s-m transition. The m-std cuts off the beginning of each space interval and the s-mtd adds to the end of each space. Each delay thus has the opposite effect on a spacing interval that it has on the marking interval. If the two delays are equal, the length of each space will be unchanged by transmission over the circuit; transmission is considered perfect if the received marking and spacing intervals are exactly of the same length.

11-332AC. BIAS DISTORTION. If the two delays are not equal, as for example if the m-std is greater than the s-mtd, all marking intervals will be lengthened, and all spacing intervals will be shortened. This is a common condition in working circuits and it is called marking bias because the circuit lengthens the signal-marking intervals. In the reverse condition, the s-mtd is greater than the m-std, all spacing intervals will be lengthened and all marking intervals shortened; this type of condition is termed spacing bias. The amount that is added to or subtracted from each marking or spacing interval due to a bias condition is equal to the difference between the values of the m-std and the s-mtd expressed in milliseconds.

11-332AD. A marking bias is called a positive bias and a spacing bias is a negative bias. If the difference between the s-mtd and the m-std is always taken as the m-std minus the s-mtd, the sign of the result will automatically be the proper sign to affix to the bias. Thus the equation for bias distortion is as follows:

$$(\text{m-std}) - (\text{s-mtd}) = \text{bias}_{\text{(milliseconds)}}$$

As an example, if the m-std of a circuit is 6 milliseconds, and the s-mtd is 3 milliseconds, the millisecond bias is + 3, indicating that every mark, regardless of length will be increased 3 milliseconds, and correspondingly, every space regardless of length will be decreased 3 milliseconds. It is important to keep in mind that a millisecond bias condition is determined entirely by equipment, line facilities, over-all length, and other factors of the circuit and will be constant for any given circuit, regardless of modulation rates of kinds of signals.

11-332AE. The effect on transmission, however, of a given millisecond bias condition, does vary with the length of the marking and the spacing intervals transmitted, even though the millisecond bias condition itself is constant. As an example of this, consider a manual telegraph circuit where the dashes (long marks) are normally about two and one-half to three times the length of the dots (short marks). For a manual telegraph, the length of the dots and the dashes decreases as the speed of transmission increases. As a typical example of this, first assume the existence of a slow transmission speed where the dots are 30 milliseconds long. A millisecond bias condition of +10 will make the dots 40 milliseconds long and the dashes 100 milliseconds long. The signals will be quite readable since the three-to-one ratio has not been significantly changed. Next assume the existence of a much faster speed where the dots are 5 milliseconds long and the dashes are 15 milliseconds long. The same +10 bias will make the dots 15 milliseconds long and the dashes 25 milliseconds long. Greater difficulty will be experienced in reading these signals since now the dashes are not even twice the length of the dots.

11-332AF. The waveshape of a typical mark signal in a neutral circuit operating with a line current of 60 ma and having capacitance to ground is shown in figure 11-40W. The horizontal lines A, B, and C represent different values of relay-biasing currents. The relay operating and releasing points (designated by heavy dots) are indicated for each of these three values of biasing current. When the normal biasing current of 30 ma (line B) is used, the length of the mark signal is that indicated by T_b. It is obvious that increasing the relay biasing current increases the s-mtd and shortens the m-std. This produces spacing bias, since it reduces the marking-signal length. A reverse condition results from lowering the bias current to a value below the normal value; that is, the marking signal is increased in length, as shown by T_c, because the s-mtd decreases while the m-std increases. The same effect as that obtained by raising the bias current, which shifted the relay operating points toward the narrow part of the wave, is obtained if the biasing current is held constant and if the line

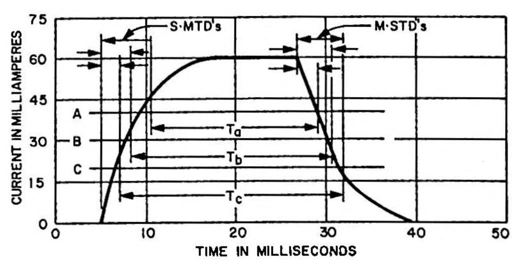


Figure 11-40W. Effect of Relay-Biasing Current on Signal Length current decreased. This in effect shifts the narrow part of the wave toward the relay operating points, and again the marking impulse is shortened. This is illustrated by the waveshape for the low line current in figure 11-40X.

11-332AG. On the other hand, increasing the line current while the biasing current remains the same, increases the current at all points on the waveshape and effectively shifts the broader part of the wave toward the operating points. This lengthens the impulse as illustrated by the waveshape for the high line current in the figure. In other words, increasing the line current in a neutral circuit tends to produce marking bias and decreasing it tends to produce spacing bias.

11-332AH. Consider a one-way polar circuit using a ground return. The sending relay connects —160 volts to the line for the marking condition and +160 volts for the spacing condition. The resistance at the sending end is adjusted by means of a potentiometer connected in the line circuit so that the current is normally about +35 ma for the marking condition and −35 ma for the spacing condition. These are the steady-state values. In this, as in other circuits, the

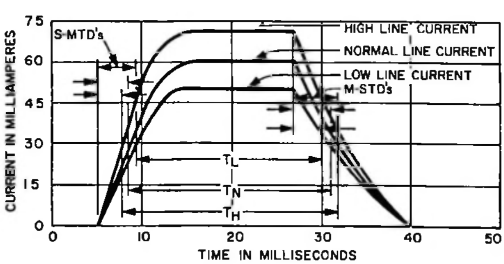


Figure 11-40X. Effect of Line Current Magnitude on Signal Length

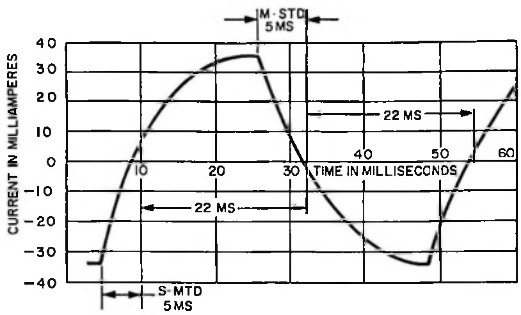


Figure 11-40Y. Signal Wave Shapes in Polar Telegraph Circuits

change of the line current from marking to spacing (m-s transition) and from spacing to marking (s-m transition) will be delayed because of the capacitance between the line and ground. The transition of the line current from spacing to marking may be analyzed in a similar manner to show the cause of the gradual change in this case, as is shown in figure 11-40Y. The fact that the m-s and s-m waveshapes are identical in form is a valuable feature of polar operation. To obtain the full advantage of this feature, however, the relay operating points must be symmetrically located on the waveshape. That is, the s-m relay operating point must be located the same distance from the start of the s-m waveshape as the m-s relay operating point is located from the start of the m-s waveshape. These relay operating points will then be the same distance on each side of the zero current line of the waveshape diagrams. The s-mtd and the m-std are equal and there is no bias in the received signals. Unbiased polar transmission thus depends upon three conditions:

a. Equal but opposite potentials must be applied to the sending end.

b. The resistance of the circuit must

remain constant for both positions of the sending relay armature.

c. The operating points of the relay must be located symmetrically about the middle of the waveshape in order that equal transition delays will be secured.

11-332AI. Figure 11-40Z shows a situation where the steady-state marking and spacing currents of a polar circuit are not equal, with the marking current being +40 ma and the spacing current being −30 ma. This condition might be due to a difference in the ground potential between the terminals, or it might be due to an unbalance between the voltages on the contacts of the sending relay. In this case, the s-m transition starts when the line current is at −30 ma and ends when the current reaches the relay operating point, while the m-s transition starts when the line current is +40 ma and ends when the current reaches the other relay operating point. When the relay is properly adjusted, the relay operating points will be the same distance on each side of the zero current line in the waveshape. The total current change of the s-m transition is slightly over 30 ma while that of the m-s transition is slightly over 40 ma. As the rate of change (slope of curves) in the two directions is still the same, the delay to the m-s transition will obviously be greater than the delay to the s-m transition. The effect of this condition on transmission is that each mark, regardless of length, will be lengthened by an amount equal to the difference between the two transition delays, and each space, regardless of length, will be shortened by the same amount. If the bias condition of the circuit were reversed, which would be the case if the spacing currents were greater than the marking current, the delay to the s-m transitions would then be greater than the delay to the m-s transitions. Under this condition all marks would be shortened and all spaces would be

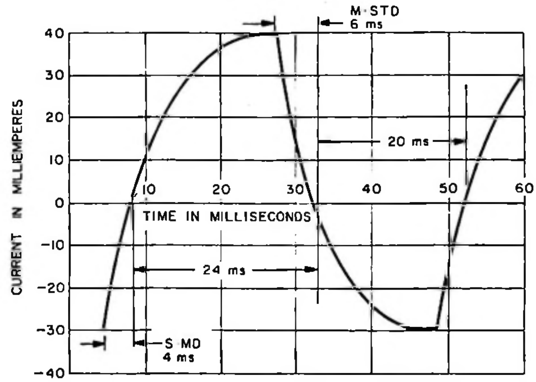


Figure 11-40Z. Effect on Signal Lengths of Unequal Polar Line Currents

lengthened and a spacing bias would exist. A similar situation could have existed if the steady-state current values had remained normal, and if the relay operating points had been shifted one way or the other on the waveshape. This could be caused by a biasing adjustment of the relay which, if it were marking, would cause the relay to operate to marking more easily than usual, and would thus shift the s-m operating point down on the waveshape. By the same token the relay would operate to spacing less readily, thus requiring more spacing current to operate it, and shifting the m-s operating point down on the wave also. This shifting of the operating points would once again make the transition from the marking condition to the m-s operating point on the waveshape different from the transition from the spacing condition to the s-m operating point on the waveshape. Unequal transmission delays and bias to transmission would result, just as in the previous instance. In either case, the important thing to note is, that although the m-s transition delays are different than the s-m transition delays, both sets of delays are constant in themselves. The difference between the two delays, which determines the amount of bias on the circuit, is therefore also a constant. Thus, if a circuit condition, like the one described, results in an m-std of 5 ms and an s-mtd of 3 ms, every m-s transition sent over the circuit will have a delay of 5 ms, and every s-m transition a delay of 3 ms regardless of the interval of time that may exist between transitions.

11-332AJ. CHARACTERISTIC DISTORTION. In the discussion so far, a transition has always been assumed to start when the line current was at the steady-state (full value) marking or spacing condition. There are situations, however, where the start of the transition does not occur when the line current is at its steady-state value. A definite amount of time is required for the line current to change from the steady-state marking condition to the steady-state spacing condition, and vice versa. Thus, in figure 11-40Y the time required for the current to make the complete change from marking to spacing and from spacing to marking is approximately 18 ms. On each transition in this case, the line current would have plenty of time to reach the steady-state value before the next transition occurred. The following transition would then start from the same current value as the preceding transitions, and the transition delay would be the same as the previous delays. In actual practice the time required to change from one steady-state condition to the other is sometimes greater than the minimum time interval between transitions in the signals. Some transitions then must occur while the line current is still in the process of changing from the previous transition. These transitions have a different delay time from transitions starting when the line current is in the steady-state condition and must therefore be distinguished from the latter type.

11-332AK. Part A of figure 11-40AA illustrates a case where the line currents require 33 ms to change from the steady-state spacing condition to the steady-state marking condition. Now assume that a marking impulse 22 ms long is being transmitted. The s-m transition at the start of the marking impulse occurs when the line current is in the steady-state spacing condition of −35 ma. This transition is thus a steady-state current transition, and as such will have the normal s-m transition delay, which is the same for all steady-state s-m transitions. The s-m transition at the beginning of the impulse starts the current changing towards the steady-state marking current value, an action which in this particular circuit will require 33 ms to complete. However, the m-s transition at the end of the marking impulse occurs only 22 ms later. At this time the line current, in the process of changing from −35 ma to +35 ma, has reached a value of 25 ma. The operation of the sending

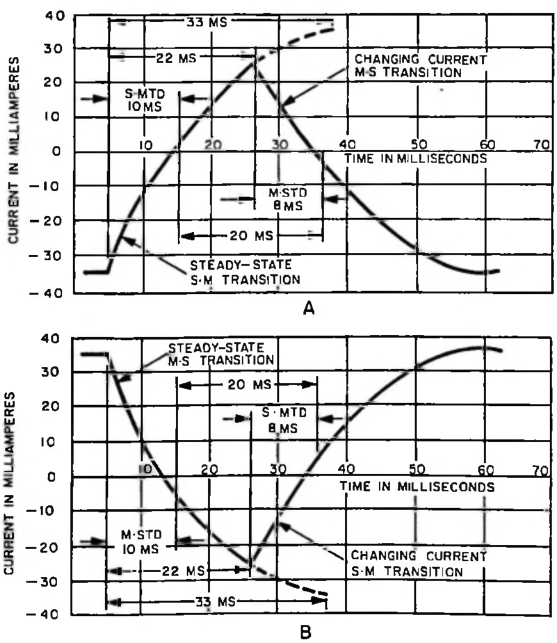


Figure 11-40AA. Changing Current Transitions relay at the end of the marking impulse reverses the voltage applied to the line, and the line current accordingly ceases changing toward the marking condition, and starts back toward the steady-state spacing condition. Since this m-s transition occurs when the line current is still in the process of changing, it is called a "changing current transition." When the line current reaches the value of −3 ma, the receiving relay operates to spacing, completing the m-s transition on the circuit.

11-332AL. The net effect on the marking impulse being transmitted in part A of figure 11-40AA will be to shorten it 2 ms. This shortening of the marking impulse occurs because the transition delay at the end of the impulse (8 ms), which adds to the impulse, is 2 ms less than the transition delay at the start of the impulse (10 ms), which subtracts from the beginning of the impulse. In a polar circuit, the rate of change of the current from spacing to marking is, of course, the same rate of change from marking to spacing. Accordingly, since in this particular circuit 33 ms were required for the current to change from −35 ma to +35 ma, 33 ms will also be required for the current to change from +35 ma to −35 ma. It also follows, then, that if a spacing impulse only 22 ms long is transmitted, the s-m transition at the end of the impulse will occur when the current is still in the changing condition and this transition will be a changing transition. Since the value of the current at the end of 22 ms was 25 ma where the current changed from spacing to marking, it follows that the current at the end of 22 ms will be −25 ma in this case. This condition is illustrated in part B of figure 11-40AA. The total current change involved in the s-m transition at the end of the spacing impulse will then be from −25 ma to +3 ma or 28 ma, the same as the total current change that took place in the former case. Likewise, the delay to this changing current transition will be 8 ms and the marking impulse being transmitted will then be reduced 2 ms in length.

11-332AM. The magnitude of the changing current transition delays just discussed is proportional to the time required for the current to change from its value at the start of the transition to the operating point value of the receiving relay. In both parts of the figure the current change was from 25 ma to 3 ma of the opposite sign, or a total change of 28 ma. It is obvious, however, from an inspection of these figures, that if the impulse transmitted had been longer than 22 ms, the line current would have been at a higher value at the time of the transition at the end of the impulse, and the transition delay would have been greater. The limiting delay will, of course, be the steady-state delay. Also if the impulse transmitted had been less than 22 ms in length, the line current would have been at a lower value at the time of the transition at the end of the impulse, and the transition delay would accordingly have been less. This is illustrated by figure 11-40AB which shows waveshapes of marking impulses from the three standard telegraph printer speeds in a circuit where the time required for the line current to change from its negative to positive value, and vice versa,

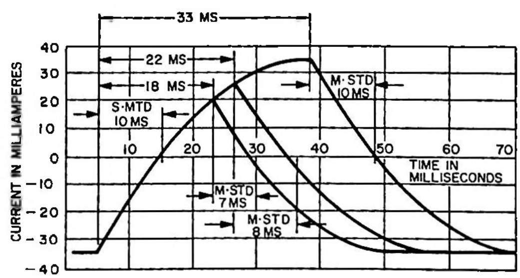


Figure 11-40AB. Characteristic Distortion Effect on Signal Lengths at Baud Operation is 33 ms. The marking impulses illustrated are 18 ms long, corresponding to 55.5 bauds; 22 ms long corresponding to 45.5 bauds; and 33 ms long corresponding to 33.3-baud modulation rates. Waveshapes for the spacing signals would, of course, be identical except for reversal of the current values. In the case of the 33-ms marking impulse, the impulse is just the required length for the current to change from one steady-state condition to the other, and the transition at the end of the impulse is thus a steady-state transition. In the case of the 22-ms impulse, the s-m transition at the end occurs when the line current is at +25 ma value, and this transition is thus a changing current transition starting at a current value less than the steady-state value.

11-332AN. As we noted before, the delay is accordingly less than the delay to the steady-state transition, 8 ms as compared with 10 ms. In the case of the 18 ms impulse, the s-m transition occurs when the line current is only at a +18 ma value. This transition is thus also a changing current transition. Due to the fact that the line current changes only 21 ma to reach the m-s operating point of the relay, as compared to the change of 28 ma for the m-s transition of the 22-ms impulse, the delay is still less. As indicated in figure 11-40AB it is now only 7 ms. The amount of a changing current transition delay is thus dependent upon the value of the line current at the start of the transition. The value of the line current is dependent upon the time interval between the changing current transition under discussion and the previous transition, which started the line current to changing. Since the time interval between the beginning of these two transitions is equal to the length of the sent impulse, it is this impulse length which finally determines the transition delay under a given set of conditions.

11-332AO. In the conditions just described, the lengths of the received signal impulses are obviously affected by the presence of the changing current transitions. This effect is called characteristic distortion. The magnitude of the effect is inversely proportional to the length of the sent impulses, and the nature of the effect is to shorten received short impulses. Since the received impulses under consideration are shortened, the effect of this case is called negative characteristic distortion. An opposite effect is possible. The characteristics of a circuit may be such that the line current tends to increase momentarily at the completion of each transition to a value greater than the steady state, due to transient effects. If the next transition occurs at such an instant, the transition delay will be greater than the delay on the preceding transition, which means that the length of the received mark or space signal, as the case may be, will be lengthened. Since the signal impulse is lengthened, this is called positive characteristic distortion. However, as the transient effect causes the line current to oscillate (increase to decrease) around the steady state, it is possible that the next transition might occur at the instant that the line current had momentarily decreased below the steady state. In such a case, the transition delay would be less, which would result in negative characteristic distortion. Thus, transient conditions may cause either positive or negative characteristic distortion, but positive characteristic distortion is not so frequently encountered as negative characteristic distortion.

11-332AP. To summarize, the change of the line current from one condition to the other on a telegraph circuit requires a definite time to be completed. If the time interval between the transitions of the signals at the sending end of the circuit is less than the time required for the line current to complete its change, changing current transitions will occur. These transitions will have delays either greater or less than the normal steady-state transition delays of the circuit, and will lengthen or shorten the short impulses of the signals to an amount dependent upon the value of the changing current transition delay which, in turn, is determined by the length of the impulse that caused the changing current transition. If the effect is to shorten the short impulse, it is negative characteristic distortion. If the effect is to lengthen the short impulse, it is positive characteristic distortion.

11-332AQ. The contrasts between characteristic distortion and bias are as follows:

a. The effect of characteristic distortion depends upon the length of the impulses transmitted. The effect of bias is independent of the length of the impulses.

b. For a given length of impulse, the effect of characteristic distortion is independent of whether it is a marking or a spacing impulse. The effect of bias is always opposite on a mark to what it is on a space.

c. Characteristic distortion is related to the amount and the arrangement of the capacitance, the inductance, and the resistance of a circuit. Except in neutral operation, these factors do not affect bias.

d. Bias is caused by unequal marking and spacing line current, biased relays, and other conditions, all of which do not effect characteristic distortion.

e. Characteristic distortion, because it is due to the capacitance, inductance, and resistance of a circuit, which, except for the resistance, are unchanging in value, varies only a small amount from day to day on a circuit. Bias, because it is caused by unbalanced voltages, ground potential, relays losing adjustment, and other similar faults, may vary from hour to hour on a circuit.

11-332AR. FORTUITOUS DISTORTION. The form of distortion, caused by such factors as cross-fire, power induction, momentary battery fluctuations, "hits," break key operation, etc, and which displaces miscellaneous received transitions by various amounts intermittently, is known as fortuitous distortion. At times this effect may be large enough to produce a complete failure of the circuit. In the transmission of miscellaneous signals, the combined effect of characteristic and fortuitous distortion on the displacement of received transitions is sometimes known as "jitter".

11-332AS. END DISTORTION. End distortion is the type of distortion in which the marking pulse is either lengthened or shortened at the end of each marking pulse with respect to the beginning of a start pulse. If the mark-to-space transition is delayed, the effect is to lengthen the marking code pulses and to shorten all spacing code pulses. This type of distortion is called marking end distortion. If the mark-to-space transition is advanced, the effect is to shorten the marking code pulses and to lengthen all spacing code pulses. This is called spacing end distortion. Marking and spacing end distorted signals are illustrated in figure 11-40AC. Note that the rest and the start pulses for end distortion are of the same duration as those for undistorted signals. This is not a condition which is normally encountered in telegraph circuits since inductance and capacitance will affect all of the signals in the train. The transmission of end distortion for test purposes makes it possible to determine the tolerance of a receiving mechanism to random or fortuitous distortions which do not affect all transitions alike. End distortion is a form of distortion that does not usually occur because of line

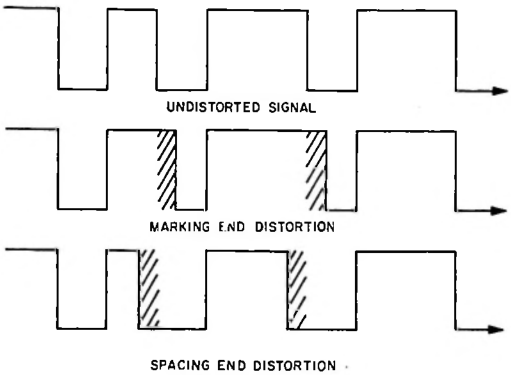


Figure 11-40AC. End Distortion Effect

conditions or characteristics. End distortion affects the length of the marking signals in much the same way as bias, but instead of affecting the front of the signal, it adds to or subtracts from the rear portion of the marking impulse.

11-332AT. The receiving mechanism of a teletypewriter requires about 20 percent of a signal for correct operation. When the receiving mechanism is oriented correctly, the 20 percent will be taken from the center of each impulse, thereby allowing the selecting interval to be shifted an equal amount in either direction. That is, the range finger may be moved toward the high and low ends by the same amount without causing errors in the printed copy. When end distortion is present in the received signals, the start and stop signals are not distorted. Therefore, any increase or decrease in length of the received marking impulse will be at the rear of that signal. This will affect the range of the receiving apparatus in just the opposite manner to the effect caused by bias distortion.

11-332AU. Figure 11-40AD is a comparison of the impulses of character Y as received with zero distortion, 40 percent marking end distortion, and 40 percent spacing end distortion. The selecting interval of each impulse is also shown in the figure. The selecting interval of A in figure 11-40AD

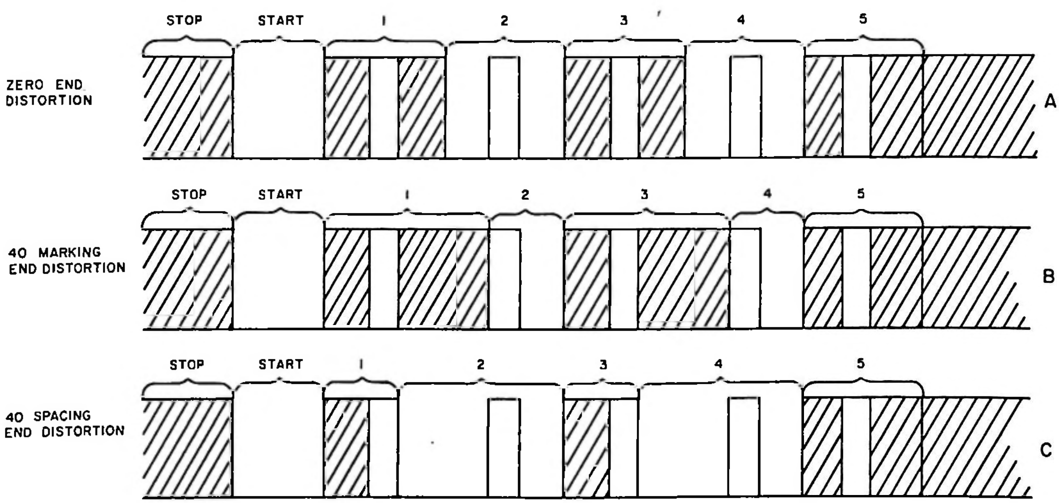


Figure 11-40AD. Comparison of Received Signals With Zero and 40 Percent End Distortion (zero distortion) may be moved 40 percent in either direction without the selection of a wrong impulse. The selection interval of B in figure 11-40AD (marking end distortion) cannot be moved to the left (low side of the range finder scale) without the selection of a marking impulse where space is required for correct reproduction of the character. The selecting interval may be moved to the right (high side of the range finder scale) by 40 percent with no effect on the selection of each impulse. The effect of spacing end distortion (C of figure 11-40AD is just the opposite. That is, the selecting interval may be moved to the left by 40 percent without affecting the receiving unit, but may not be moved to the right and still have the correct character reproduced. Marking end distortion therefore causes the range of the receiving unit to be decreased by increasing the low side, and spacing end distortion decreases the range by decreasing the high side.

11-332AV. EFFECTS OF SELECTOR MAGNET. Fixed-plant teletypewriter equipment normally possesses a relay on its mounting base. Too often, the explanation given for the presence of the line relay is "for operating the receiving equipment over long distance circuit." It is to be emphasized that while this is certainly true, there is a more valid and fundamental reason for the presence of the relay. The reason for the inclusion of the line relay is to isolate the effect of the individual teletypewriter, especially its selector magnets, from the working line circuit. Essentially, it can be demonstrated that the effect of line relay removal has more undesirable side effects than an occasional relay maintenance problem. The line relay should be bypassed only under specific circumstances, and in any case, one must accept the derogatory effects produced in external line circuits.

11-332AW. When the selector magnet of a teletypewriter is inserted into a neutral line circuit, a large inductance is inserted into the circuit. This comparatively large inductance component itself has a serious enough effect on the electrical signal, but even worse is the variable inductance reflected into the circuit from the motion of the selector armature as it moves toward and away from the pole faces of the selector magnet, as is illustrated in figure 11-40AE. The transient effect noted on the space-to-mark transition is the back emf produced when the selector armature is placed adjacent to the selector magnet pole pieces (core). The relative position (phase) of this transient may be moved earlier or later by adjusting the rangefinder arm of the teletypewriter.

11-332AX. Figure 11-40AF shows the resultant input signal to relay Z in figure 11-40AG, for a 45.5-baud, unit stop, locked letter Y. Part A of the figure is a polar trace; relay output to the line is 6 percent spacing bias. Part B of the figure is a neutral waveform; input to the line relay is 1 percent distortion from the keying device. Note the effect of the monitor printer selector magnet in series with the input to the relay on the mark-to-space and space-to-mark transitions. Figure 11-40AH is made under the same conditions as before except that the rangefinder arm has been moved to produce selection at a later point in time. Notice that the relay transit time has slowed down in the space-to-space transitions. Relay output to the line is now increased to 15 percent spacing bias. Figure 11-40AG

Figure 11-40AE. Waveform Distortion Resulting From Selector Magnet in Line

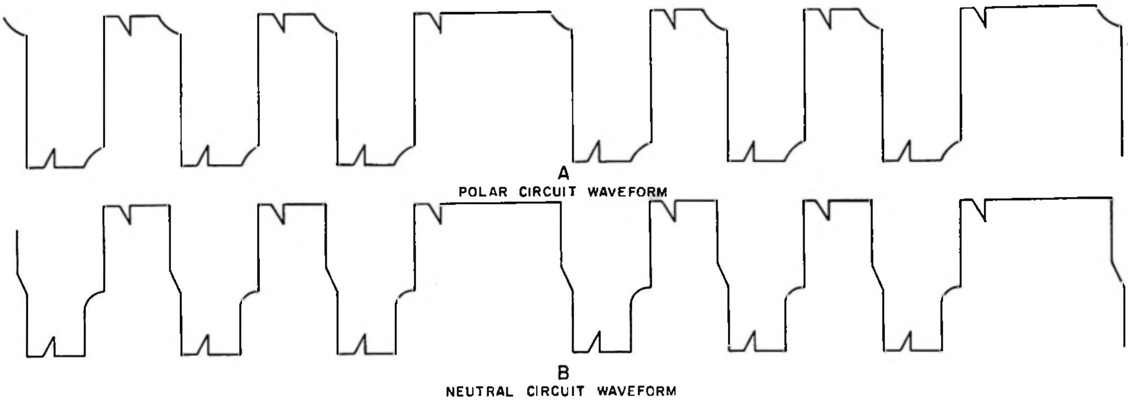


Figure 11-40AF. Waveforms Associated With Figure 11-40AG

shows the same circuit conditions except that the monitor printer has been removed from the input to the relay. Notice that the relay transit times are now equal and much faster than the space-to-mark transit times of either figure previously shown, as is shown in figure 11-40AI. Output distortion of relay to the line is now 2 percent. The serious effect of the movement of the range-finder arm of the receiving teletypewriter is thus incontestably displayed. It should be remembered that if the varying inductive effect is moved in time away from the leading edge of the waveform it will have less effect (8 percent versus 15 to 20 percent space bias or characteristic distortion). The changing effect with range scale setting is due to the magnet flux built up in the relay signal winding. This magnetic flux is not disturbed significantly until the signal flux build up has overcome the bias winding flux, and the armature has completed its travel. The transient effect of the elector armature has less chance of cancelling the flux of the signal winding and while it lowers the flux in the signal winding it does not normally

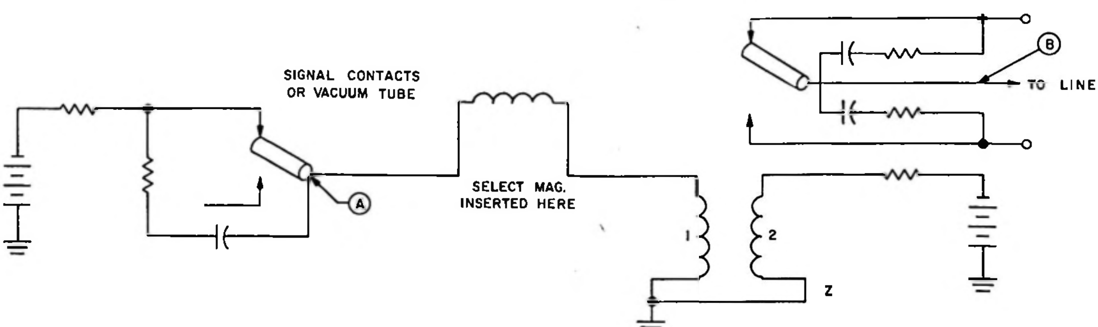


Figure 11-40AG. Typical Neutral Telegraph Circuit

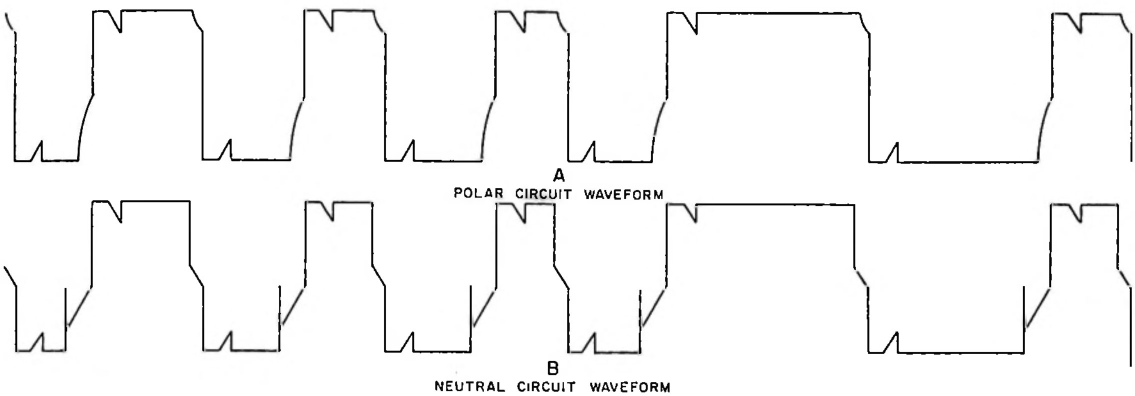


Figure 11-40AH. Waveforms Associated With Figure 11-40AG

cause the relay armature to move again.

11-332AY. There is still another reason why selector magnets should not be inserted directly into a line circuit. As mentioned previously, the polar relay is designed to operate at and about the half-power point, while the selector magnet can respond to signals considerably below this level. That is, the selector magnet armature can follow signals which decrease in amplitude yet do not necessarily pass through the half-power point. This facet of operation causes the teletypewriter to be more susceptible to random noises and circuit disturbances.

11-332AZ. INTERNAL DISTORTION. Assuming that a rheostat and milliammeter are connected in series with basic elements of a teletypewriter, a study of energizing currents can be made. Also, under the assumption that all mechanical adjustments

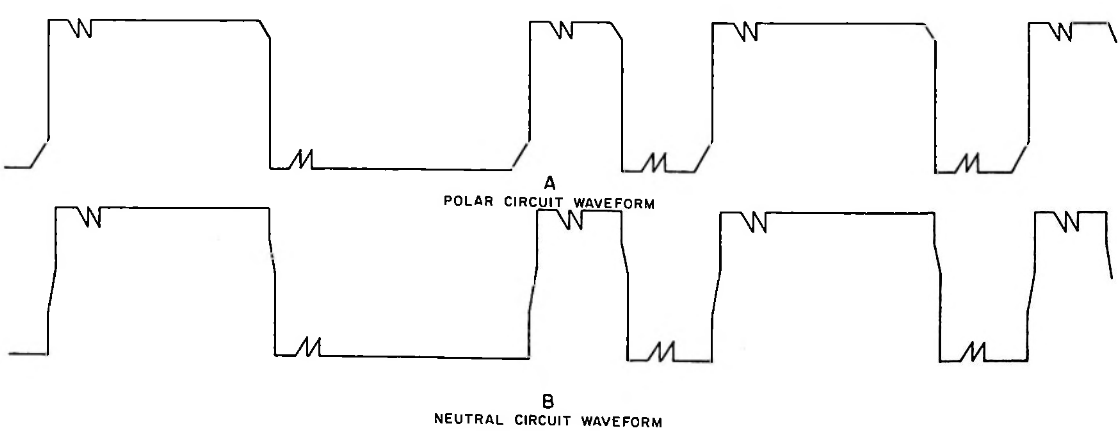


Figure 11-40AI. Waveforms Associated With Figure 11-40AG, Selector Magnets Removed From Circuit at optimum value, the rheostat should be first adjusted so that the selector magnet armature is in the spacing position. As the rheostat is carefully adjusted, the circuit current will increase steadily; there will be a definite milliammeter indication at which the selector armature operates to the marking position. This indication will be approximately 36 ma and is called the operating current value of the unit for a particular set of mechanical adjustments. As the rheostat is once again adjusted for decreasing values of circuit current, there will be another definite milliammeter indication at which the selector magnet armature operates to the spacing position. This indication is called the release current value for the same set of mechanical adjustments as before.

11-332BA. It is useful to use the selector magnet unit in a teletypewriter circuit without any changes in mechanical or electrical parameters from the previous discussion. The first two impulses of a character are shown in figure 11-40AJ; the transmitted pulses are shown in idealized form by means of dashed lines. The received pulses are shown superimposed on the same coordinates. The first impulse of the character is the start pulse beginning with the marking-to-space transition at time t_0. When the current through the selector magnet coils decreases to its release value at approximately 24 ma, the selector magnet armature moves to the spacing position. The difference in time of transmitting contact opening at time t_1 and the time of armature movement to the spacing position at time t_2 is slightly greater than 1 ms in the figure. At time t_3, which is 22 ms later than time t_1, the transmitter contact closes, terminating the start pulse interval and beginning the first intelligence impulse in the marking direction. The line current must increase to its operate value of approximately 36 ma before the selector magnet armature can be attracted to the marking position once again. Notice that the difference in time of transmitting contact closure at time t_3 and time of selector armature movement to the marking position is slightly greater than 1 ms once again. It is now apparent that the received impulses for this reference condition are 22 ms long as are the transmitted pulses,

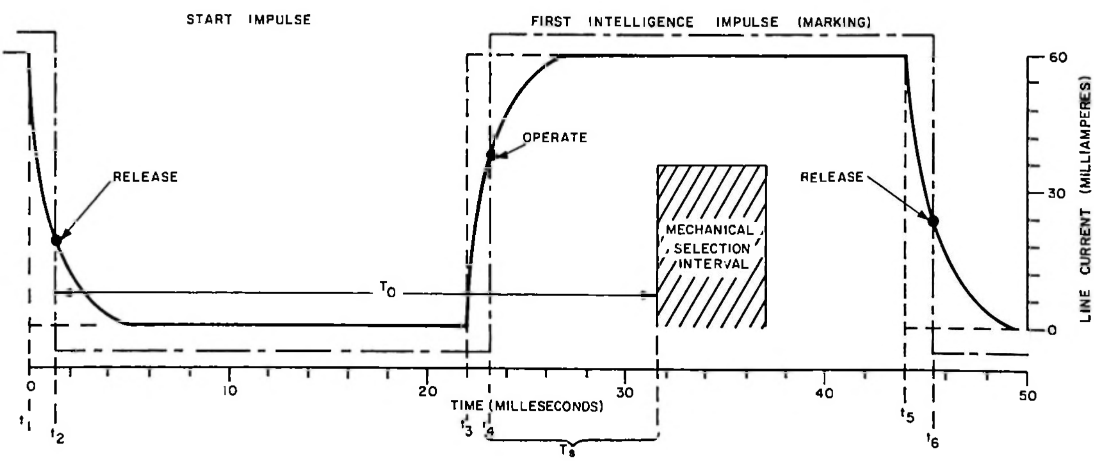


Figure 11-40AJ. Selector Magnet Operation Versus Electrical Signal and Mechanical Selection—Zero Bias however, the received signal is delayed by 1 ms, as is shown by the dot-dash lines.

11-332BB. The release of the selector armature at time t_2 begins a series of mechanical operations which results in a mechanical selection interval 30.5 ms later than time t_2; it is designated as interval T_o in the figure. The interval T_o is independent of the electrical signal once the selector armature is released on the mark-to-space transition at time t_2. Inspection of the beginning of the mechanical selection interval with respect to the operate point of the selector armature for the first intelligence impulse shows that mechanical selection has taken place 8.5 ms later than time t_4 indicated as interval T_s in the figure. Furthermore, inspection of the interval from the end of the selection time to the end of the first intelligence impulse shows a difference of 8.5 ms. For the conditions shown in the figure, the orientation range is adjusted so that mechanical selection will occur exactly in the middle of the time allotted for the reception of the first intelligence impulse and all remaining intelligence impulses, which is the optimum condition.

11-332BC. Consider now that the tension of the selector armature spring is increased; a new set of conditions will apply even for the same electrical signal as before, which is indicated in figure 11-40AK. When the mark-to-space transition is transmitted into the line, selector magnet current will decrease to the release value to permit selector armature movement to the spacing position to commence the received start impulse interval. Notice that due to the increased armature spring tension, the selector armature moves to the spacing position for a coil current of approximately 36 ma. This value is 12 ma greater than for the previous conditions. The corresponding time of selector armature release at time t_2 is reduced to about 0.5 ms; hence, increasing armature spring tension will cause the selector armature be released earlier than it would for optimum conditions. At time t_3, which is 22 ms later than time t_1, the transmitter contact closes, terminating the start pulse interval and commencing the first intelligence impulse in the marking direction. The

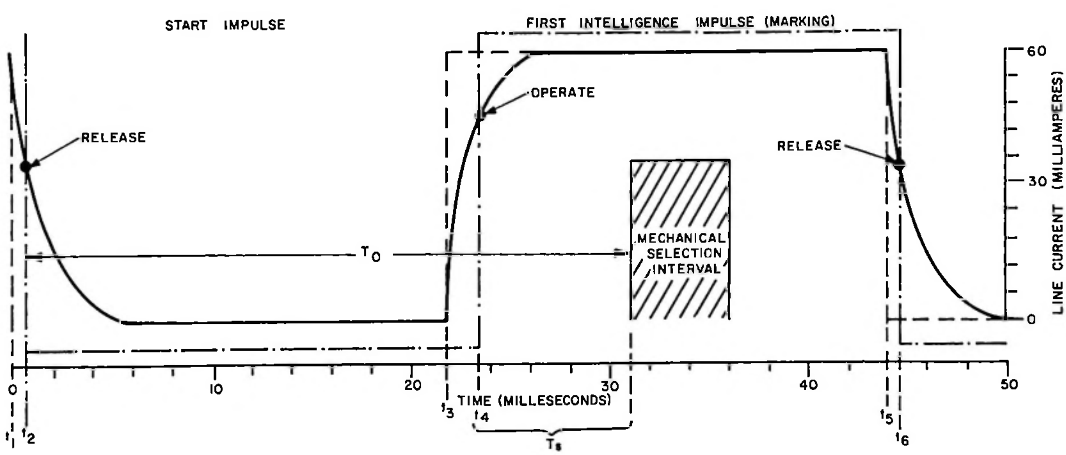


Figure 11-40AK. Selector Magnet Operation Versus Electrical Signal and Mechanical Selection — Spacing Bias line current must increase to about 45 ma to produce enough magnetic flux to overcome the increased armature spring tension. The corresponding time difference between transmitting contact closure at time t_3 and selector magnet armature movement to the marking condition at time t_4 is about 1.5 ms in this case. The figure indicates that although the transmitted start impulse occupied the full 22 ms time allotted from time t_1 to t_3, the time of mechanical operation occupied a 23-ms interval from time t_2 to t_4, an increase of 1 ms over and above the transmitted impulse. At time t_5, which is 44 ms later than t_1, the transmitting contact opens to terminate the first intelligence impulse and initiate the second intelligence impulse. The line current will decrease once again to the release current value and the selector armature will move to the spacing position. The time of release, which is the interval t_5 to t_6, will be equal to the interval of t_1 and t_2; release occurs earlier than optimum conditions. The figure indicates that although the time of transmission for the first intelligence impulse lasted for the scheduled 22 ms from t_3 to t_5, the time of mechanical operation occupied a 21-ms interval from time t_4 to t_6 a decrease of 1 ms under the transmitted impulse.

11-332BD. The spacing impulses have been effectively increased in length while the marking impulses have been shortened in length, by means of increasing selector armature spring tension; this condition is termed internal spacing bias. The term spacing bias is used because the length of the spacing impulses is greater than the length of the marking impulses; the term internal is used because the bias condition is caused by a mechanical adjustment rather than an electrical signal defect. Now that the conditions for spacing bias have been established, a determination should be made with respect to the orientation range under increased armature spring tension. Assume that the orientation range remains unchanged from its previous setting. The release of the selector armature at time t_2 begins a series of mechanical operations which results in a mechanical selection interval 30.5 ms later than time t_2, which is designated as interval T_o in the figure. Inspection of the beginning of the mechanical selection interval, with respect to the operate point of the selector armature for the first intelligence impulse, shows that mechanical selection began 7.5 ms later than time t_4, which is indicated as interval T_s in the figure. Mechanical selection no longer occurs in the exact center of the intelligence bit. A comparison with the conditions shown in the previous figure 11-40AJ shows that mechanical selection occurred 1 ms earlier than it should have for optimum conditions. Whenever selector-armature spring tension is increased, the increase is equivalent to advancing the mechanical selection interval. In terms of the rangefinder mechanism, it is equivalent to moving the rangefinder arm toward the lower numbers of the scale. Compensation for a limited amount of internal spacing bias can be obtained by adjusting the rangefinder arm towards the higher numbers of the scale.

11-332BE. All of the arguments for producing internal spacing bias have been predicated on increasing the tension of the armature spring, but this is not the only mechanical means of producing the same result. For instance, the magnet pole faces can be moved farther away from the armature instead of increasing spring tension. The magnetic field exerted by the coil toward the armature is considerably weakened and spring tension will again be predominant to produce spacing bias. To counteract this effect, armature spring tension must be reduced, or the coil current increased, or the rangefinder arm must be moved toward the higher end of the scale, or any combination of these. While it is certainly true that the spacing between the selector armature and the magnet pole faces could be decreased by a readjustment of the lower selector armature stop in an effort to correct spacing bias, it is generally not advisable since other interoperating adjustments could interfere to the extent of producing complete operating failure with very small amounts of armature stop adjustment. In electrical terms, spacing bias can also be produced by lowering the line current through the selector magnet coils.

11-332BF. Suppose that the selector armature spring tension is reduced. It should be quite apparent to you that the selector armature will be held in the marking position until relatively low values of line current are attained. In a typical situation, assume that 14 ma of line current are being used (figure 11-40AJ). The start of the selector mechanism will be delayed for approximately 2 ms. The operate point will be moved into the region of 25 ma instead of the reference 36 ma, and the corresponding time between t_3 and t_4 is reduced considerably to about 0.5 ms. The start pulse has been effectively shortened and will therefore produce a corresponding increase in marking impulse operating time. This condition where the marking impulses are longer than the spacing intervals is termed marking bias, further, since the bias condition was precipitated by a mechanical adjustment, it is termed internal marking bias. Introducing friction within the selector mechanism will therefore produce marking bias. The beginning of the mechanical selection interval will now be retarded with respect to the leading edge of the marking impulses. To counteract this condition without changing spring tension, you must move the range-finder arm toward the lower numbers in its scale. Since moving the magnet pole faces away from the armature produces spacing bias, marking bias can be overcome by increasing the distance between magnet pole faces and the armature.

11-332BG. When a teletypewriter is misadjusted, it is subject to internal distortions which reduce the operating margin of the selector mechanism. After all other adjustments in a teleprinter are properly made, the proper balance between the selector armature spring and the magnet air gap must be established for the line current on which the teleprinter will be operating. Since the spring and air gap adjustments given in the adjustment specifications of the teleprinter were developed by taking the average of a large number of teleprinters, they do not necessarily result in optimum adjustments for each teleprinter due to variations between teleprinters. If the magnet air gap is not properly adjusted so that the pull of the magnet is in proper balance with the armature spring tension, internal bias will result. If the armature spring load is not of the magnitude, erratic operation of the selector armature will occur which will result in internal skew. All teleprinters should be so adjusted that the internal bias and skew are below 3 percent.

11-332BH. When initially adjusting a teletypewriter for elimination of bias and skew, the armature spring tension and air gap are adjusted in accordance with the specifications for the teletypewriter. The teletypewriter and a bias/distortion test set are then connected into a local test circuit which has the same line current as in the actual operating condition of the teletypewriter. Since wave-shaped signals affect the internal bias of the teleprinter, no wave shaping should be used in the test circuit. A prepared test tape is then inserted in the test set and a fixed amount of spacing bias is transmitted to the teletypewriter and a range taken. The same amount of marking bias is then transmitted and the range again measured. The internal bias and skew are then calculated using the following relationships.

$$\text{Bias} = \frac{(H_s - L_s) - (H_m - L_m)}{2}$$

$$\text{Skew} = \frac{(H_m - L_s + 2D) - (H_s - L_m)}{2}$$

Where H_m = high end of marking bias range

L_m = low end of marking bias range

H_s = high end of spacing bias range

L_s = low end of spacing bias range

D = percentage of distortion in transmitted signals

Positive bias indicates internal marking bias and negative bias indicates internal spacing bias. Positive or negative skew in a teleprinter is the result of irregular operation of the selector armature, or some other selector member, from the marking to the spacing position or from the spacing to the marking position, respectively.

11-332BI. If a teleprinter has internal skew greater than plus or minus 3 percent, the skew should be removed first by adjusting the armature spring. Positive skew can be reduced by increasing the armature spring tension and negative skew can be reduced by decreasing the spring tension. After the internal skew has been reduced to 3 percent or less, the internal bias should be reduced to 3 percent or less by adjusting the armature air gap. Negative bias can be reduced by decreasing the air gap and positive bias can be reduced by increasing the air gap. The range scale should then be set at the center of the bias range. This center is found by the following formula:

$$\text{bias center} = \frac{H_m + L_s}{2}$$

With the range scale set at this point, a teleprinter operates when signals containing 40 percent marking or spacing bias are transmitted to it. It should also print correctly on signals containing 35 percent marking or spacing end distortion. A teletypewriter operating at a 75-baud modulation rate should operate satisfactorily on signals containing 35 percent marking or spacing bias and 30 percent marking or spacing end distortion. A teletypewriter receiver operating at a 50-baud modulation rate, and adjusted in accordance with the foregoing method will usually have a range of at least 80 points on undistorted signals. Ranges of 85 to 90 points are not uncommon. However, ability to operate satisfactorily on distorted signals is a far more important test of a receiver than is its range on undistorted signals.

11-332BJ. TELETYPEWRITER MARGIN.

11-332BK. Teletypewriter margin is usually defined as the maximum distortion of the incoming modulation for which the teletypewriter produces a series of errors less than some arbitrary figure; it is affected by the kind of text transmitted and the rhythm of transmission among other parameters already discussed. Such a definition is rather incomplete and much experimental work is being carried out in order to develope a better definition. There are generally four margin figures for each receiver. Two of these margins must distinguish between mark-to-space transitions and space-to-mark transitions so that allowances can be made for bias in the receiving relay. The two remaining margin figures are relative to modulations having early or late distortions to advance-distorted or delay-distorted modulation. All four receiver margin figures fluctuate around a mean value because of friction, inertia, and mechanical depreciation.

11-332BL. Provided that the selector armature is in the correct spacing or marking position during the middle quarter of each intelligence code unit, the selecting mechanism should correctly translate the electrical signals into properly recorded characters. This corresponds to a maximum distortion of 7.5 ms, either early or late. Since it is common to express distortion as a percentage of the signal unit, and since the particular modulation rate is 50 baud:

$$\frac{7.5 \text{ ms}}{20 \text{ ms}} \times 100 = 37.5 \text{ percent}$$

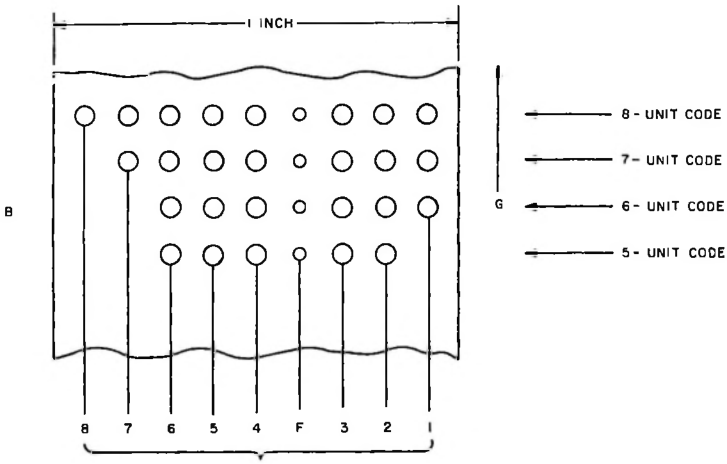
A well adjusted receiver, able to work with this amount of distortion is said to have a margin of 37.5 percent; in practice, a margin of 35 percent is considered satisfactory.

11-332BM. TAPE-RECEIVED EQUIPMENT.

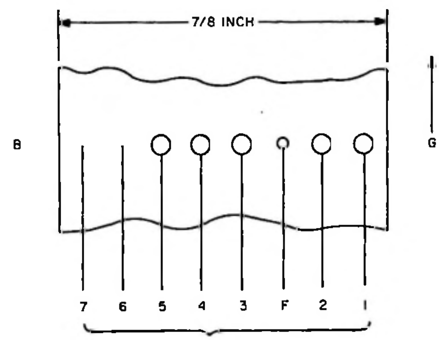
11-332BN. Teletypewriter messages can also be received as perforations in a paper tape. The information recorded on the tape contains intelligence data, however, synchronization date is not included; in addition to the perforated information a smaller hole is punched in the tape for feeding the tape through the device. The standard for identifying information tracks is established in two parts. The reason the two part standard exists is because of the adoption of an 8-unit code; the 5-unit code is now an interim standard. For machine systems with the capability of using 8 information tracks, the feed hole track shall be situated between the third and fourth information tracks from the guiding edge of the tape as illustrated in figure 11-40AL. For machine systems with the capability of using 5 information tracks (this includes teletypewriters), the feed hole shall be situated between the second and third information tracks from the guiding edge of the tape as illustrated in figure 11-40AL. Since the latter standard is an interim standard, it is not to be used in future design.

11-332BO. Tape width has evolved from a standard of 11/16 inch to 1 inch. Messages recorded by a typing reperforator appear both in typewritten characters and in code perforations. On the 11/16-inch wide tape, the typewritten characters and code perforations occupy the same tape area; complete perforation of the tape would remove part of the printing and impair the readability of the typewritten message. Therefore, a chadless form of perforation is used. The punchings, or chads, are not completely severed from the tape; about 75 percent of the circumference of the chad is severed, the rest remaining attached and forming a lid for the perforation cut in the tape. At a later point in time, the tape width was changed to 7/8 inch. The additional area provided room for two printing tracks at the bottom of the tape, one track for alphabet-letter printing and one track for numerical-figure printing, which were free of code perforations. The 1-inch wide tape is capable of providing the extra area for perforating tracks 6, 7, and 8. A minimum tape width of 1 inch is established for systems with the capability of using three information tracks between the guiding edge and the feed holes. Minimum tape widths of 11/16 inch and 7/8 inch are established for systems with the capability of employing two information tracks between the guiding edge and the feed hole.

11-332BP. While chadless tape has solved the problem of legibility it has not permitted the perforated tape to occupy the same volume as the unperforated tape. If tight reeling were attempted, it would be very likely that the adjacent lids would interlock with the perforation, causing retransmission problems. The standard is now chadded tape, that is, the perforations are completely punched.



THREE INFORMATION TRACKS - G TO F
TAPE AS VIEWED FROM TOP PRINTING SIDE



TWO INFORMATION TRACKS - G TO F
TAPE AS VIEWED FROM TOP PRINTING SIDE

LEGEND

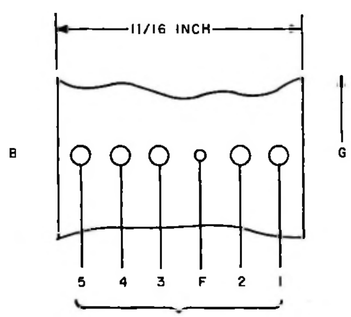
↑ DIRECTION OF TAPE FEED

G GUIDING EDGE OF TAPE

1-8 INFORMATION TRACKS START FROM TAPE GUIDE

F FEED HOLE

B BOTTOM EDGE OF TAPE



TWO INFORMATION TRACKS - G TO F
TAPE AS VIEWED FROM TOP PRINTING SIDE

Figure 11-40AL. Perforated Paper Tape Standards

11-332BQ. PAGE PRINTING MECHANISM.

11-332BR. Teletypewriter messages can be received by teletypewriter equipment containing page printing mechanisms. In general, most types of page printing mechanisms vary little in design. There are minor differences that exist in page printing mechanisms that are used in teletypewriters designed for use as fixed station equipment and teletypewriters designed for use as field equipment, but the basic principles of operation for all printing mechanisms are the same.

11-332BS. The theory of selecting and printing on both fixed plant and field teletypewriters is basically the same. In the design of fixed plant and field teletypewriters, every effort is made to make parts interchangeable with similar parts in other teletypewriter mechanisms. Minor electrical and mechanical differences are inevitable, due to the fact that the basic page printing mechanism is used as a component of a variety of field and fired plant sets.

11-332BT. Page printing teletypewriters are capable of sending and receiving messages in standard start-stop, five unit code. Teletypewriters generally print on neutral signals, but teletypewriters can be adopted to receive incoming polar signals. This adoption is accomplished with the use of certain wiring changes. In some instances involving field equipment, the adoptation is accomplished through additional equipment and adjustments.

11-332BU. Operation of the page printer is controlled by the teletypewriter code groups. These teletypewriter code groups are received from either the keyboard-transmitter of the same teletypewriter, or from the transmitter of another teletypewriter. When the code groups are received from the transmitter of another teletypewriter, both teletypewriters are connected to the same signal circuit.

11-332BV. The start impulse (first impulse of each code group) causes the page printer to start a cycle of mechanical operation. The five code impulses, received in sequence in the selector magnet of the page printer, are used by the selector mechanisms to select the character to be printed, or to select the nonprinting mechanical operation that is to be performed. After the fifth code impulse is received, the selected character is printed, or the selected mechanical operation is performed automatically by the page printer. The stop impulse is used to stop the page printer until the start impulse of the next code group is received.

11-333. RADIO TELETYPEWRITER EQUIPMENT.

11-334. GENERAL. Radio teletypewriter signals are usually transmitted by either the tone-modulated method or the carrier-frequency-shift method. The tone-modulated method is normally used for short-distance communications, and the carrier-frequency-shift method is used for long-distance communications.

11-335. TONE-MODULATED METHOD. Figure 11-41 shows the basic equipment required for a tone-modulation radio teletypewriter terminal. The tone converter for this terminal is used for both transmitting and receiving. The switching control unit associated with the teletypewriter has a single "transmit-receive" position for tone-modulated operations.

11-336. A start impulse from the local teletypewriter instantly and automatically switches the tone converter to the transmit condition. While the tone converter is in the transmit condition, signals from the receiver are blocked. This equipment cannot

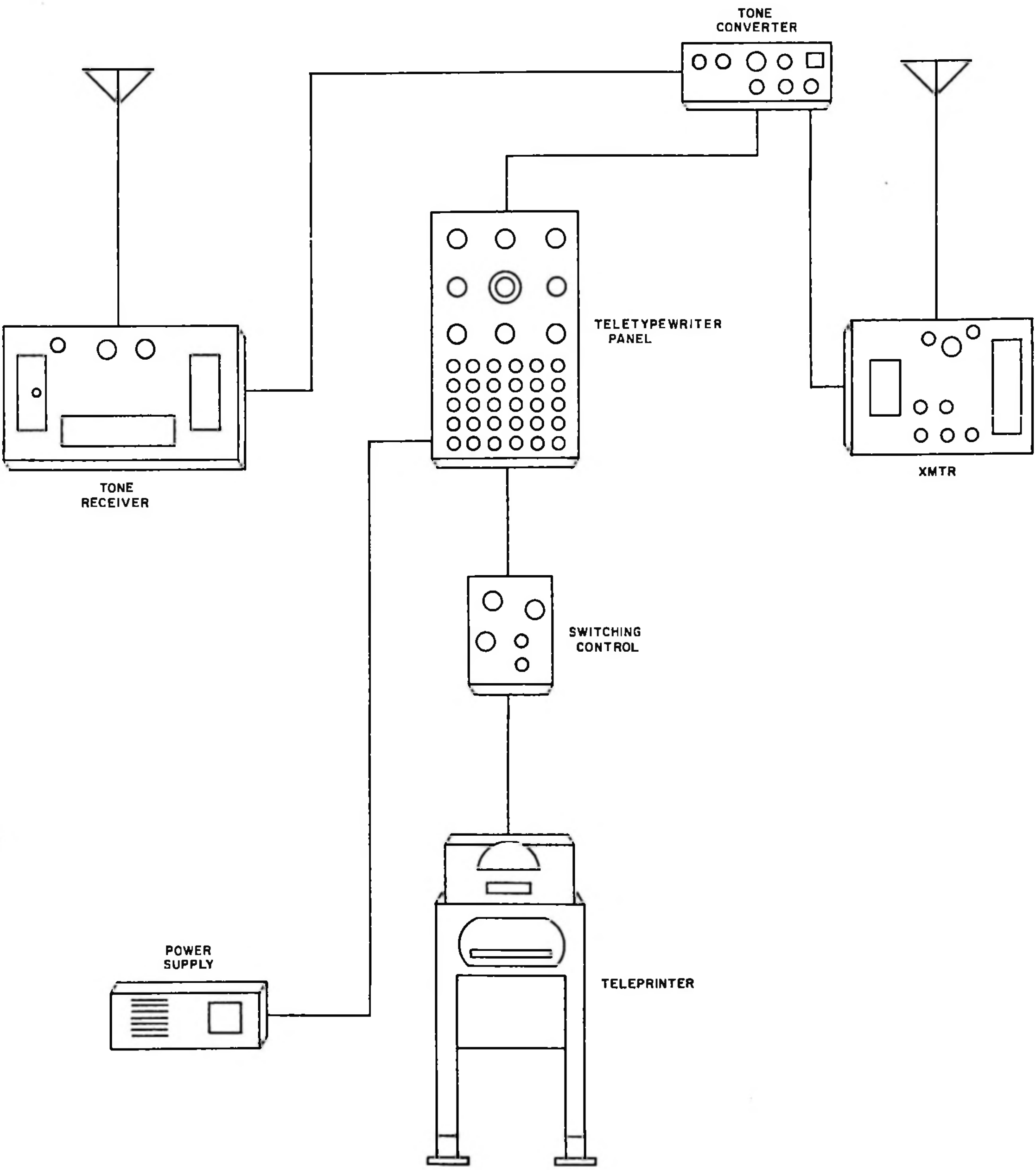


Figure 11-41. Tone-Modulation Radio Teletypewriter Terminal

transmit and receive simultaneously. When transmission stops, there is a slight delay before the converter automatically reverts to standby "receive" condition. This delay compensates for short interruptions in the transmitting message. The equipment will then remain ready to receive until local transmission is resumed.

11-337. The two basic circuits in the tone-modulation radio teletypewriter terminal are the transmit and receive circuits. A circuit common to both systems is the direct-current power supply that furnishes the local "loop" current to the teletypewriter panel. The transmit circuit passes from the teletypewriter through the tone-modulated transmit-receive position of the switching control unit, through a channel in the teletypewriter panel, through the transmit side of the tone converter, and thence to the radio transmitter. When an incoming signal is received, it moves through the receiver and through the receive side of the tone converter. From there it takes the same path through the teletypewriter panel and the switching-control unit as that taken by the transmit signal. Then the signal proceeds to the teletypewriter, where the mark and space signals are converted to the printed (and/or punched) message. Between the teletypewriter and the tone converter, the circuits for transmitting and receiving are identical. The tone converter is the point of divergence. It is the central component in both circuits.

11-338. The same message appears in three forms of signals, in both transmitting and receiving. The teletypewriter signal is a direct-current sequence of on-and-off, or mark-and-space, pulses from the transmit section of the teletypewriter to the tone converter. From the tone converter to the transmitter there is an alternating sequence of two different audio-frequency tones. From the transmitter out over the antenna, the signals are propagated by a tone-modulated radio-frequency carrier wave. Thus a tone-modulated carrier wave is produced, with its mark and space modulations corresponding to the original direct-current mark and space pulses. In the tone converter, the circuit has been automatically switched to the transmit side, placing the receive relay out of operation. Inside the transmit relay, the on-and-off teletypewriter pulses control a two-tone oscillator. A mark decreases the resistance, and a space increases the resistance. In this manner, the transmit side of the tone converter changes the current and no-current pulses into a corresponding sequence of two audio tones, one tone during mark signals and a different tone during space signals.

11-339. In the receiving circuit the arrangement is reversed. From the antenna to the receiver the signal is a tone-modulated carrier wave. From the receiver to the tone converter it is a sequence of audio tones. And from the tone converter through the teletypewriter panel and switching control to the teletypewriter it is direct-current mark-and-space pulses. The receiver picks up the tone-modulated carrier wave with its mark-and-space modulations, and detects the signal, separating the audio tones from the carrier wave. The mark-and-space audio tones then move on to the receive side of the tone converter. The converter, which has now been automatically switched to the receive condition, converts the sequence of audio tones into corresponding direct-current marks and spaces. Thus the message is changed to a form that will actuate the selector magnet in the teletypewriter.

11-340. CARRIER FREQUENCY-SHIFT METHOD. The carrier frequency-shift method is the most dependable radio method for long-distance transmission of teletypewriter signals. This method differs fundamentally from the tone-modulated method by the manner in which the direct-current mark-and-space signals from the teletypewriter modulate the transmitter, and are converted after reception to similar dc pulses for operation of a receiving teletypewriter.

11-341. Figure 11-42 shows a basic radio teletypewriter frequency-shift terminal. During transmission, the teletypewriter may be visualized as a switch that opens and closes the loop circuit furnished by the power supply and thereby produces dc mark-and-space pulses. The mark-and-space pulses are directed through a channel of the teletypewriter panel, and then through the keyer and transmitter. In the keyer, the signal is changed into a corresponding sequence of shifted frequencies that are amplified and multiplied by radio equipment for transmission. Basically, the transmitted signal is frequency-modulated. The marks and spaces are formed by frequency shifts higher and lower than the carrier frequency.

11-342. The mark is the carrier frequency plus the modulating frequency—called a marking shift. The space is the carrier frequency minus the modulating frequency—known as a spacing shift. Thus the frequencies actually being transmitted are always higher or lower than the carrier frequency from which the shift is referenced. The carrier frequency is not transmitted. The keyer receives the carrier frequency and the direct-current on-and-off pulses from the teletypewriter, and shifts it before transmission. The shift from the basic frequency is "plus" for marks and "minus" for spaces.

11-343. The frequency shifts take place in the following manner: In the transmitter, a master oscillator generates the basic radio frequency that is fed into a balanced modulator in the keyer. Coupled to the balanced modulator is a shift oscillator which is controlled by its own shift modulator. As the direct-current mark-and-space pulses feed into the shift modulator, the shift oscillator is actuated to produce a higher frequency for a mark and a lower frequency for a space. The balanced modulator correspondingly increases or decreases the carrier frequency wave to produce the frequency-shifted carrier wave that is fed back into amplifier stages of the transmitter. Transmitter operational changes in the master oscillator frequency do not affect the frequencies of the shift oscillator. Hence, it is the keyer unit through which the signal is converted and the frequency is shifted.

11-344. For reception, a typical installation has two radio receivers, a receiver transfer switchboard, a converter-comparator group, a teletypewriter panel, a switching control, a teletypewriter, and a power supply, as shown in figure 11-42. For simplicity of explanation, the operation of one receiver, one of the converters, and the comparator is described. Assuming that the switching control is set for "carrier frequency shift receive," and that the power supply is furnishing the necessary loop current to the teletypewriter panel, the signal may be followed. The signal in its path from the antenna to the teletypewriter takes various forms; that is, the same message is in three types of signals: from the antenna into the receiver, it is a frequency-shifted radio carrier wave; from the receiver to the converter, it is one of two audio frequencies; and from the converter through the teletypewriter panel and switching control to the teletypewriter, it is direct-current marks and spaces.

11-345. The signal changes take place in the receiver and converter. The receiver is used to tune in the radio-frequency carrier wave with its mark-and-space frequency shifts. Also, the receiver separates the audio frequencies from the carrier wave. The receiver output of mark-and-space audio frequencies is then fed into the converter, where it operates a keyer unit that converts

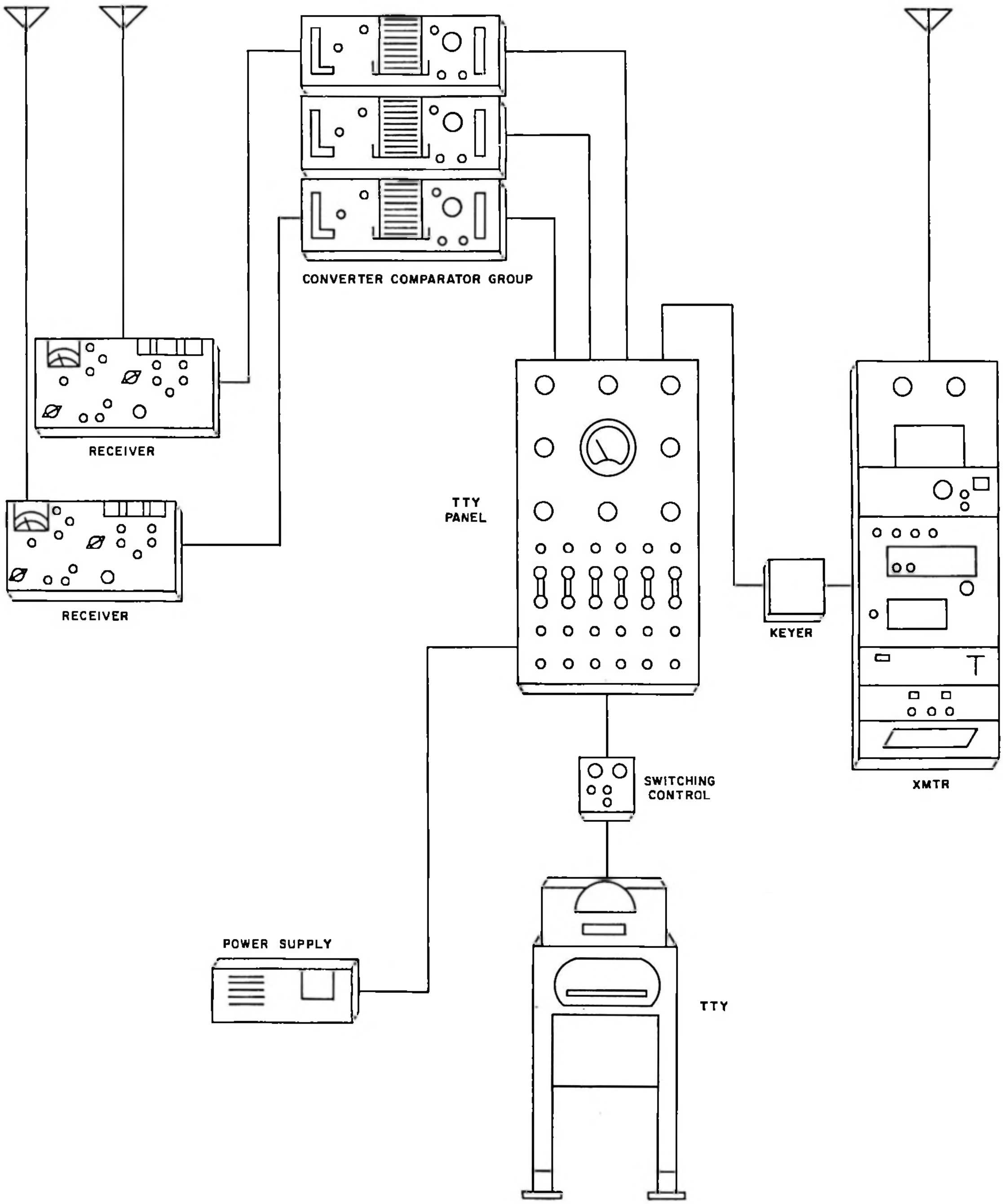


Figure 11-42. Carrier Frequency-Shift Radio Teletypewriter Terminal

the frequencies into their corresponding dc marks and spaces. These mark-and-space signals operate a selector magnet in the teleprinter machine that changes the electrical impulses into printed characters, to form intelligence in a message format. For reliable long-range radio teletypewriter operations, the Air Force generally uses diversity reception. In diversity operation, the same message is received simultaneously by separate receivers. One or the other should get the message through, regardless of weather conditions, interference, and atmospheric variations.

11-346. Figure 11-42 shows the converter-comparator group connections terminating into three separate loops in the teletypewriter panel. For diversity reception, the receivers can be connected to separate antennas and tuned to receive the same message by signals that may continually vary in strength because of atmospheric conditions. The strengths of the audio outputs of the two receivers will vary accordingly. In the converter-comparator group a simple manual adjustment of two switches, on the front of the operating panels, cuts out the converter keyer units and channels the audio signal into a selector unit in the comparator. The selector unit compares the two signals, automatically selects the one that is momentarily the stronger, and sends only the stronger signal to the comparator's keyer unit. The keyer unit changes the shifting audio frequencies into dc marks and spaces. Thus it can be be understood that the converter-comparator group is what makes diversity reception possible. Any two different radio frequency signals carrying the same message simultaneously, but varying in strength, become one series of dc pulses of fairly uniform strength. The comparator output is channeled to the teletypewriter panel, and then into the teletypewriter.

11-347. Note that a single carrier frequency-shift signal can be received through this same converter-comparator. If only one converter is receiving intelligence, its signal is obviously always stronger, and the selector in the comparator favors it continuously. With the diversity circuit, reception can come through one side of the converter-comparator or from the other side. This is a definite advantage, for if there is trouble in one of the receivers or converters, this circuit is unlikely to fail completely. When necessary, a simple switching adjustment in the converter-comparator group can permit two different messages to be received and converted simultaneously. In the adjustment, two different loops in the teletypewriter panel are keyed, and the dc mark-and-space signals are channeled to two separate teletypewriters.

11-348. THE FREQUENCY-SHIFT RADIO SIGNAL. Below 30 mc, where long-distance communications are handled, the most common type of teletypewriter transmission used is the frequency-shift method. This method shifts a transmitter's carrier above and below an assigned operating frequency in accordance with the mark-and-space signals from the sending teletypewriter machine. The amount of frequency shift to the sides of center is generally fixed at plus and minus 425 cps, for a total shift of 850 cps. This standard is in use for operating frequencies between 600 kc and 30 mc. Below 600 kc, narrower shifts of 200 cps for the lf range and 50 cps for the vlf range are in use. In general, no modulation is applied to the two carriers. However, it is possible at most frequency-shift keyers to provide for 200-cps phase modulation, which sometimes proves beneficial under conditions of signal fade or flutter due to multipath reception.

11-349. Two methods of receiving the frequency-shift keying are in use presently; both employ a standard communications receiver, but differ in the means of obtaining output from it. One method, employing one type of converter comparator (AN/URA-6 or 7), obtains its output from the receiver i-f strip by means of a special adapter; another, employing the more common converter-comparator (AN/URA-8), uses the standard audio output from the receiver. The latter method has the advantage of using standard radio communications audio transfer panels, allowing almost any receiver to be used in an emergency. Both types of converter-comparators contain two radio teletypewriter converters with individual keyers to operate teletypewriter machines independently. Also included is a comparator which may be employed during weak-signal reception to select the stronger signal.

11-350. THE DC TELETYPEWRITER SIGNAL. Selector magnets in the older Model 15 machines (TT-5/FG), using a pulling magnet selector unit, require a current of 60 milliamperes (ma) for operation. The model 15 machines, using the newer holding magnet selector unit, have a switch to convert from 60-ma operation (position "P") to 20-ma operation (position "S"). Model 28 machines may be used either way by changing the connections at the selector magnets. Printing errors and erratic operation may result if the current is allowed to drop below its normal value. The current may be monitored and regulated at the teletypewriter transfer panel, where a meter is available. A loop current control is provided for each of the channels.

11-351. The dc source for operating the teletypewriter signal loop is generally obtained from a rectifier power supply unit connected to the teletypewriter transfer panel. In this manner the converter-comparator controls the signal circuit to the machine, providing current and no-current in accordance with the incoming mark-and-space tones from the receiver.

11-352. TELETYPEWRITER MACHINES. Receiving teleprinters are provided with three different types of display: the printed page, the printed tape, and screen projection of page copy. Transmitting teletypewriter equipment is arranged for standard keyboard transmission and perforated tape transmission. Most machines may be fitted with ac-governed, dc-governed, or synchronous motors. Air Force installations, however, should employ the ac-governed motors if frequency and voltage regulation problems are commonly experienced. Governed motors are provided with a black-and-white spotted rotating target painted on the governor used, in conjunction with a tuning fork of appropriate frequency, to adjust the governor to the proper speed. Once set, this speed will hold constant with variations in the ac line of from 50 to 60 cps, and from 104 to 126 volts, ac.

11-353. RADIO TERMINAL CONSIDERATIONS.

11-354. ANTENNAS. The physical condition of the antennas used for radio teletypewriter communications is more critical than of those used in manual telegraph applications. Loose connections, dirty insulators, and excessive leakage to ground along the transmission line will generate noise, which will cause misprints or garbling on the machine. Antennas should be checked with a "megger" ohmmeter at least monthly; the reading for an open-circuited cable should be above 100 megohms. The physical location is next in importance. Receiving antennas are often placed near radar and radio transmitting antennas and still perform satisfactorily for manual cw reception. But if the antenna picks up radiation from the radar, radio transmitter, or other radiating apparatus, it will materially reduce or even completely blank out radio teletypewriter reception. Be sure that the antenna chosen for teletypewriter reception is free from intermittent noise of this nature. Paralleling receivers on one antenna is poor practice because it materially reduces the strength of the received signals. During the reception of weak signals this practice can render the teletypewriter service inoperative. Antenna multicoupler equipments may be employed to overcome this limitation.

11-355. DIVERSITY OPERATION. During weak and fading signal reception conditions, diversity operation is necessary for dependable operation. Several methods are possible; if practicable, all should be tested and the best method selected. All methods require a minimum of two receivers, both converters, and operation of the comparator to key the teletypewriter machine. The diversity methods may be classified as: frequency diversity, polarization diversity, space diversity, and combinations of frequency, space, and polarization diversity. Frequency diversity is set up by copying the same message on two different frequencies with separate receivers, each feeding its output to its own converter. Polarization diversity is set up by copying the same message on the same frequency with separate receivers. One of two receiving antennas can be vertical and the other horizontal. Space diversity is employed at most stations where adequate space is available. It requires a minimum of two receivers to copy the same signal on the same frequency. For this type of diversity operation, the antennas should be separated by two or more wavelengths.

11-356. CHOICE OF FREQUENCIES. The factors controlling frequency selection for radio teletype reception are much the same as those for other radio services. The following characteristics should be given consideration in the choice of frequency:

a. Local- or short-distance reception is good on any hf channel, day or night, with most of the atmospheric noise usually found below 5 mc.

b. Long-distance morning reception is good above 14 mc, with light fading developing in the afternoon, and heavy fading and loss of signal in the evening.

c. Long-distance morning and afternoon reception is good between 9 and 14 mc. Some fading develops in the evening.

d. Long-distance afternoon and evening reception is good between 5 and 9 mc. Signals drop out in the morning.

e. Long-distance evening reception is generally available between 3 and 5 mc. Signals drop out in the early morning.

f. Frequencies and time cannot be exact, and will vary with the season, local weather, year, and sun spot activity. The information given above should be used only as a general guide.

g. In addition, of prime importance in radio teletypewriter reception is receiver stability and selectivity. Receivers which cover the upper half of the hf range (15 to 30 mc) have sufficient stability to make use of sharp selectivity as an aid in eliminating adjacent-channel interference and noise for manual telegraph applications; you may correct drift from time to time with no loss of contact. For radio teletypewriter communications, however, the output tones from the receiver must be maintained closely at their proper pitch. Many receivers will drift out of tune after 10 or 15 minutes, requiring periodic retuning. Also, vibration of mobile equipment can cause serious detuning to some receivers when used above 15 mc. If radio teletypewriter operation is intended above 15 mc, you should first determine whether all voltages and frequency compensating devices employed within the receiver are functioning properly, in order to minimize this drift. Frequency stability and atmospheric conditions are the controlling factors in any upper range hf teletypewriter receiving circuit.

11-357. RECEIVERS. Receiver selection will, of course, depend on the receivers included in the installation. As an aid in selecting the best receiver, you should check the technical order for each receiver, and choose the one which provides the most ac and dc voltage regulation, the best oscillater frequency compensating components, and the best stability against drift due to all causes.

11-358. The following information concerns tuning and adjustment of the receivers. When a new antenna or receiver has been set up for teletypewriter reception, the signal should be found with the receiver set for cw reception, and the receiver then switched to avc-phone operation. Next, the antenna trimmer should be peaked for maximum signal on the input meter. The receiver should then be returned to cw, and only the best oscillator control (vernier, cw pitch, or bfo control) adjusted for proper input to the teletypewriter converter. If this adjustment cannot be made, then and only then the receiver should be detuned slightly, to bring the adjustment in as indicated on the oscilloscope of the teletypewriter converter. If it is necessary to detune the receiver to make this adjustment, the beat oscillator in the receiver needs readjustment. This should be done before weak-signal teletypewriter reception is attempted. Excessive noise may be reduced by operating the receiver set for sharp or narrow (i-f) selectivity. The use of the noise limiter may also be of help.

11-359. TELETYPEWRITER EQUIPMENT TESTING.

11-360. GENERAL. Since all teletypewriters respond to the same functions and operations, they can all be tested for a specific type of operation. However, you must bear in mind that the response of a page printing mechanism is entirely different from the response of a typing reperforator to the same function. Teletypewriter mechanisms are grouped below, so that you can know what to expect from a given group of machines.

11-361. Teletypewriters in the group including machines such as the TT-5/FG, TT-7/FG, TG-7-( ), etc, are to be designated as Model 19 series. The group including the TT-47/UG, TT-48/UG, TT-69/UG, and TT-70/UG are to be designated as Model 28. Model 14 series will include typing reperforators.

11-362. SIMPLE KEYBOARD AND TAPE TRANSMITTER TESTS. Observation of equipment during operation is usually the first test of a suspected printer. Transmitting with the receiver converter-comparator set on "tune" will provide a closed, continuous loop for testing purposes. If a tape transmitter is included with the installation, it should be patched into the testing channel and used instead of the keyboard. Tape transmitters generally operate at a constant 60 words per minute, and thus provide a more rigid test of the receiving printer.

11-363. Test tapes may be prepared on the keyboard of a reperforator machine, and connected end-to-end with transparent tape to provide a continuous test signal. For the text of the tape, many good tests are available; the chief requirement is that it contain all the functions and characters used. R and Y are used commonly as the two letters, since they require the selector unit to reverse all its code pulse components from R to Y.

11-364. Using the keyboard of the suspected machine is a good test, but not a conclusive one. If the machine's motor is of the governed type, it will be possible for its speed to be too high or too low and still work correctly with its own keyboard. Both are kept in synchronism by means of the main shaft gears. If a machine will not copy an incoming signal, check the machine first with its own keyboard. If it passes this test, try sending to it from another machine or tape transmitter known to be in good condition. If it passes this test, the machine is operating satisfactorily, and the incoming signal or the keyer in the converter is at fault. If it fails either test, the machine needs attention.

11-365. ORIENTATION AND SELECTION. Check the range of the teletypewriter machine by alternately printing the letters R and Y while slowly moving the range-finder pointer toward the lower end of the scale until printing errors occur. Note the position of the pointer. Check the upper limit of the range by moving the range-finder pointer toward the upper end of the scale until errors are printed. Note the position of the pointer again. Then set the range finder at the midpoint of these two extremes. Print the "test sentence" to check the selection process. The range, under local test conditions with a well adjusted machine, should be 80 points; however, a 70-point range is acceptable.

11-366. FIGURES AND LETTERS SHIFT ALIGNMENT. The shifting of a teletypewriter is accomplished in an operational sequence by operating FIGS key, figure "2" key, LTRS key, and "W" key for several operations. Note that the carriage shifts correctly, that the upper and lower edges of the letters and figures print uniformly, and that the characters line up evenly across the page. This same test is valid for the typing reperforators; however, you must remember that the shifting motion is in a horizontal plane from front to back, and vice versa.

11-367. Since the shifting is accomplished on a horizontal plane by the typebox, this test is not very useful for the Model 28 series of teletypewriters. One of the easiest ways to check the Model 28 series for shifting is to depress the LTRS key and then the letter "M." Stop the motor of the machine, manually trip the printing hammer, and checking to make sure that the rear edge of the type pallet is centered with respect to the printing hammer. Grasp the track which mounts the typebox and gently push from side to side to see whether there is excessive end play; the edges of the type pallet should always be lined up with the edge of the printing hammer in either direction. Energize the motor, depress the FIGS key, and depress the PERIOD key; then stop the motor. Follow the same procedure as before.

11-368. LINE FEED TEST. Position the single-double line feed lever to the single position, and operate the LINE FEED key and some character key alternately. The characters should be printed in a straight diagonal line across the page in single spaces. Double line feed should be similarly checked. This method is applicable to all page printer mechanisms; however, a typing reperforator will only punch the tape for a line feed and print a character having three horizontal bars (≡).

11-369. The Model 28 series of teletypewriters have a red keybutton on the keyboard, labeled LF. Depressing this keybutton will cause the platen to rotate, continuously feeding the paper through the platen until the key is released.

11-370. CARRIAGE RETURN. With the typebar carriage at the extreme right position, operate the CAR RET key. The carriage should return to the extreme left position smoothly and quickly, without hesitation, bouncing, or jarring. Each line should begin directly beneath the previous line.

11-371. The Model 28 series of teletypewriters have a red keybutton on the keyboard, labeled CAR RET. Depressing this keybutton will cause the carriage to return to the starting point at any time it is depressed.

11-372. The typing reperforator will punch the code combination for carriage return and print a "< ", since these machines are not fitted with a movable carriage mechanism.

11-373. SPACING. Depress the space bar partially to make sure that the carriage travels one space at a time. Depress the space bar fully and make sure that the machine continuously repeats the operation. Model 28 series teletypewriters will not repeat unless the repeat key (red keybutton) is also depressed.

11-374. LEFT MARGIN. The left side of the letter "M" should be printed approximately 7/8 inch from the left edge of the platen—not the paper edge. This does not apply to typing reperforator mechanisms.

11-375. RIGHT MARGIN. Check to make sure that the machine will print lines of exactly 72 characters. The 73rd character should operate the carriage return and line feed functions, and overprint the character sent. This is not applicable to typing reperforators.

11-376. BIAS AND END DISTORTION. You can determine the amount of signal bias and end distortion that the teletypewriter selector unit can tolerate by using Test Set TS-2B/TG or a similar signal distortion tester. If the test set or teletypewriter is equipped with governed motors, the motor speed must first be adjusted. Normally, this is done by using a tuning fork and the target spots on the governor to check the speed. The character selecting switch of the test set is then set to transmit either R or Y. The adjustable distributor disk of the test set is set to the desired percentage of bias and end distortion; this control is usually set for 35 percent.

11-377. Bias is a type of teletypewriter signal distortion which affects all impulses uniformly. It has the effect of advancing or delaying the beginning of each mark impulse with respect to the beginning of the character cycle. For the bias check, set the test set distortion switch to the BIAS M position; this provides an output signal with a fixed amount of marking bias. Check the upper limit of the range by moving the teletypewriter range-finder pointer toward the upper end of the range scale until errors are printed. Then set the distortion switch to the BIAS S position and check the lower limit of the range by moving the range-finder pointer toward the lower end of the range scale until errors are printed. Midway between these two settings is the point on the range scale where the teletypewriter will tolerate the maximum bias. This point is then considered as the optimum setting for bias. When the adjustable distributor disk is set for 35-percent bias, you can determine the maximum amount of signal bias that can be tolerated by the formula:

$$\begin{matrix}\text{Maximum}\\ \text{bias}\\ \text{(percent)}\end{matrix} = 35 + \frac{\begin{matrix}\text{Upper limit}\\ \text{bias M}\end{matrix} - \begin{matrix}\text{Lower limit}\\ \text{bias S}\end{matrix}}{2}$$

11-378. End distortion is another type of teletypewriter signal distortion which affects all impulses uniformly. It has the effect of advancing or delaying the end of each mark impulse with respect to the beginning of the character cycle. To measure the amount of end distortion, set the distortion switch to ED M; this provides an output signal with a fixed amount of marking and distortion. Check the lower limit of the range by moving the teletypewriter range-finder pointer toward the lower end of the range scale until errors are printed. Then set the distortion switch to the ED S position and check the upper end of the range by moving the range-finder pointer toward the upper end of the range scale until errors are printed. Midway between these two settings is the point on the range scale where the teletypewriter will tolerate the maximum end distortion. This point is then considered as the optimum setting for end distortion. When the adjustable distributor disk is set for 35-percent end distortion, you can determine the maximum amount of end distortion the teletypewriter will tolerate by the formula:

$$\text{Maximum end distortion (percent)} = 35 + \frac{\dfrac{\text{Upper limit}}{\text{ED S}} - \dfrac{\text{Lower limit}}{\text{ED M}}}{2}$$

11-379. The internal bias of the teletypewriter receiving selector unit may be calculated by subtracting the optimum setting for bias from the optimum setting for and distortion.

11-380. MUTUAL INTERFERENCE PROBLEMS.

11-381. In a communications system, the number of radio circuits employed may necessitate the use of many transmitters and receivers in the same general locality. The selection of frequency assignments to avoid mutual interference between communication equipments becomes increasingly more complex as the proximity and/or number of equipments is increased. While it is possible to estimate reasonably well the frequencies at which interference is likely, and to draw general conclusions as to the degree of interference, it is impracticable to predict accurately the seriousness of interference at specific frequencies. The information given in the following paragraphs will assist in system planning, by providing a discussion of interference sources and a means of reducing the mutual interference problem.

11-382. TRANSMITTER-TO-RECEIVER INTERFERENCE.

11-383. GENERAL. Transmitter-to-receiver interference is the major type of interference, and is difficult to avoid in an area where many transmitters and receivers are located in proximity to one another, because the desired signals arriving at the receivers from distant transmitters are usually weak and can thus be interfered with easily by nearby transmitters.

11-384. Each transmitter radiates small amounts of energy at many frequencies other than its fundamental carrier frequency, and each receiver is responsive, although very inefficiently, to signals of many frequencies other than the desired frequency to which it is tuned. In addition, strong signals from two nearby transmitters operating on different frequencies may heterodyne in a receiver to produce interference. These spurious radiations and responses greatly increase the number of frequencies at which interference can occur. For example, when transmitting and receiving antennas are located close together, interference may result, not only in receivers tuned to frequencies near that of a strong transmitter carrier, but also in receivers tuned to frequencies corresponding to the weaker spurious transmitter harmonic radiations. In addition, interference may result when the strong transmitter carrier frequency corresponds to one of the many weak spurious responses of the receiver. It is also possible for interference to occur when a spurious transmitter radiation corresponds to one of the receiver spurious responses, but serious interference of this type is not normally experienced.

11-385. ANTENNA LOCATION. Much of the interference that occurs when transmitting and receiving antennas are located close together disappears when the intensities of interfering signals are reduced by separating the transmitting and receiving antennas. Such attenuation by physical separation of antennas has proved successful in practice, since the receiver sensitivity at spurious response frequencies is dependent upon the intensity of the interfering signal. For these cases, a decrease of 20 db in the field intensity, resulting from increased antenna separation, may reduce the receiver sensitivity to spurious response by another 20 db or more, thus rapidly eliminating such interference.

11-386. Because of the advantages brought about through the separation of transmitting and receiving antennas, it follows that careful consideration should be given to the possibility of providing separate sites for transmitting and receiving antennas. Antenna separations of as much as 5 miles are not uncommon where transmitter outputs of 400 watts and above are involved; however, a spacing of 1 or 2 miles should permit reasonably flexibility in the assignment of operating frequencies in the 3- to 30-megacycle range. The separation of transmitting and receiving antennas and the associated transmitters and receivers generally requires the use of remote control equipment.

11-387. Mutual interference can exist between communication equipments operating in widely separated frequency bands if these equipments are in close proximity to one another. For example, high-frequency transmitters emit spurious radiations at many times the assigned operating frequency, and these radiations may cause interference in very-high-frequency receivers located nearby. Also, high-frequency receivers are frequently susceptible to spurious responses in the very-high-frequency band, and may pick up interference from a nearby vhf transmitter.

11-388. As an alternative to space attenuation by means of antenna separation, the directional characteristics of antennas may be used to advantage in reducing mutual interference. Orientation and physical separation of the transmitting and receiving antennas may provide satisfactory results in some installations. Where the transmitting and receiving antennas must be located near each other, consideration should be given to the possibility of using the natural contours of the area to provide antenna isolation. Terrain shielding can be satisfactorily employed in locations where the orientation of the antenna radiation patterns permits such an arrangement.

11-389. When space is not provided to separate transmitting and receiving antennas, as with mobile equipment, for example, certain types of interference cannot be so easily eliminated. With numerous transmitting and receiving antennas separated by only a few hundred feet or so, the number of interference possibilities increases to such an extent that the solution of the operating frequency assignment problem becomes very difficult, if not impracticable. If such a situation cannot possibly be avoided, it is then essential to understand the equipment characteristics responsible for the interference, to aid in selecting initial operating frequencies that will have some prospect of proving reasonably satisfactory when operations are begun. When interference does occur under these conditions, a knowledge of the various possible interference factors will permit a more rapid determination of the type of interference involved, and a more logical approach to the frequency rearrangement problem.

11-390. SPURIOUS TRANSMITTER RADIATION. Spurious transmitter radiations account for much of the transmitter-to-receiver interference when transmitting and receiving antennas are located in close proximity to each other. Most of the power in the output of a transmitter is at the operating carrier frequency and its relatively narrow sidebands. However, a small amount of power is always present at all harmonics of the master-oscillator or crystal-oscillator frequency. Such harmonic radiation is particularly undesirable from the interference standpoint, as well as the fact that the generation of these harmonics represents a definite power loss and, therefore, lowered efficiency.

11-391. In high-frequency transmitters, the operating carrier frequency is usually the same as, or a multiple of, the master-oscillator or crystal-oscillator frequency. For example, if the crystal-oscillator frequency is 3 megacycles, the operating carrier frequency, in megacycles, could be chosen as 3, 6, 9, 12, 15, or any other multiple of 3 megacycles, depending on the transmitter design. Assuming the operating carrier frequency to be 3 megacycles in this case, spurious transmitter radiations could take place at multiples of 3 megacycles throughout the high-frequency spectrum. If a master oscillator were employed as the frequency-determining device, rather than a 3-megacycle quartz crystal, it is quite possible that the fundamental frequency of the master oscillator might be chosen as 1.5 megacycles for an operating carrier frequency of 3 megacycles. In this case, in addition to the desired output at 3 megacycles, spurious transmitter radiations could take place at multiples of 1.5 megacycles, as, for example, radiations at 1.5, 4.5, 6, 7.5, 9, etc, megacycles. The design of a transmitter usually incorporates features which help to eliminate, or at least suppress, these harmonics.

11-392. Because of possible harmonic radiation from transmitters, guy wires which support antenna masts should be divided into insulated sections, the lengths of which are not harmonically related to the operating frequency. Failure to observe this practice could lead to excitation of, and reradiation from, a resonant length of guy wire by spurious transmitter radiations.

11-393. Since the frequencies of these spurious outputs are different for different operating carrier frequencies, it is apparent that the number of receiving channels which may be interfered with increases rapidly with the number of transmitters in operation, as each transmitting frequency will be accompanied by a series of spurious radiations displaced in frequency from any other series. Spurious transmitter radiations can also be caused by cross-modulation between two transmitters.

11-394. CROSS MODULATION. When an undesired signal feeds through into an operating circuit, so that the undesired signal modulates or interferes with the circuit, the defect is known as cross modulation. Cross modulation occurs in a receiver when the carrier to which the receiver is tuned is modulated (or interfered with) by the modulation of an adjacent-channel signal. Cross modulation, in this case, is most apparent during the temporary lapses of modulation of the received carrier. Cross modulation can also result when two transmitters are in close proximity. When the antennas or transmission lines of two transmitters are located near each other, an appreciable radio-frequency voltage from one transmitter may be impressed across the output tank circuit of the other. Because of nonlinear voltage-current characteristics in the final amplifier circuit, this induced voltage causes the generation and radiation of spurious frequencies at other than the operating carrier frequency of either transmitter.

11-395. For example, if one transmitter operating at 24 megacycles has impressed across its output tank circuit a voltage at 19 megacycles from a nearby transmitter, a spurious frequency of 29 megacycles will be formed in the output circuit of the first transmitter. This spurious signal, which is the sum of the difference frequency and the lower transmitted frequency, will be radiated after being attenuated by the tank, antenna coupling, and antenna circuits. Other spurious frequencies will be generated similarly at 14, 34, 9, 39, etc, megacycles in the order of their relative strengths; however, these spurious frequencies will be of importance only in extreme cases. A difference frequency of 5 megacycles (24 — 19) will also be generated. In this example, interference with a nearby high-frequency receiver operating at 5 megacycles may be caused, as well as interference with other nearby high-frequency receivers operating at frequencies that correspond to the spurious frequencies generated by the cross modulation.

11-396. Interference of this type should not prove serious if the receiving antennas are well separated from the transmitting antennas, since the spurious frequencies are highly attenuated. If they are not, the interference may be reduced by: increasing the separation between the transmitting and receiving antennas; reorienting the antennas of transmitters which are reacting upon each other so as to take advantage of the minima of the directional radiation patterns; avoiding or removing any condition where two adjacent transmitting antennas are tuned to the same or nearly the same operating frequency; and observing the proper separation of adjacent transmission lines.

11-397. RECEIVER-TO-RECEIVER INTERFERENCE.

11-398. Spurious receiver radiations account for much of the receiver-to-receiver interference. This type of interference is usually not serious, but must sometimes be considered in situations where receiving antennas are closely grouped or when several receivers are connected to a common antenna. Almost without exception, communication receivers are of the superheterodyne type. The fundamental or harmonic frequencies of the local oscillator in one receiver may reach the input of a nearby receiver through several signal paths, the most important of which is probably through coupling between the antennas used in conjunction with the receivers.

11-399. In some cases, several receivers operating on different frequencies may be connected to a common antenna, in which case suitable isolation between the receivers must be provided, or the operating frequency of each receiver chosen so that local-oscillator radiation does not present an interference problem. A multicoupler unit is often used to reduce the effect of local-oscillator radiation, and is basically a distributing amplifier unit operating from a single antenna, which affords some isolation for each of the several receivers. Since the efficiency of multicoupler units is rather low (about 25 percent), a far more satisfactory arrangement is to provide separate receiving antennas which are physically separated and/or oriented to avoid the receiver local-oscillator interference problem.

11-400. SPURIOUS RECEIVER RESPONSES.

11-401. Spurious receiver responses may be defined as responses which occur in superheterodyne receivers when, by some combination of circumstances, a frequency equal to the intermediate frequency is produced in the output of the mixer (or converter) stage, or in the intermediate-frequency stages themselves.

11-402. A superheterodyne receiver which is tuned to receive a signal of a certain frequency will respond best at that frequency. However, the receiver will also respond inefficiently to signals at numerous other frequencies scattered over a wide band of frequencies extending above and below the normal operating frequency of the receiver.

The exact location in the radio spectrum of the spurious response frequencies is a function of receiver design. The relative efficiency or sensitivity of the receiver at these spurious response frequencies depends on such factors as the selectivity of the antenna, the selectivity of the receiver, the degree of filtering in the local-oscillator output circuit, and, of course, the intensity of the interfering signal at the spurious response frequency.

11-403. When the receiver is properly aligned, the sensitivity of the receiver at spurious response frequencies is always much less than at the desired operating frequency; however, the sensitivity may be sufficient to permit interference from a local transmitter which radiates a relatively strong carrier (or harmonic) at a spurious response frequency. If such a spurious signal reaches the limiter grid circuit in an fm receiver, or the detector in an a-m receiver, with an amplitude comparable to the desired signal, serious interference with the reception of voice or teletypewriter transmission will result. In the reception of cw (by ear), substantially more interference can be tolerated by employing skilled radio operators on the affected circuit.

11-404. Spurious intermediate-frequency responses occur when signals within the i-f pass band are picked up directly in superheterodyne receivers having insufficient shielding or insufficient rf selectivity. The rf selectivity is an important factor when the operating frequency is within 20 percent of the intermediate frequency, which is generally possible in receivers designed for use in the low-frequency or medium-frequency bands. In such cases it is ordinarily specified that the receiver should be at least 80 db less sensitive to an interference signal at the intermediate frequency applied to the receiver input terminals than to a signal at the desired operating frequency.

11-405. Spurious receiver responses which are most common occur at the "image" frequency of the receiver. This response is caused by the fact that there are two possible signal frequencies at which normal superheterodyne receiver action can take place to produce the intermediate frequency: one of these signal frequencies is above the local-oscillator frequency by the amount of the i-f, and the other is below the local-oscillator frequency by the same amount. The antenna and rf amplifier circuits are tuned to amplify the desired operating (signal) frequency. These circuits are therefore detuned (or off-resonance) with respect to the image frequency, so that a signal at this frequency is attenuated considerably before reaching the mixer (or converter) grid, thus reducing the receiver sensitivity at the image frequency. The degree of such image rejection depends on the receiver design.

11-406. A series of spurious receiver responses can also occur at submultiples of the operating frequency, such as one-half, one-third, or one-fourth of the frequency to which the receiver is tuned. When such frequencies are applied to the rf stage(s) with sufficient signal strength, the rf amplifier output will include harmonics of these frequencies because of the nonlinear characteristics of the rf stage(s). One of these harmonics thus produced will equal the operating frequency to which the receiver is tuned, and this signal will be amplified and detected in the same manner as the desired signal. Submultiple frequencies which are one-fifth (or less) of the operating frequency are usually of no consequence, and may be weak enough to be neglected. These frequencies are sufficiently removed from the operating frequency to be attenuated by the antenna and rf amplifier circuits, and the resulting high-order harmonics produced by nonlinear action are much weaker. Responses of this type are easily eliminated by increased separation between the transmitting and receiving antennas.

11-407. Spurious receiver responses can also be caused as a result of harmonics generated by the local oscillator of the receiver. Any harmonic of the local oscillator can beat with certain signal frequencies present at the mixer stage to produce the intermediate frequency. The frequencies at which such spurious responses can occur are equal to the local-oscillator harmonic frequencies plus or minus an amount equal to the intermediate frequency. The most important response frequencies to consider are those which are associated with the second, third, and fourth harmonics of the receiver local oscillator.

11-408. Spurious receiver responses can occur as a result of the combination of a strong local signal and the resultant harmonics produced in the rf stage, and local-oscillator harmonics. As mentioned above, harmonics of any strong incoming signal can be produced in the output of rf stages. These harmonic frequencies heterodyne with the fundamental or harmonic frequencies of the local oscillator to produce a multitude of beats, the most important of which are those that fall near the desired operating frequency. These spurious responses are so numerous near the desired operating frequency that it is very difficult to operate transmitters at frequencies within 10 percent of the receiver operating frequency without causing interference. The likelihood of interference from this source is decreased rapidly as the transmitting and receiving antennas are separated.

11-409. The desired carrier frequency may be modulated by the modulating frequencies of a strong undesired carrier in the rf or mixer stages to cause cross modulation. Cross modulation may also be caused by rectifying junctions, as a result of poor electrical contact between metal surfaces. In addition to interfering with the desired signal, the undesired signal may be strong enough to develop considerable avc voltage in the receiver, or self-biasing of rf stages, thus causing the receiver to be less sensitive to the weaker desired signal.

11-410. SUPPRESSION OF HARMONIC AND PARASITIC SIGNALS.

11-411. GENERAL. The desired carrier frequency may be modulated by the modulating frequencies of the strong undesired carrier in the rf or mixer stages to cause cross modulation. (Another cause of cross modulation is the introduction of nonlinear conductors, either in the antenna system or in proximity to it.) Also, it should be noted that the undesired carrier may be strong enough to develop considerable avc voltage in the receiver, or self-biasing of rf stages, thus causing the receiver to be less sensitive to the weaker desired signal, together with an attendant reduction in audio output.

11-412. The degree to which transmitter harmonics or other transmitter spurious radiations must be suppressed depends principally on two factors: the strength of the signal that is being interfered with, and the relationship between the spurious radiation and the frequency of the signal that is being interfered with. When the affected signal is strong, the problem of transmitter suppression is relatively simple; when it is weak, a much greater suppression must be effected, and the problem becomes much more complex. In either case, the intensity of the interference depends very greatly on the exact frequency of the interfering signal. Effective harmonic suppression has three separate phases.

a. Reducing the amplitude of harmonics generated in the transmitter. This is a matter of circuit design and operating conditions.

b. Preventing stray radiation from the transmitter and associated wiring. This requires adequate shielding and filtering of all circuits and leads from which radiation can take place.

c. Preventing harmonics from being fed into the antenna. There is no transmitter that will not generate some harmonics, but it is obviously advantageous to reduce their strength, by circuit design and choice of operating conditions, by as large a factor as possible before attempting to prevent them from being radiated.

11-413. REDUCTION OF HARMONIC GENERATION. Harmonic radiation from the transmitter or from its associated wiring will obviously cause interference just as readily as radiation from the antenna; therefore, measures taken to prevent harmonics from reaching the antenna will not reduce interference if the transmitter itself is radiating harmonics. But once it has been found that the transmitter itself is free from harmonic radiation, devices for preventing harmonics from reaching the antenna can be expected to produce results. Since reasonably efficient operation of rf power amplifiers is always accompanied by harmonic generation, good transmitter design calls for the operation of frequency-multiplier stages at a low power level (plate voltages not exceeding 250 or 300 volts). When the final output frequency is reached, it is desirable to use as few stages as possible in reaching the output power level, and to use tubes that require a minimum of driving power.

11-414. Harmonic currents of considerable amplitude flow in both the grid and plate circuits of rf power amplifiers, but they will do relatively little harm if they can be effectively bypassed to the cathode of the tube. The paths followed by harmonic currents in an amplifier circuit, because of the high reactance of the tank coil, are usually through the tank capacitor, the plate (or grid) blocking capacitor, and the tube capacitances. The lengths of the leads forming these paths are of great importance, since the inductance in a circuit will resonate with the tube capacitance at some frequency. (The tank and blocking capacitances are usually so large as compared with the tube capacitance that they have little effect on any parasitic resonant frequency.) If such a resonance happens to occur at or near the same frequency as one of the transmitter harmonics, the effect is just the same as though a harmonic tank circuit had been deliberately introduced; the harmonic at that frequency will be tremendously increased in amplitude.

11-415. Such resonances are unavoidable, but the use of short, direct leads in the plate and grid circuits can increase the parasitic resonant frequency to such a high frequency that it does not affect the operation. If this method fails, then low-inductance capacitors can be connected between the grid and cathode and between the plate and cathode, to lower the parasitic resonant point to a frequency which will not add to the harmonic output. These capacitors will also form new resonant circuits with the tube capacitance and connecting leads, but this generally occurs at a sufficiently high frequency that no harm is done.

11-416. DETECTION OF RESONANT POINTS. The vhf resonance points in amplifier tank circuits can be found by coupling a grid-dip meter to the grid and plate leads. If a resonance is found in or near the affected frequency, a method such as that described above may be used to move it well out of the range. The grid-dip meter may also be used to check for vhf resonances in the tank coils. In making this check, the coil must be disconnected entirely from the transmitter, and the grid-dip meter coil moved along the coil while exploring for a dip. If a resonance fails at the point of interference, changing the number of turns might be possible, and will move the point of resonance to a frequency where it will not be troublesome.

11-417. OPERATING CONDITIONS. Grid bias and current have an important effect on the harmonic content of the rf currents in both the grid and plate circuits. In general, the harmonic output increases as the grid bias and grid current are increased, but this is not necessarily true of a particular harmonic. The third and higher harmonics, especially, will go through fluctuations in amplitude as the grid current is increased, and sometimes a rather high value of grid current will minimize one harmonic as compared with a low value of grid current. This characteristic can be used to advantage where a particular harmonic is causing interference, keeping in mind that the operating conditions that minimize one harmonic may greatly increase another.

11-418. For equal operating conditions, there is little or no difference between single-ended and push-pull amplifiers with respect to harmonic generation. Push-pull amplifiers are frequently trouble-makers on even harmonics, because with such amplifiers the even-harmonic voltages are in phase at the ends of the tank circuit and hence appear with equal amplitude across the whole tank coil, if the center of the coil is not grounded. Under such circumstances the even harmonics can be coupled to the output circuit through stray capacitance between the tank and coupling coils. This does not occur in a single-ended amplifier if the coupling coil is placed at the grounded end of the tank.

11-419. REDUCTION OF TRANSMITTER RADIATION.

11-420. GENERAL. Transmitter radiation can be a very serious problem when a transmitter is operated at high power and radio receivers are in close proximity.

11-421. CHECK FOR TRANSMITTER RADIATION. A check for transmitter radiation should always be made before attempting to use low-pass filters or other devices for preventing harmonics from reaching the antenna. The only really satisfactory indicating instrument is a sensitive receiver or panoramic adapter. In regions where the received signal is strong, an indicating wave meter such as one having a crystal or tube detector may be useful; if it is possible to get any indication at all on harmonics, either on the supply leads or around the transmitter itself, the harmonics are probably strong enough to cause interference. However, the absence of any such indication does not mean that harmonic interference will not be caused. If the techniques of shielding and lead filtering are incorporated in the transmitter, the harmonic intensity on any external leads should be far below what any such instruments can detect.

11-422. Radiation checks should be made with the transmitter delivering full power into a dummy antenna, such as an incandescent lamp of suitable power rating, preferably installed inside the shielded enclosure. If the dummy antenna must be external, it is desirable to connect it through a coaxial-matching circuit. Shielding the dummy antenna circuit is also desirable, although it is not always necessary.

11-423. Make the radiation test on all frequencies that are to be used in transmitting, and note whether or not interference is picked up in the receiver or shows in the panoramic adapter. (These tests must be made while the signal being interfered with is being received, since the best tones will not be formed if the signal carrier is not present.) If interference exists, its source can be detected by grasping the various external leads (by the insulation—not the live wire), or by bringing the hand near meter faces, louvers, and other possible points where harmonic energy might escape from the transmitter. If any of these tests cause a change—not necessarily an increase—in the intensity of the interference, the presence of leakage at that point will be indicated. The location of such spots will usually simplify the determination of the remedy. If the receiver and the transmitter can be operated side-by-side, a length of wire connected to one antenna terminal on the receiver can be used as a probe to go over the transmitter enclosure and external leads. This device will very quickly expose the spots from which serious leakage is taking place.

11-424. As a final test, connect the transmitting antenna or its transmission line terminals to the outside of the transmitter shielding. Interference created when this test is applied indicates that weak currents are on the outside of the shield, and can be conducted to the antenna when the normal antenna connections are used. Currents of this nature represent interference that can be conducted over low-pass filters, etc, and which, therefore, cannot be eliminated by such filters.

11-425. SHIELDING. Direct radiation from the transmitter circuits and components can be prevented by proper shielding. To be effective, a shield must completely enclose the circuits, and the parts must have no openings that will permit rf energy to escape. Unfortunately, ordinary metal boxes and cabinets do not provide good shielding, since such openings as louvers, lids, holes for running in connections, etc, allow leakage.

11-426. A primary requisite for good shielding is that all joints must make a good electrical connection along their entire length. A small slit or crack will let out a surprising amount of rf energy; so will ventilating louvers and large holes, such as those used for mounting meters. On the other hand, small holes do not impair the shielding very greatly, and a limited number of ventilating holes may be used if they are small (not over 1/4 inch in diameter). Also, wire screen makes quite effective shielding if the wires make good electrical connection where they cross over; therefore, the leakage through large openings can be very much reduced by covering such openings with screening that is well bonded to all edges of the opening.

11-427. The intensity of rf fields about coils, capacitors, tubes, and wiring decreases very rapidly with distance; thus shielding is more effective, from a practical standpoint, if the components and wiring are not too close to the fields. It is advisable to have a separation of several inches, if possible, between the "hot" points in the circuit and the nearest shielding.

11-428. For a given thickness of metal, the greater the conductivity the better the shielding. Copper is best, with aluminum, brass, and steel following in that order. However, if the thickness is adequate for structural purposes (over 0.02 inch) and the shield and a "hot" point in the circuit are not in close proximity, any of these metals will be satisfactory. Greater separation should be used with steel shielding than with the other materials, not only because it is considerably poorer as a shield, but also because it will cause greater losses in neary-by circuits than will copper or aluminum at the same distance. Wire screen used as a shield should also be kept at some distance from high-voltage or high-current rf points, since there is considerably more leakage through the mesh than through solid metal.

11-429. Where two pieces of metal join, as in forming a corner, they should overlap at least a half inch, and be fastened together firmly with screws or bolts spaced at close-enough intervals to maintain firm contact all along the joint. The contact surfaces should be clean before joining, and should be checked occasionally—especially steel, which is almost certain to rust after a period of time.

11-430. The leakage through a given size of aperture in shielding increases with frequency; therefore, such factors as good continuous contact, proper screening of holes, etc, become critical for the ultra-high-frequency band. Hence, transmitters, which in general have frequency-multiplier harmonics of relatively high intensity in this region, require special attention in this respect if there is a possibility of interfering with a channel received locally.

11-431. Even very good shielding can be made completely useless when connections are run from external power supplies and other equipment to the circuits inside the shield. Every conductor so introduced into the shielding forms a path for the escape of rf, which is then radiated by the connecting wires. Hence a step that is essential in every case is to prevent harmonic currents from flowing on the leads leaving the shielded enclosures.

11-432. Harmonic currents always flow on the dc or ac leads connecting to the tube circuits. A very effective means of preventing such currents from being coupled into other wiring, and one that provides desirable bypassing as well, is to use shielded wire for all such leads, maintaining the shielding from the point where the lead connects to the tube or rf circuit, all the way through to the point where it is about to leave the chassis. The shield braid should be grounded to the chassis at both ends and at frequent intervals along the path.

11-433. These bypasses are essential at the connection-block terminals, and desirable at the tube ends of the leads also. Properly installed, they have been found to be so effective that there is usually no need for further harmonic filtering. However, if a test shows that additional filtering is required, an rf filter should be installed at the tube end of the shielded lead; if more than one circuit is filtered, care should be taken to keep the rf chokes separated from each other, and so oriented as to minimize coupling between them. This is necessary for preventing harmonics present in one circuit from being coupled into another.

11-434. Meters that are mounted in an rf unit should be enclosed in shielding covers, the connections being made with shielded wire, and with each lead bypassed to ground. The shield braid should be grounded to the panel or chassis immediately outside the meter shield. A bypass may also be connected across the meter terminals, principally to prevent any fundamental current that may be present from flowing through the meter itself. As an alternative to individual meter shielding, the meters may be mounted entirely behind the panel, and the panel holes needed for observation may be covered with wire screen that is carefully bonded to the panel all around the hole.

11-435. HARMONIC TRAPS. If only one harmonic is particularly bothersome, a parallel-resonant circuit trap tuned to the harmonic frequency may be installed in the plate lead of the amplifier (or in each lead for push-pull circuits). At the harmonic frequency the trap represents a very high impedance, and hence reduces the amplitude of the harmonic current flowing through the tank circuit. In the push-pull circuit, both traps have the same constants. The inductance-to-capacitance ratio is not critical, but a high-capacity circuit will usually have the least effect on the performance of the plate circuit at the normal operating frequency.

11-436. Since there is a considerable harmonic voltage across the trap, it may radiate unless the transmitter is well shielded. Traps should be placed so that there is no coupling between them and the amplifier tank circuit.

11-437. A trap is a highly selective device, and is useful, therefore, only over a small range of frequencies. For example, a second- or third-harmonic trap for a 30-mc tank circuit will usually not be effective over more than 50 kc at the fundamental frequency, depending on how serious the interference is without the trap. Because traps are critical of adjustment, it is better to prevent interference by other means, when possible.

11-438. PREVENTING HARMONICS FROM REACHING THE ANTENNA. Equally important in reducing harmonic interference is to keep the spurious energy generated in or passed through the final amplifier stage from traveling over the transmission line to the antenna. It is seldom worthwhile to attempt this until the radiation from the transmitter and its connecting leads has been reduced to the point where, with the transmitter delivering full power into a dummy antenna, it has been determined by actual testing that the radiation is below the level that can cause interference. If the dummy antenna test indicates enough radiation to be readily received, it is a practical certainty that harmonics will be coupled to the antenna, no matter what preventive measures are taken external to the transmitter.

11-439. In inductively coupled transmitter outputs, some harmonic energy will be transferred from the final amplifier through the mutual inductance between the tank coil and the output coupling coil. Harmonics of the output frequency transferred in this way can be greatly reduced by providing sufficient selectivity between the final tank and the transmission line. A good deal of selectivity, amounting to a 20 to 30 db reduction of the second harmonic, and a much higher reduction of higher-order harmonics, may be obtained by use of a matched antenna coupler (tuner).

11-440. Transmitters operating above 30 mc may produce harmonics not directly associated with the output frequency, such as those generated in low-frequency early stages of the transmitter, and coupled to the antenna by stray paths. This may cause a spurious signal of such a nature that the selectivity of the amplifier tank circuits is not sufficient to prevent its being coupled to the antenna. Spurious signals of this type are reduced when link coupling is used between the driver stage and the final amplifier (and between earlier stages), in addition to the suppression afforded by using an antenna coupler.

11-441. CAPACITIVE COUPLING. Inasmuch as a coil is a sizable metallic object, there is capacitance between the final tank coil and its associated link coil, and between the antenna tank coil and its link. Energy coupled through these capacitances travels over the link circuit and the transmission line as though these were merely single conductors. The tuned circuits simply act as masses of metal, and offer no selectivity at all for capacitance-coupled energy. Although the actual capacitances are small, they offer a very good coupling medium for frequencies in the vhf range. Therefore, harmonics and other spurious signals transferred from the tank by stray capacitance are not suppressed by an antenna coupler to the same extent as those transferred by pure inductive coupling.

11-442. Capacitance coupling can be reduced, with either single-ended or balanced tank circuits, by the grounding of the coupling coil to the chassis by a short, direct connection, as shown in figure 11-43. If the coil feeds a balanced line or link, it is preferable to ground its center, but if it feeds a coaxial line or link, one side may be grounded.

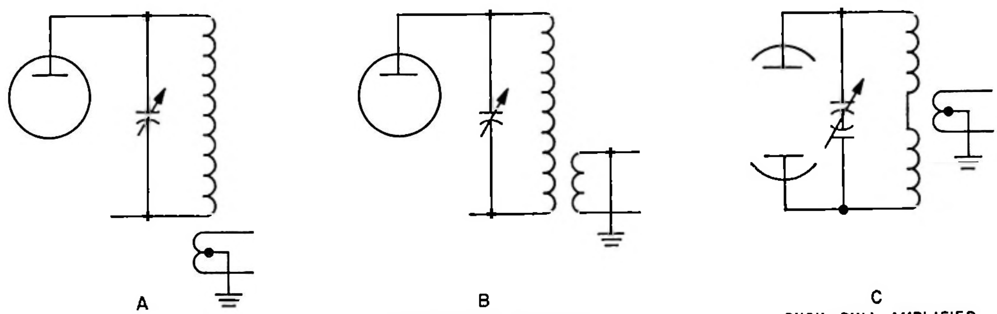


Figure 11-43. Suppression of Capacitance-Coupled Harmonics Grounding Antenna Coupling Coils

11-443. Capacitive coupling can also be reduced by using a shielded coupling coil. One way of obtaining the shielding at high frequencies is to form a one-turn coupling coil from a length of coaxial cable. The inner conductor of the coaxial cable is used to form a one-turn coupling coil. The outer conductor, which is grounded to the chassis, serves as an open-circuited shield around the turn. This shielding has no effect on the inductive coupling. Because this construction is suitable for one turn only, the coil is not well adapted for use on the lower frequencies, where many turns are required for good coupling. However, shielded coupling coils having a larger number of turns are available. A shielded coil is particularly useful with push-pull amplifiers when the suppression of even harmonics is important.

11-444. A shielded coupling coil or coaxial output will not prevent stray capacitive coupling to the antenna if harmonic currents can flow over the outside of the coaxial line. The proper way to use coaxial cable is to shield the transmitter completely and to make sure that the outer conductor of the cable is a continuation of the transmitter shielding. This prevents rf energy inside the transmitter from being emitted by any path except the inside of the cable. Harmonics flowing through a coaxial line can be stopped from reaching the antenna by an antenna coupler or by a low-pass filter installed in the line.

11-445. LOW-PASS FILTERS. A low-pass filter properly installed in a coaxial line, feeding either a matching circuit (antenna coupler) or the antenna directly, will provide high attenuation of harmonics. When the main transmission line is of the parallel-conductor type, the coaxial-coupled matching-circuit arrangement is highly recommended as a means for using a coaxial low-pass filter. A properly designed low-pass filter will not introduce appreciable power loss at the fundamental frequency if the coaxial line in which it is inserted is terminated so that the standing wave ratio is low. Such a filter has the property of passing, without loss, all frequencies below its "cutoff" frequency, but simultaneously has large attenuation for all frequencies above the cutoff frequency.

11-446. In order to provide the harmonic attenuation of which it is capable, a low-pass filter must be installed in such a way that all the output current of the transmitter flows through it. If harmonic currents are permitted to flow on the outside of the connecting coaxial cables, they will simply flow over the filter and on up to the antenna, and the filter will not have an opportunity to stop them. That is why it is so important to reduce the radiation from the transmitter and its leads to negligible proportions.

11-447. The proper way to install a filter between a shielded transmitter and a matching circuit is to form a continuous shield by connecting the filter to the transmitter by a shielded cable so as to keep all the rf energy inside. It is thus forced to flow through the filter, and the harmonics are attenuated. If there is no harmonic energy left after passing through the filter, shielding from that point on to the antenna is not necessary; consequently, the matching circuit or antenna coupler does not need to be shielded. If the antenna is driven through coaxial line, the matching circuit is unnecessary; in that case the line goes directly from the filter to the antenna.

11-448. When a filter does not seem to give the harmonic attenuation of which it is capable, the probable reason is that harmonics are bypassing it because of improper installation and inadequate transmitter shielding, including lead filtering. However, occasionally there are cases where the circuits formed by the cables and the apparatus to which they connect become resonant at a harmonic frequency. This greatly increases the harmonic output at that frequency. Such troubles can be completely overcome by substituting a slightly different cable length. The most critical length is that connecting the transmitter to the filter. Checking with a grid-dip meter at the final amplifier output coil will usually show whether an unfavorable resonance of this type exists.

11-449. NOISE (INTERFERENCE) AND FIELD STRENGTH MEASUREMENTS.

11-450. One of the many factors to be considered in planning high-frequency radio reception is the predicted circuit reliability. Satisfactory communication is possible only when the signal-to-noise ratio at the point of reception is sufficient to produce results capable of meeting the requirements for the type equipment in use on a given communication circuit (voice, telegraph, teletypewriter, etc). For this reason, the conditions existing at the area where reception of signals is contemplated are of utmost importance, as noise and interference become important limiting factors in determining circuit reliability.

11-451. Interference that is external to the receiving equipment may be broadly classified as either natural or man-made. Atmospheric interference, or static, caused by lightning discharges in local storms or arriving from distant tropical storms, and snow and dust static represent natural interference. Man-made interference is caused by many types of electrical installations, including engine ignition, power units of various types, power and other utility lines, rotating devices, etc. Interference caused by man-made installations may reach a receiver by any of the following means:

a. Direct radiation. Directly radiated interference is transmitted from the noise source to the receiving antenna, and does not travel any great distance; it is local to the area.

b. Conduction. Interference may be conducted directly from the source through the power lines to the receiver. Such interference generally occurs in ac-operated receivers in the form of a low-frequency hum or noise, or both.

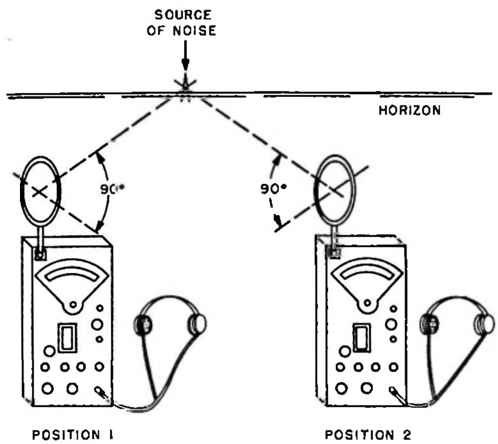


Figure 11-44. Locating Source of Noise by Triangulation

c. Reradiation. Reradiated interference is a common type of interference. Power lines, telephone and telegraph lines, and ungrounded metal structures conduct interference which has been transmitted to them from the noise source by direct radiation, by conduction, or by induction, and then reradiate the noise directly to the receiving antenna.

11-452. NOISE LOCATION TECHNIQUES.

11-453. GENERAL. Noise sources can be located by using a radio receiver or a field strength meter in conjunction with a loop, rod, or probe antenna.

11-454. LOOP ANTENNA METHOD. If you use the loop antenna method, the loop is turned so that maximum noise is heard in the output of the receiver. The loop is then turned until the noise output is minimum, indicating that the bearing of the noise source is 90 degrees from the direction of the plane of the loop. The azimuth bearing of the noise source should be plotted on a suitable map, noting that there is a possible 180-degree error in bearing at the moment. The noise-measuring instrument is then moved to a second position, and the loop is again turned for minimum noise. The azimuth bearing of the noise source should be plotted as before; the point of intersection of the lines plotted from both positions establishes the approximate location of the noise source, as illustrated in figure 11-44. In many cases the bearing plotted from a third position will be necessary in order to fix the location of the noise source more accurately. If the noise is determined to be local to the area, the source may be further localized by the signal intensity method, once the direction has been established as described.

11-455. ROD ANTENNA METHOD. A non-directional antenna, such as a vertical rod, can be used to determine the source of local interference. The use of an audio output meter in conjunction with a headset is advisable, since the human ear is not sensitive to small changes in noise level. The noise-measuring instrument is then moved toward the noise source, in the direction of the greatest noise intensity as indicated by the output meter. Since the output meter indicates all types of interference received within the effective bandwidth of the receiver, it is only by listening that the kind of interference and its probable source can be estimated.

11-456. PROBE ANTENNA METHOD. The probe antenna restricts the pickup of noise to that produced in the immediate vicinity of the probe. Therefore, you can use this device in close proximity to the actual noise source with much more accurate results than can be obtained by the previous two methods.

11-457. FIELD STRENGTH MEASUREMENTS.

11-458. GENERAL. Field strength is the effective value of the electric field intensity in microvolts or millivolts per meter, produced at the given point by radio waves from a particular station. The antenna radiation pattern (often called the directional characteristic) is the aggregate data collected from all points. The field strength produced by an antenna is proportional to the current flowing in it. Those parts of the antenna that carry the most current, as a result of standing waves, have the most radiating effect. (All resonant antennas have standing waves; only terminated types like the terminated rhombics have relatively uniform current distribution.) The power gain of an antenna is a measure of the directivity of the field pattern of the antenna under measurement, as compared with the field pattern of a "comparison" antenna (isotropic radiator) that radiates uniformly in all directions. It is the ratio of the power necessary to produce a given field strength at the given point by the comparison antenna to the power that must be radiated by the antenna under measurement to obtain the same field strength at the same point.

11-459. Field strength is measured in volts per meter. Since most field intensities are very small, it is convenient to employ the terms millivolts per meter or microvolts per meter. Thus, a 1-millivolt potential difference would exist between two points 1 meter apart in a 1-millivolt-per-meter field, assuming that the points lie in the direction of the greatest rate of potential change. Therefore, an antenna with an effective height of 5 meters, which is subjected to a field intensity of 20 millivolts per meter, would develop a 100-millivolt signal.

11-460. ANTENNA GAIN. The efficiency of a transmitting antenna is usually stated in such terms as power gain and field gain. These terms are based on the field intensity produced by a half-wave antenna in free space at a distance of one mile, under conditions for which there are no reflected waves. Power gain is then defined as the ratio of the power required by a half-wave antenna to produce a particular field strength at a distance one mile to the power required by the antenna under test to produce the same power, or, in equation form:

$$\text{Power gain} = \frac{\text{Power required with vertical half-wave antenna}}{\text{Power required with antenna under test}}$$

This ratio is usually given in db.

11-461. The antenna field gain of a high-frequency broadcast antenna is the ratio of the effective (free space) field intensity produced at one mile in the horizontal plane (vertical antenna radiating waves vertically polarized), expressed in millivolts per meter for 1 kilowatt antenna input power, to 137.6 mv/m (the field strength from a standard antenna under similar conditions). It is assumed in this definition that no waves are reflected from the earth or surrounding objects. Hence, the field gain of a multielement antenna expressed as a ratio is:

$$\text{Field gain} = \frac{\text{Field intensity of antenna at 1 mile for 1-kw input}}{\text{137.6 millivolts per meter}}$$

11-462. Since power is proportional to voltage squared, the relationship between field gain and power gain is expressed in the following equation:

$$\text{Field gain} = \sqrt{\text{Power gain}}$$

consequently a power increase of four times is required in order to double the signal intensity. If it is possible to increase the field intensity by using a multielement antenna rather than by increasing the power, then less expensive equipment can be employed and operating costs will be lower.

11-463. FIELD STRENGTH. The field strength of a radio wave is determined by measuring the voltage which the wave induces in a test antenna. The relationship between this induced voltage and the field strength can be obtained by calculation in the case of a loop or similar antenna, and by measuring the effective height of the antenna in other cases. (The effective height can be described as the calculated true electric height, corresponding to a perfect antenna that will produce the same field strength and is the ratio of the equivalent lumped induced voltage that can be thought of as acting in the antenna, divided by the field strength inducing this voltage.) There are two general methods for the determination of the field strength of a radio wave, the standard antenna method and the standard field generator method. The standard antenna method uses an antenna in which is known, by calculation or experiment, the amount of voltage that will be induced in the antenna by radio waves of known field strengths. The field strength is then determined by measurement of the induced voltage. The standard field generator method uses a standard signal generator to produce a radio wave of a known field intensity, which is then compared with the radio wave being measured; the field strength is then determined by the relationship between them.

11-464. The loop antenna is the most commonly used antenna for the measurement of field strength of radio waves below 40 mc. This is because its properties are predictable by calculation, it is directional, it is independent of ground constants, it can be designed for use with low-frequency measurements, and it is portable. A loop antenna is essentially a large coil of any convenient section, and extracts energy from the wave front as a result of phase differences between the voltages induced in the opposite legs. For example, when the plane of the loop is perpendicular at the direction of travel (ie, the wave front passes through the two legs at the same time), the voltages induced in the two vertical legs are equal and in the same phase, but, being directed around the loop in opporite directions, they cancel each other and result in zero response. As the plane of the loop is brought nearly parallel with the direction of wave travel, the wave front reaches the two vertical legs at slightly different times; this causes a phase difference between the voltages induced in those two legs, resulting in a voltage that circulates current around the loop; this voltage is maximum when the plane of the loop is parallel to the direction of wave travel. The induced voltage can be calculated for all shapes of loops by the formula below if the loop is small as compared with the wavelength, and if the frequency of the radio wave is no greater than one-third the self-resonant frequency of the loop.

$$\text{Induced voltage (E)} = 2E_oN \frac{\text{(loop area)}}{\text{wavelength}}$$

where Eo is the field strength of the radio wave in volts per unit length

N is the number of turns of the loop.

The effective height of the antenna is the ratio of the induced voltage (E) to the field strength (Eo).

11-465. The loop size that would be required for measurements above 40 mc is too small for use without complication; for this reason, loops are rarely used for measurements above this frequency. A half-wave dipole is a more useful antenna for these measurements. Other suitable alternates to the loop antenna are the doublet, which is preferable to the dipole at frequencies where the dipole would be too large, and the short grounded vertical antenna, which is best used for the evaluation of weaker fields. However, the vertical antenna has the disadvantage that its effective height must be determined for each particular installation by experiment.

11-466. The induced voltage for a half-wave dipole can be closely approximated from the formula,

$$E = \frac{\text{wavelength x Eo}}{\pi}$$

and the effective height from the formula,

$$h = \frac{\text{wavelength}}{\pi}$$

For the doublet, the induced voltage can be determined by the formula,

$$E = \frac{EoL}{2},$$

provided that the radio wave is no greater than one-third the resonant frequency of the doublet. Its effective height is $\frac{L}{2}$, where L is the length of the doublet. The voltage induced into a short grounded vertical antenna can be found by the same formula used for the doublet $E = \frac{EoL}{2}$ if the antenna is considerably shorter than one-quarter wavelength and if L is the length of the grounded antenna. The effective height, as previously mentioned, is best determined by experimentation. If the vertical antenna is exactly one-quarter wavelength long, the induced voltage may be computed by the formula,

$$E = \frac{\text{wavelength x Eo}}{2}$$

where Eo is the field strength in volts per unit length.

11-467. The standard antenna employed for use with microwave applications usually consists of a high-gain antenna, such as a horn or parabola, having a known power gain. When the antenna is matched to a load impedance so that the load absorbs maximum possible power from the antenna, the field strength can be calculated from the following equation

$$Eo = \sqrt{\frac{480\,\pi^2 P_r}{\lambda^2 G}}$$

where Eo is the field strength

P_r is the load power

λ is the wavelength

G is the power gain of the antenna relative to an isotropic radiator

11-468. STANDARD ANTENNA METHOD. The usual method of measuring the field strength with a standard antenna makes use of a field strength meter incorporating a superheterodyne receiver, as shown in figure 8-147. The receiver has an adjustable attenuator in the intermediate frequency amplifier for adjusting the gain to the receiver in accurately known increments. This method of measuring field strength is essentially a comparison method in which the signal induced in the antenna is compared with a signal of the same frequency, which is induced into the antenna by an auxiliary signal generator. (The auxiliary, or calibrating, voltage source is often built into the receiver.) The ratio of the two voltages is obtained by the attenuator settings that are required to maintain a constant output for both signals. The accuracy of this method of measurement is high, since the receiver characteristics (assuming that the mixer is linear) do not affect the measurement.

11-469. When portability of the instrument is a consideration, the loop antenna is commonly used to develop the input rf signal. The use of a loop antenna ordinarily leads to input arrangements like the circuit shown in figure 8-147. The output of the auxiliary oscillator can be coupled to the loop by means of a transformer, as shown, or applied directly across a resistor in series with the loop. The voltage transfer of the transformer is assumed to be a 1:1 ratio for the sake of simplicity. The local oscillator must be arranged and coupled to the input circuit so that its output is not varied as a result of tuning. Either the input circuit can be metered by an electronic voltmeter, or the plate circuit of the mixer can be metered by a milliammeter calibrated to indicate a 1-volt signal at the grid. A calibrated attenuator precedes the i-f amplifier. This placement of the attenuator avoids saturation of any of the i-f stages. Care must be used to prevent overloading of the mixer stage. A detector incorporating a metering circuit follows the i-f amplifiers. The steps listed below describe a typical field-strength test.

a. Tune in the signal, and adjust the attenuator to provide a convenient deflection of meter M2. Record this attenuator setting, which shall be referred to as A1.

b. Turn on the auxiliary oscillator, and set it to the frequency of the incoming signal.

c. Adjust the output of the auxiliary oscillator so that a preselected amplitude of signal, usually 1 volt, is indicated at the grid of the mixer.

d. Readjust the attenuator so as to make the deflection of meter M2 the same as that of step 2. Record this attenuator setting, which shall be called A2.

NOTE

It would appear that the signal induced in the loop is (A2 — A1) db below the oscillator voltage measured at the grid of the mixer. This is not quite true, however, since there is a resonant rise of voltage

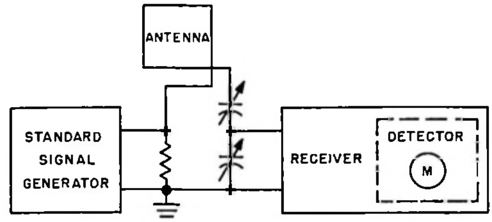


Figure 11-45. Substitution Method of Determining Field Strength

in the loop. As a matter of fact, rf amplifying stages are sometimes placed between the loop and the mixer grid. It is convenient to refer to the ratio of the actual signal present at the grid of the mixer to the received loop signal voltage as the voltage transfer ratio. In order to make proper allowance for this ratio, the following steps should be performed.

e. Remove the auxiliary oscillator from the coupling transformer, and connect it directly to the grid circuit. Do not change the output of the oscillator from the setting arrived at in step c.

f. Readjust the attenuator setting so that meter M2 again shows the same deflection as that obtained in step a. Call this attenuation A3.

g. Compute the quantity (2A2 — A1 — A3). The loop signal voltage is below the voltage measured in step c by this amount in db.

11-470. SUBSTITUTION METHOD. The substitution method of determining field intensity makes use of a standard signal generator and a sensitive receiver connected to a standard antenna, as shown in figure 11-45. The receiver is tuned to the unknown signal frequency and the antenna is oriented for maximum response, as indicated on the microammeter connected in the circuit of the second detector. The receiver gain is adjusted for a convenient reading on the meter. The antenna is then tuned for a null, where little or no signal is received, and the signal generator is turned on and set for the signal frequency, thereby inducing a voltage in series with the antenna (a loop in the case illustrated) by means of a 1- or 2-ohm resistor. The output of the signal generator is then adjusted until the receiver meter indicates the same reading as previously obtained from the induced signal. The induced voltage is equal to the output voltage of the signal generator.

11-471. This method of measurement has the disadvantage that the signal generator must be well shielded so that no stray radiation is picked up by either the antenna or the receiver. The substitution method is well adapted to long-wave measurements, and the accuracy of this method is comparable to the adjustable i-f attenuator method.

11-472. STANDARD FIELD GENERATOR METHOD. This method of measurement is widely used for measurements at frequencies above 30 mc. The standard field generator is a compact portable oscillator having a small loop (or other type) antenna provided with a thermocouple meter to measure the antenna current. The antenna must be of such a design that when a known quantity of current or power is transferred to the antenna, the intensity of the field can be calculated from the dimension and construction of the antenna. It is thereby possible to compare the standard field produced with the unknown field, and to calibrate the intermediate-frequency attenuator type of measuring equipment for the higher frequencies.

11-473. Standard antennas and the formulas for calculating the free space values of field intensity (in the direction of maximum radiated field) of each are as follows:

Loop:

$$E_o = \frac{120\,\pi^2 NAI_a}{d\,\lambda^2}$$

Short vertical antenna carrying uniform current:

$$E_o = \frac{60\,\pi\, hI_a}{d\,\lambda}$$

Half-wave dipole:

$$E_o = \frac{60I_a}{d} = \frac{7.02}{d}\sqrt{P_a}$$

Directional antenna:

$$E_o = \frac{5.48}{d}\sqrt{P_a G}$$

In these equations,

E_o = field strength, volts per meter, at distance d

d = distance in meters

λ = wavelength in meters

h = height of antenna, assumed less than $\lambda/10$

N = number of turns in loop

A = area of loop, sq meters

P_a = power radiated by antenna in watts

I_a = current flowing in antenna in amperes

G = gain of antenna over isotrophic radiator

11-474. RELATIVE FIELD STRENGTH MEASUREMENTS.

11-475. GENERAL. There are many occasions when it is only necessary to measure relative field strength, such as when plotting an antenna radiation pattern or just checking the operation of a transmitter. Relative field strength measuring equipments consisting only of a simple tunable receiver or circuit, are available for this type of measurement. These equipments are calibrated in terms of relative voltages, relative power, or decibels referred to a particular reference level. The equipments are direct-reading, and require only proper tuning and antenna orientation for comparative readings.

11-476. GRID-DIP METER METHOD. It is possible to use the grid-dip meter as a relative field-strength meter. Refer to paragraph 8-741 for a discussion concerning grid-dip meters. For the grid-dip meter to perform the measurement mentioned above, the plate voltage must be turned off, and a loop antenna must be connected to the coil terminals of the instrument. The appropriate plug-in coil is inserted, and the meter tuned to the transmission. If the received signal is sufficiently strong, current will flow in the grid-cathode circuit. The relative magnitude of the field is indicated by the amount of meter deflection. Since the meter deflection may not be linear, the meter should be calibrated if accurate indications are required.

11-477. SIMPLE METER APPLICATION. Simple instruments intended specifically for the measurement of field intensity usually employ a crystal diode connected into a circuit like the one shown in figure 8-146. A whip antenna is often used as the pickup antenna. To increase the sensitivity, it is advisable to employ a microammeter as the indicating device. The use of plug-in coils extends the instrument operation over a wide frequency range.

11-478. ADVANCED METER APPLICATION. When greater sensitivity is necessary, a more elaborate field-strength meter is required. One technique is to employ a specially designed receiver, which has an attenuator calibrated in db at the input of its first i-f amplifier. The output of the mixing stage must be exactly proportional to the rf signal voltage present in the grid circuit, and this property must hold for input amplitudes up to at least 1 volt. This type of measuring device is shown in figure 8-147.

11-479. ANTENNA RADIATION PATTERN.

11-480. The antenna radiation pattern (field strength pattern), when used for radio transmission, can be defined as the relative radiation that the antenna produces in different directions. For reception the pattern is the same, but represents the relative response of the antenna to radio waves arriving from different directions. The various radiation properties of an antenna apply both to transmitting and receiving applications, subject to certain qualifications when the path of the waves between the transmitting and receiving points involves propagation through the ionosphere. Thus, the more effective the antenna for transmitting, the more effective it will be for receiving. Also, the directive properties will be the same for transmission and reception, and in the case of directive antennas the gain will be the same for both transmitted and received signals. For this reason, antennas are said to follow the theorem of reciprocity. Because of antenna receprocity, the directional characteristics of an antenna can be determined either by radiating power from the antenna and measuring the field-strength distribution that results, or by measuring the voltage induced in the antenna as a portable transmitter is moved about it (or as the antenna is rotated with the transmitter stationary).

The former arrangement is usually considered the more convenient. In performing these measurements, the distance between the transmitter and the receiver should be at least two to three wavelengths, to reduce any error introduced from induction fields to a minimum. For the high frequencies, a complete pattern measurement requires the use of an airplane. Measurements taken at the ground level have limited significance, since only the field-strength distribution in the horizontal plane of the vertically polarized component of the ground wave may be measured; horizontally polarized antennas at these frequencies radiate virtually no field strength along the earth's surface.

11-481. The free-space field patterns for most common antennas have been calculated and are readily available from many sources. These patterns can be of great assistance in making field intensity measurements, since a complete pattern plotted experimentally requires an enormous number of readings. Usually, it is only necessary to take readings where it is expected that a deviation will occur from the free-space pattern. The usual procedure for plotting the pattern is to show the radiation as a function of aximuth angle for different values of vertical angle, or vice versa. In the case of highly directional antennas, the most important characteristics are the shape and width of the main lobe, and the magnitudes and directions of the side lobes.

11-482. The simple beam width of a microwave parabolic antenna can be determined approximately by calculation. Because of the tremendous gain of a parabolic antenna at the microwave frequencies, the information obtained from this calculation is all that is necessary for most applications. The calculation is given in paragraph 11-787.

SECTION III

RADAR EQUIPMENT TESTING

11-483. GENERAL.

11-484. It is possible to determine to some degree the satisfactory operation of a radio receiver by listening to its output and observing how much the gain control must be advanced to obtain sufficient volume. However, unlike most radio communication equipments, satisfactory performance of radar sets cannot be determined by observation alone.

11-485. In the field of radar, reliance cannot be placed solely on visual observation to judge the range capability and data accuracy of radar systems. It has been found to be so inaccurate as to be completely valueless. Numerous tests made on radar equipments in the field strongly emphasize this fact. In one instance, the performance of approximately 100 different radar sets was carefully measured with test equipment of known accuracy. In each case, the radar set under test was considered to be in normal operating condition by the radar personnel concerned. The results of these tests revealed that, on the average, the maximum effective range of the radar equipments under test was only one half the maximum range possible, had the equipments been operating at peak efficiency. In fact, five radar sets were found to be operating at less than 10 percent of their possible maximum range, which means in effect, that these radar equipments were protecting only 1 percent of their assigned tactical areas. Since such poor performance, as demonstrated by these tests, may have serious consequences, it can readily be seen that performance testing is of the utmost importance in radar work.

11-486. Investigation to find the cause of this unsatisfactory situation showed that many technicians are not sufficiently familiar with the techniques and procedures necessary to test microwave-radar systems properly. In this section an effort is made to remedy this situation. It is assumed that the reader is already familiar with the basic principles of radar. By way of review, a brief discussion of the functional requirements of the range-determining components of a radar set is included below.

11-487. RADAR FACILITY FUNDAMENTALS.

11-488. The operational requirements of the components of typical pulse radar facilities are briefly described here. For purposes of study, the typical radar facility can be reduced to the following functional components: synchronizer (modulator), transmitter, antenna, receiver, and indicator. Because power supplies differ so greatly in different radar sets, they will not be considered in this general discussion. While the complexity of each of the above components may vary considerably, depending on the use for which each particular radar equipment was designed (ie, search, navigation, fire control, etc), basically the function of each type of unit is identical in all radars. It is not the purpose of this discussion to describe any specific radar set, but rather to deal with the functional requirements of each unit necessary for the efficient over-all performance of any radar facility.

11-489. SYNCHRONIZER.

11-490. Figure 11-46 illustrates the typical timing requirements of a radar facility which are developed and supplied by the synchronizing component. The timing circuits may all be located in one unit, or as is often the case, they may be distributed throughout one or more additional components, such as the transmitter, indicator, or receiver. In an externally synchronized equipment, a trigger pulse is fed to the transmitter for the purpose of timing the firing of the rf oscillator with the rest of the radar facility. In some sets, a gate pulse is also fed to the receiver in order to turn it on immediately after the transmitter fires. When the rf performance of the transmitter and receiver is measured, some of the test equipments used must be synchronized with the radar facility. To supply the necessary timing signal for these test equipments, test receptacles are usually located conveniently on these units which provide sync voltages of the proper amplitude and polarity.

11-491. TRANSMITTER.

11-492. The transmitter consists of an rf generator and a modulator, or pulsing component. In microwave applications, the generator is usually a magnetron, because of its relatively high output power. In lower-frequency applications a ring or conventional type oscillator is used. For satisfactory operation of the transmitter, a nearly rectangular modulator pulse is required. A pulse with a flat top is desired because a magnetron tends to shift frequency if its high voltage (furnished by the modulating pulse) varies during the period of oscillation. A steep leading edge is also required, particularly in fire-control equipment, where accurate range data is necessary. In facilities where minimum range data is needed, a pulse with a steep trailing edge is essential.

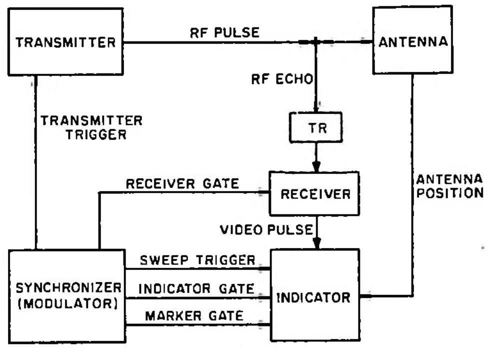


Figure 11-46. Radar Facility Showing Timing Data Supplied by the Synchronizer

11-493. The required width, or duration, of the pulse also depends upon the type of radar system in which it is to be used. In the case of long range air-search systems, wide pulses (2 to 20 microseconds) are utilized in order to maintain a high average of transmitted power. For surface-search and fire-control equipments, which require high resolution, narrow pulses (0.1 to 2 microseconds) are used.

11-494. Any sizable variation from the norm in pulse width, shape, or amplitude seriously affects the range capability and accuracy of the facility. These undesired pulse variations are not necessarily apparent to a radar operator. For this reason, test methods and procedures have become incorporated in the preventive maintenance program to observe and measure transmitter operation as it affects the overall performance of a radar facility.

11-495. ANTENNA.

11-496. A typical antenna group consists of such components as the transmission or rf lines, with associated accessories such as rotating joints, duplexer, slotted-line section and directional coupler, and the antenna and antenna reflector, some of which are shown in figure 8-132. Although the rf lines may be either coaxial cable or wave guides, their function is the same, that is, to deliver the maximum amount of power from the transmitter to the antenna and to receive power from the antenna to the receiver. Many factors tend to reduce the efficiency of rf lines. Improper coupling, faulty rotating joints, dents in the line, or poor solder joints may produce an impedance mismatch between the rf lines and the antenna or transmitter, resulting in a high standing-wave ratio, magnetron instability, and a reduction in facility performance.

11-497. Since radar facilities both transmit and receive through the same antenna, some method must be employed to block the receiver during transmission. The most common device used is the duplexer, which consists of a transmit-receive (tr) switch, and usually an anti-transmit-receive (atr) switch. The tr switch is a gas-filled tube which is designed to short circuit across its spark gap each time the transmitter fires, thus preventing saturation of the receiver, and damage to the crystal mixer (when used) by a strong signal.

11-498. To prevent loss of the return signal in the transmitter during reception, many radar facilities incorporate an atr switch, located an odd number of quarter wave lengths from the receiver T-junction. During transmission, both the tr and atr gaps are fired, producing short circuits at these points. In this way transmitter power is conducted to the antenna without loss. During reception neither is fired, and the impedance relations of the atr switch are such that the received echo is reflected into the receiver with minimum loss. In addition to this, the tr and atr switches must be capable of quick recovery after the transmitter fires, so that echoes from nearby targets may be received. Recovery times of tr tubes range from 3 to 20 microseconds. Fire-control radar facilities require tubes with longer recovery periods. Since the recovery time of these tubes tends to increase with use, it must be checked periodically to ensure proper operation of the radar facility. Methods of measuring recovery time will be discussed later.

11-499. Early radar antennas often required considerable adjustment for efficient operation, but most present-day antennas are preset at the factory, and require no further adjustments in the field. However, when making rf tests or measurements, it is important to position the antenna so that it will not be affected by strong echoes from nearby fixed targets.

11-500. To provide efficient rf test points, two devices—directional couplers and slotted-line sections—are used. In some equipments these devices are permanently built into the antenna group. In others they are included as part of the test equipment, and may be inserted in the rf line whenever tests are to be made.

11-501. RECEIVER.

11-502. An efficient receiver is one having good sensitivity, short recovery time, and sufficient bandwidth to pass a received pulse echo without undue distortion. Figure 11-47 shows the necessary stages required to obtain these. Because of the high frequencies at which most radar receivers operate, the received signal is fed directly to a crystal mixer or rf amplifier. The local oscillator and preamplifier stages are normally separated from the receiver sec-

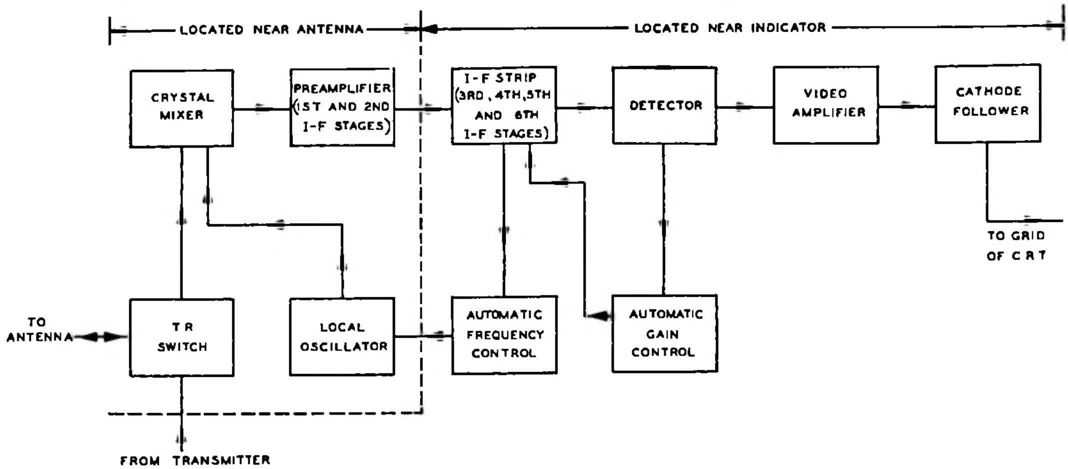


Figure 11-47. Microwave Receiver

tion and are located close to the mixer or rf amplifier which is often near the antenna. A silicon crystal is usually used as a mixer because of its low noise level. Two preamplifier i-f stages follow immediately to provide a signal boost so that the i-f output signal from the crystal mixer will not be lost because of attenuation in the transmission line, before it reaches the remotely located receiver unit.

11-503. Although the tr switch is designed to protect the receiver each time the transmitter fires, a strong signal from the transmitter may leak through. Unless additional precautions are incorporated, this undesired transmitter pulse may overdrive and block the receiver, rendering it insensitive to signals reflected from nearby targets. This blocking usually occurs in one of the resistance-coupled video stages. Several methods are used to minimize this undesirable effect, such as using a receiver gate pulse, or feeding a negative-going signal from the second detector to the first video stage.

11-504. The bandwidth requirements of the receiver also depend on the type of radar facility in which it is used. Fire-control radar facilities require broad-band receivers for accurate range data. Search facilities, on the other hand, operate satisfactorily with narrow-band receivers, since merely the presence and approximate range of a target are usually all the data required.

11-505. INDICATOR.

11-506. The indicator performs the important function of transforming electrical information gathered by the radar into a visual presentation on the face of one or more cathode-ray tubes which are part of the component. If the facility includes an A scope or a ppi scope, visual observation of receiver performance, as indicated by echo box "ring time," may be had, as explained later in the text.

11-507. It should be emphasized that while the display on the A scope may indicate, to the inexperienced operator, that all the components of a radar facility are operating, it will not show how efficiently they are performing. The degree of efficiency can be determined only by careful performance testing.

11-508. FREQUENCY MEASUREMENT.

11-509. The measurement of frequencies employed in radar operation falls into two general categories: transmitter frequency and receiver frequency.

11-510. A range of frequencies is assigned for transmitter operation of any given radar facility. Transmitter operation is restricted to a certain range of frequencies for several reasons: First, radar beacon stations, which are assigned to each radar band, respond only to signals within a given frequency range. Second, the associated waveguide tuning adjustments cover only a limited range of frequencies. Third, interference between radar facilities used for different types of services could result if all radar equipments were permitted to operate in the same band. For this reason, airborne S-band and ground S-band radar equipments usually operate in different parts of the band.

11-511. Testing the radar receiver frequency consists of measuring the frequency at which the receiver operates most efficiently, or of measuring the local-oscillator frequency. For radar reception, a knowledge of the receiver frequency is not important as long as the receiver is carefully tuned to the transmitter frequency. In receivers using afc, the local oscillator must, of course, be operated either above or below the signal frequency in accordance with the design specifications, but here again a knowledge of the exact frequency is not very important. However, in beacon reception, a knowledge of the exact receiver frequency is often necessary in order that the receiver may be accurately tuned (in the absence of a beacon signal) to the beacon-signal frequency, which is different from that of the radar-transmitter frequency. In beacon reception it is important that the receiver bandpass be centered about the frequency used for interrogating the beacon. The bandwidth of a beacon receiver is an important factor and should be checked along with the beacon receiver frequency. This measurement is often made with test equipment incorporated in the radar facility.

11-512. FREQUENCY TESTING COROLLARY DATA.

11-513. GENERAL. The following subjects are pertinent to the measurement of radar frequencies: frequency testing standards, frequency coupling methods, test equipments commonly used in radar frequency testing, and the accuracy limitations of these instruments.

11-514. FREQUENCY TESTING STANDARDS. Equipments employed in frequency testing are classified as either primary or secondary standards. For a detailed discussion concerning frequency standards, refer to paragraph heading 8-320. To summarize briefly, a primary standard provides an extremely accurate frequency source which is checked against the rotation of the earth, and is used to calibrate secondary standards. Secondary standards are extremely accurate and are not unnecessarily cumbersome or difficult to use. Consequently, for radar maintenance, accurate test equipments comparable to secondary standards are used.

11-515. METHODS OF COUPLING FREQUENCY STANDARDS. Frequency testing instruments are not designed to be coupled directly into the radar equipment, since the high-power transmitter pulse develops very high voltage within resonant circuits associated with the instruments and arcing would result. Satisfactory methods of frequency coupling are described for power sampling techniques and are discussed under paragraph heading 11-548.

11-516. FREQUENCY TESTING EQUIPMENTS. Two test equipments are satisfactorily used in the measurement of microwave frequencies. They are the resonant-coaxial-line frequency meter and the resonant-cavity frequency meter. Both test instruments depend upon a condition of resonance to provide an accurate test indication.

11-517. RESONANT-COAXIAL-LINE FREQUENCY METER. This type of frequency meter is discussed in paragraph 8-365, and is illustrated in figure 8-67. To summarize, coaxial-line frequency meters may be connected to operate as either transmission or reaction type indicators. When used as the transmission type, energy is fed into one coupling loop and the indicating device is connected to the other loop. When the circuit is resonant, the greatest energy transfer takes place and the indicator shows the greatest output signal. When used as the reaction type, the resonant circuit functions as an absorption device, so that at resonance the indicator shows a dip in the reading.

11-518. RESONANT-CAVITY FREQUENCY METER. Figure 8-65 shows a common type of resonant-cavity frequency meter, which consists essentially of a hollow metal cylinder coupled to a waveguide by means of a small hole or a coupling loop and coaxial connector. The cavity within the cylinder is resonant by virtue of its dimensions. Two end plates may be thought of as the capacitance elements, and the adjoining walls as the inductance. Frequency is varied by adjusting the position of one of the end plates with a micrometer screw, which is calibrated to indicate frequency. The degree of accuracy obtained with this type of meter is greater than that obtained with the resonant-coaxial-line frequency meter.

11-519. FACTORS AFFECTING MEASUREMENT ACCURACY. The accuracy of a microwave frequency measurement is expressed in terms of maximum error in megacycles, and may be either absolute or relative. Absolute accuracy states how much error with respect to a standard (usually WWV) is involved in a single frequency measurement, whereas relative accuracy indicates how much error is involved in the difference in frequency (or increment) between two microwave signals. For example, assume a measurement accuracy of ± 4 mc absolute and ± 1 mc relative; if the frequency of a certain transmitter measures 9300 mc and the local oscillator frequency is 9330 mc, the following conclusions can be reached:

a. The transmitter frequency is somewhere between 9304 mc and 9296 mc (or 9300 mc ± 4 mc absolute).

b. The local-oscillator frequency is somewhere between 9334 mc and 9326 mc (or 9330 mc ± 4 mc absolute).

c. The local-oscillator frequency is 29 mc to 31 mc above the transmitter frequency (9330 mc minus 9300 mc or 30 mc ± 1 mc relative).

Thus, it can be seen that the difference in frequency between two measured values is much more accurate than the measured values themselves.

11-520. In beacon-receiver frequency testing, an absolute accuracy of ± 4 mc is not considered good enough. In order to obtain the required accuracy, manufacturers of frequency testing equipments carefully calibrate by hand that part of the test equipment which concerns beacon frequency testing, with the result that these equipments have an absolute beacon-frequency accuracy better than ± 1 mc in the X band.

11-521. Most of the early frequency meters were tuned by means of a micrometer screw, and the readings were converted into frequency with the aid of a calibration chart. Some of the newer type meters make use of a dial, geared to the screw, which indicates frequency directly. Thus dial readings are greatly simplified, but the gear mechanism associated with the dial introduces a certain amount of backlash which affects the accuracy of the indication. The backlash effect may be minimized by always approaching the final dial setting from the same direction. This direction should be the same as that used during factory calibration, and should be specified in the instructional literature accompanying the test equipment.

11-522. Atmospheric conditions of temperature, relative humidity, and atmospheric pressure have an appreciable effect upon the accuracy of a frequency meter. This effect is minimized by constructing the resonant sections of materials that minimize or compensate for changes in temperature, and by hermetically sealing the units against moisture, with the result that sufficient accuracy is obtained for most applications. Where extreme accuracy is necessary, the correction charts for varying atmospheric conditions (supplied with the equipment) must be used. Atmospheric pressure variations, under normal conditions, are not great enough to require compensation; however, the effect of reduced pressure at high altitudes can be appreciable. It should be emphasized that the conditions discussed above are the conditions existing inside the frequency meter rather than the conditions outside it.

11-523. TRANSMITTER FREQUENCY TESTING.

11-524. GENERAL. Radar transmitters may be either fixed frequency or tunable. If the frequency of a fixed-frequency transmitter is measured and found to be outside the operating band, the magnetron or the defective component of the magnetron assembly must be replaced. Tunable transmitters may be adjusted throughout the operating band; this is a desirable feature when several radars are in use in a limited area, since the radar may be tuned to a particular frequency to prevent or avoid interference. In addition, when jamming signals are present, a change in radar frequency may be effective in eliminating or reducing their effect. Since the operating bands are fairly wide, however, transmitter frequency tests do not require extreme accuracy.

11-525. REACTION-TYPE INDICATION METHOD. An early method of frequency measurement is shown in figure 11-48. The meter absorbs power from the crystal detector at resonance; thus, a reaction type indication is obtained. Figure 11-49 shows the appearance of the rf envelope as the frequency-meter tuning is varied. Resonance is obtained when the center of the pulse reaches its lowest point, indicating maximum reaction. If desired, a microammeter may be used instead of the synchroscope; for this instrument, the frequency meter is adjusted for a dip in the current reading.

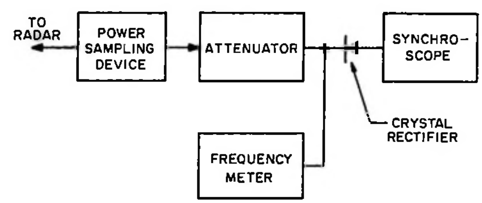


Figure 11-48. Frequency Measurement, Reaction-Type Indication

11-526. TRANSMISSION-TYPE INDICATION METHOD. Figure 11-50 shows a widely used test method for frequency measurement, in which the transmission type indication is used. The procedure for this method is as follows:

a. The equipment is connected as shown in figure 11-50 and the power sample is coupled into the frequency meter.

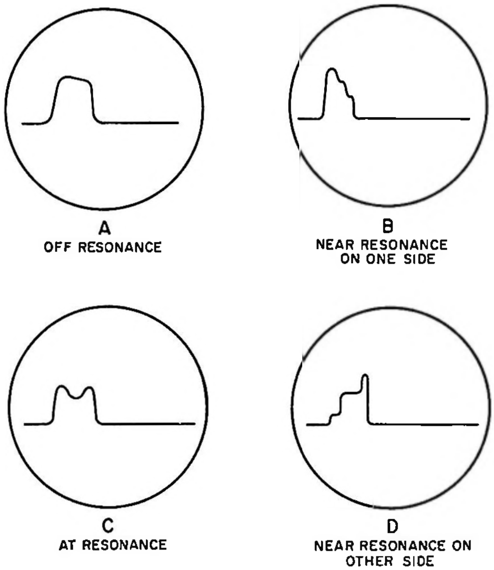


Figure 11-49. Change of Waveform Observed During Frequency Measurement

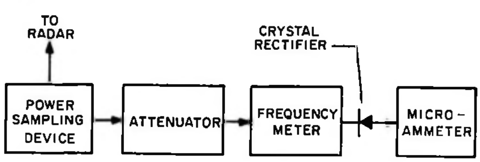


Figure 11-50. Frequency Measurement, Transmission-Type Indication

b. The measurement is started with maximum attenuation and the frequency meter is tuned through the frequency range.

c. If no indication is observed, the attenuation is reduced about 10 db and the frequency control is tuned through the frequency range again. This process is repeated until a reading is observed.

d. The frequency dial is set for maxi-

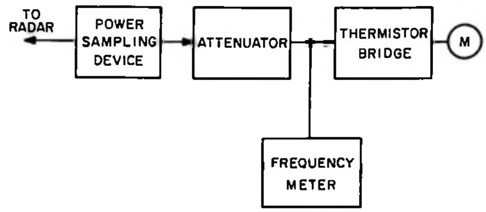


Figure 11-51. Combination Power and Frequency Measurement

mum reading with sufficient attenuation to keep reading below full-scale value.

e. Ordinarily it is necessary to convert the dial reading to frequency.

11-527. COMBINATION POWER AND FREQUENCY TESTING. In modern power testing equipment, a frequency meter is often included as an integral part of that equipment. The frequency meter is usually connected as shown in figure 11-51. For power testing, the frequency meter must be tuned off resonance so as not to affect the accuracy of the power measurement. The test procedure is very simple and is usually performed directly after a power measurement. Frequency measurement is performed as follows:

a. A 1-mw reading is established as discussed in the power testing procedure.

b. The frequency meter is tuned for minimum meter reading. (The meter must be tuned slowly or the resonance point may be passed before the thermistor can respond.)

c. Ordinarily it is necessary to convert the dial reading to frequency.

11-528. RECEIVER FREQUENCY TESTING.

11-529. GENERAL. For most types of radar equipment the receiver is tuned to the

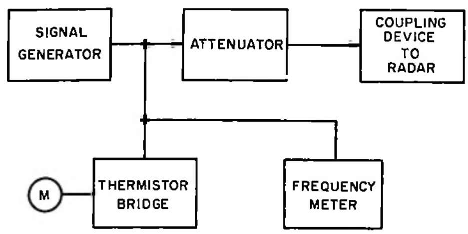


Figure 11-52. Receiver Frequency Measurement

transmitter frequency, and it is not necessary to make a receiver frequency test. However, for beacon operation, the receiver must be accurately tuned to a specified frequency and, in the absence of a beacon signal, the receiver frequency may have to be determined.

11-530. TEST PROCEDURE. The test equipment is connected as shown in figure 11-52; the receiver frequency measurement is performed as follows:

a. The frequency of the signal generator is tuned to the receiver center frequency by observing the receiver output on either the radar indicator or a synchroscope. The resonant point is indicated by maximum receiver output.

b. The frequency meter is tuned to the frequency of the signal generator. This is indicated on the radar indicator or the synchroscope by a dip in output.

c. Ordinarily it is necessary to convert the frequency meter dial reading to frequency. This frequency is the receiver center frequency.

11-531. ALTERNATE TEST PROCEDURE. In some cases, such as in beacon operation, it may be necessary to set the frequency of the receiver to some predetermined value. For these cases, the test equipment is converted as shown in figure 11-52; the procedures are performed as follows:

a. The frequency meter is adjusted to the desired frequency.

b. The signal generator is tuned for maximum dip as indicated by the thermistor bridge.

c. This dip is checked by varying the frequency of the frequency meter to determine if it is caused by the meter.

d. The receiver is tuned for maximum output on an output indicator. (A synchroscope may be connected to the receiver output to serve as an indicator.)

11-532. LOCAL-OSCILLATOR FREQUENCY MEASUREMENT. The local-oscillator frequency can be measured by feeding the output of the local oscillator, directly if possible, to the frequency meter and making the test previously described. If desired, the local oscillator may be set to some predetermined frequency by setting the frequency meter above or below (as specified in instructional or maintenance literature) the frequency to be received by an amount equal to the intermediate frequency, and tuning the local oscillator for the required indication. This method is especially usefor for a radar facility employing afc, which requires that the local-oscillator frequency be set on a certain side of the signal frequency. For manually tuned radars, either side works well as far as receiver performance is concerned.

11-533. POWER MEASUREMENTS.

11-534. When testing radar facilities, it is often necessary to measure the power output of the radar transmitter, or to determine the output level of a signal generator so that the test equipment can be used to make accurate measurements on the radar receiver. It is important, therefore, that the technician have a thorough knowledge of the principles involved in testing power output.

11-535. Modern testing methods require that the absolute power in watts be the unit of measurement. Power can be measured in terms of relative values, but for most purposes such measurements are considered unsatisfactory. For example, a crystal rectifier and dc meter may be used to indicate power in units of meter deflection instead of watts. If periodic tests were made on a certain type of equipments, with the same crystal and meter, using the same procedure, any change in power would probably be discovered. However, this method has several faults. First, there is a danger that the initial reading, which must serve as a reference for comparison with later readings, may be taken at a time when the equipment is not operating properly. Second, manufacturers' specifications are rated in watts, rather than in relative values. Third, the results of this method cannot be compared with the results obtained by another person or by means of other test methods. Fourth, a different crystal and meter combination is very likely to give a different reading, so that if either the meter or crystal were damaged, the test procedures might have to be started all over again. It is clear, therefore, that accurate, calibrated test standards must be used if maximum performance and maintenance efficiency are desired.

11-536. POWER TESTING DATA.

11-537. GENERAL. The material under this heading is included to provide basic information which is necessary to understand microwave power measurements and the techniques involved. The subjects covered are pulse power and average pulse power, the decibel and its use, power sampling methods, attenuators, and a brief description of the thermistor.

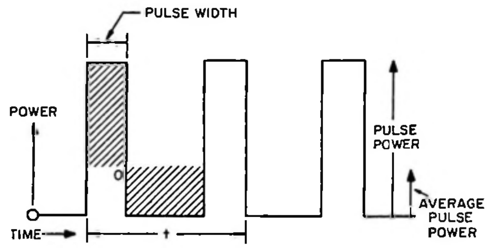


Figure 11-53. Transmitting Pulses, Showing Peak and Average Power

11-538. PULSE POWER AND AVERAGE PULSE POWER. Power measurements are classified as either pulse power or average pulse power. The actual transmitter output occurs at peak level, but most modern test methods measure the heating value of the rf energy, to obtain the average value. It is correct to use either value for reference so long as one or the other is consistently used. Frequently it is necessary to convert from pulse power to average pulse power, or vice versa; therefore, the relationship between the two must be understood. Figure 11-53 shows the comparison between pulse power and average pulse power when a square pulse is used. The average value, which represents the actual heating value of the pulses, is located at a point somewhere between zero and peak. The level of the average value is defined as that level where the pulse area above average equals the area below average between pulses. If the pulses are evened off in such a way as to fill in the space between pulses, the level obtained is the average value, as shown in figure 11-53, where the shaded area of the pulse is used to fill in the space between pulses. In the same figure, the area of the pulse is equal to pulse width multiplied by pulse power, and the area of the average value is equal to average pulse power multiplied by the pulse period (T). Since the two values are equal, it is permissible to express the equation as follows:

$$\text{Pulse width x pulse power} = \text{Average pulse power x T}$$

Transposing terms in the equation produces:

$$\frac{\text{Average pulse power}}{\text{Pulse power}} = \frac{\text{Pulse width}}{\text{T}}$$

and since T = 1/prf
then:

$$\frac{\text{Average pulse power}}{\text{Pulse power}} = \text{Pulse width x prf}$$

$$= \text{Duty cycle}$$

The ratio of average pulse power to pulse power is called the duty cycle, and represents the time the transmitter is on, each second. Duty cycle is simply a numerical value, and may be used to describe power, voltage, or current as long as the terms are consistent. For example, if a certain radar has a pulse width of 1/2 microsecond and a pulse recurrence frequency of 2000 pulses per second, the duty cycle is $1/2\ (10^{-6}) \times 2000$, or 0.001 (0.1%). If the pulse power is 200 kw, the average pulse power is 200 kw x 0.001, or 200 watts. If the pulse current is 10 amp, then the average pulse current is 0.01 amp.

11-539. THE DECIBEL AND ITS USE. The decibel is part of a larger unit called the bel. As originally used, the bel represented a power ratio of 10 to 1 between the strength of two sounds. To gain a better understanding of the bel, consider three sounds of unequal power intensity. If the power intensity of the second sound is 10 times the power intensity of the first, its power level is said to be 1 bel above that of the first. If the third sound has a power intensity 10 times that of the second, its level is 1 bel above that of the second. But, since the third sound is 100 times as intense as the first, its level is 2 bels above that of the first. Thus a power ratio of 100 to 1 is represented by 2 bels; a power ratio of 1000 to 1, by 3 bels; a power ratio of 10, 000 to 1, by 4 bels, etc. It is readily seen, therefore, that the concept of bels represents a logarithmic relationship, since the logarithm of 100 to the base 10 equals 2 (corresponding to 2 bels), the logarithm of 1000 equals 3 (corresponding to 3 bels), etc. The exact relationship is given by the formula:

$$\text{Bels} = \log \frac{P2}{P1}$$

where $\frac{P2}{P1}$ represents the power ratio.

11-540. This logarithmic characteristic of the bel makes it a very convenient means for expressing power ratios. For example, assume that we desire to find the attenuation ratio of an rf attenuator which is to be used to measure transmitter power output. On test, it is found that 60, 000 watts of rf input to the attenuator produces an output of 6 milliwatts. To find the attenuation ratio use the equation:

$$\text{Attenuation ratio} = \frac{P2}{P1}$$

$$= \frac{60,000}{0.006}$$

$$= 10,000,000$$

This ratio can be expressed much more conveniently in terms of bels.

$$\text{Bels} = \log \frac{P2}{P1} = \log \frac{60,000}{0.006}$$

$$= \log 10,000,000 = 7 \text{ bels}$$

In this case, the attenuation ratio is 7 bels. In other words, P2 is said to be 7 bels up with respect to P1.

11-541. In all instances where P2 is numerically greater than P1, as in the above example, the final result is expressed as a positive quantity. When P2 is smaller than P1, the numerical result is the same, but it is expressed as a negative quantity. If, for example, P2 is 0.006 watt and P1 is 60,000 watts, then:

$$\text{Bels} = \log \frac{P2}{P1} = \log \frac{0.006}{60,000}$$

$$= \log 0.0000001 = -7 \text{ bels}$$

In this case, P2 is said to be 7 bels down with respect to P1.

11-542. Since the bel is a rather large unit, its use may prove inconvenient. Usually, therefore, a smaller unit, the decibel, is used. Ten decibels equal 1 bel. A 10-to-1 power ratio, which can be represented by a bel, can also be represented by 10 decibels (10 db), a 100-to-1 ratio (2 bels) can be represented by 20 db, a 1000-to-1 ratio (3 bels) by 30 db, etc. The previous formula for bels may be written to give a result in decibels merely by multiplying by 10. Thus the formula becomes:

$$\text{Decibels (db)} = 10 \log \frac{P2}{P1}$$

It should be clearly understood that the term decibel does not in itself indicate power, but is rather a ratio or comparison between two power values. In radar testing, however, it is often desirable to express performance measurements in decibels. This can be done by using a fixed power level as a reference. The original standard reference level was 6 milliwatts (0.006 watt), but to simplify calculations, a 1-milliwatt standard has been adopted and will be used hereafter in the part of this manual dealing with radar testing. (Note: A few equipments use one watt as a standard.)

11-543. REFERENCE LEVEL (DBM). When 1 mw is used as a reference level, the ratio is expressed in dbm's. The abbreviation dbm indicates decibels relative to a 1-milliwatt standard. Thus a pulsed radar transmitter having an average power output of 100 watts is said to have an average power output of 50 dbm. The conversion from power to dbm is made as follows:

$$\text{Average power (dbm)} = 10 \log \frac{P2}{P1}$$

$$= 10 \log \frac{100}{.001}$$

$$= 10 \log 100,000$$

$$= 50 \text{ dbm}$$

11-544. Conversions from power to dbm can be made more readily by means of the graphs shown in figures 11-54 and 11-55. If, as in the above example, the average power output is 100 watts, reference to the graph in figure 11-54 shows that the line representing 100 watts intersects the curve at point A, indicating that 100 watts is equivalent to 50 dbm.

11-545. Voltage and current ratios may also be expressed in terms of decibels, provided the resistance remains constant. For equal resistances, the formulas are:

$$\text{db} = 20 \log \frac{E2}{E1}$$

$$\text{db} = 20 \log \frac{I2}{I1}$$

The difference in the multiplying factor in these formulas (20 rather than 10, as in the case of power ratios) arises from the fact

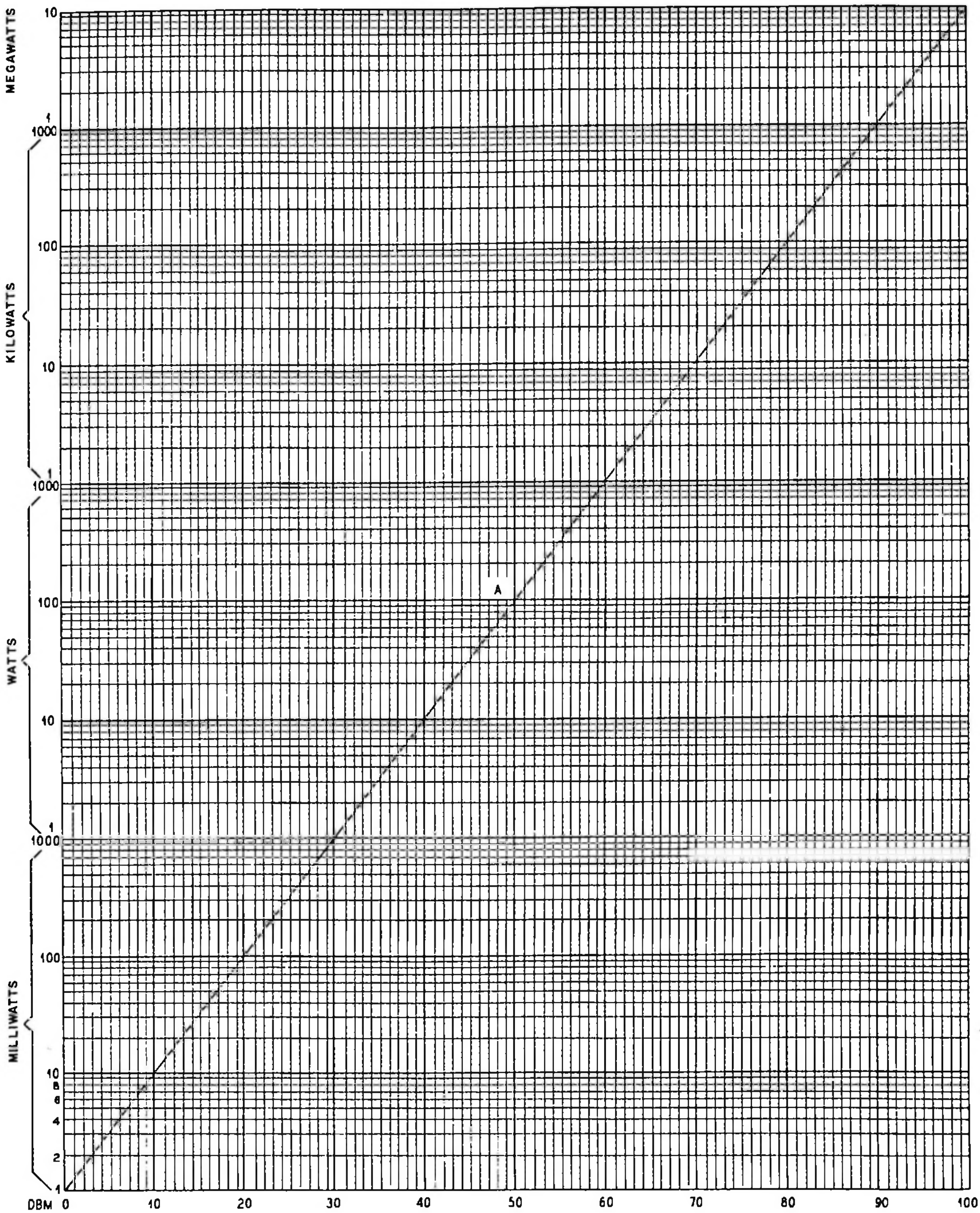


Figure 11-54. Power to DBM Conversion Chart, 1 Milliwatt to 10 Megawatts

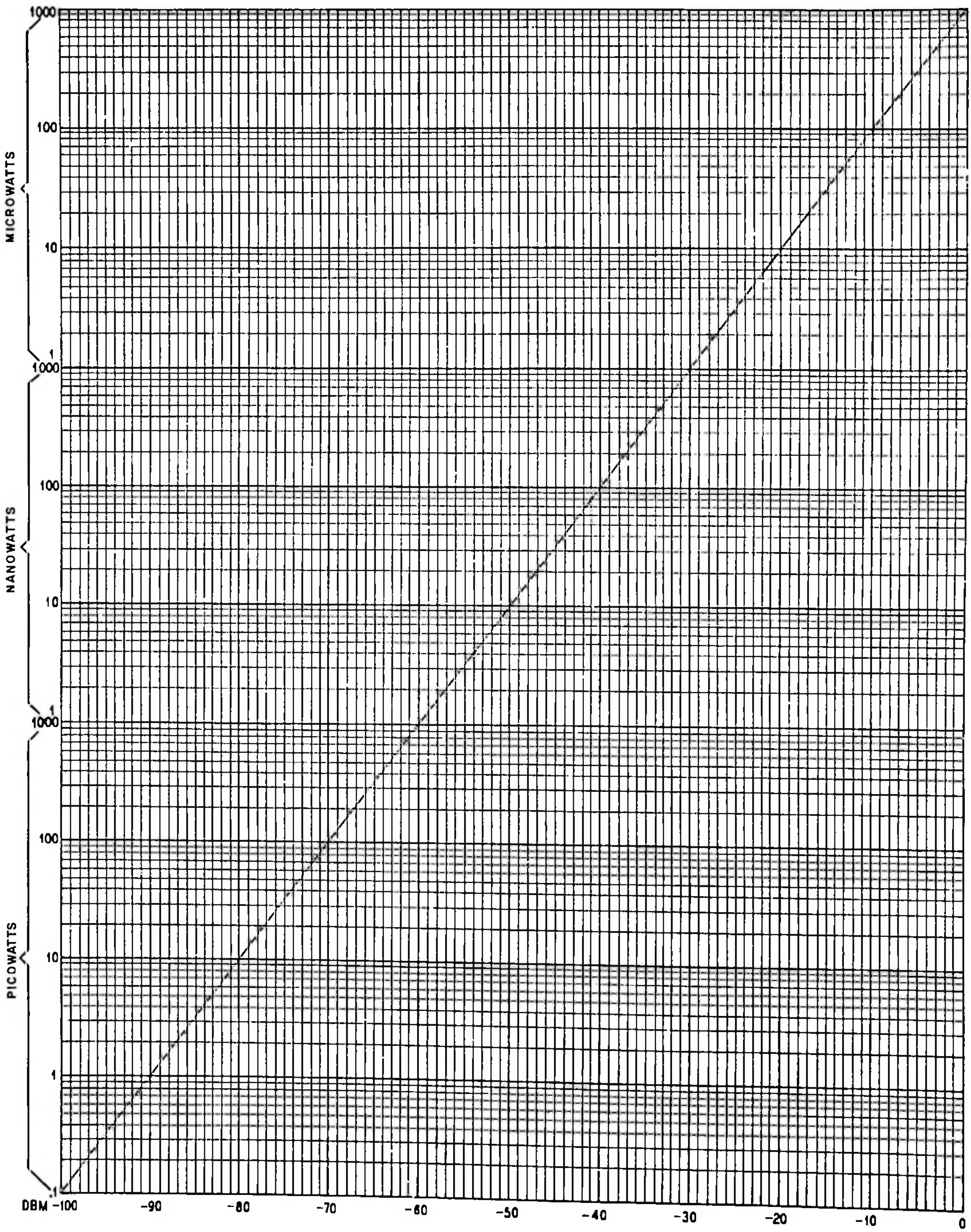


Figure 11-55. Power to DBM Conversion Chart, 1 Milliwatt to 0.1 Picowatt

that power is proportional to voltage or current squared, and when a number is squared, the logarithm of that number is doubled. For power ratios, the db value is 10 times the logarithm of the ratio. For voltage or current ratios, the db value is 20 times the logarithm of the ratio.

11-546. As was stated previously, power measurements are classified as either pulse power or average pulse power. When testing the over-all performance of a radar system, it is necessary to know the pulse power output of the transmitter. However, modern test equipments can measure only the average power output which, depending on the specific test equipment, may be given either in watts or in dbm. Therefore, the relationship between pulse power and average pulse power for a rectangular pulse must be determined. It is most convenient to express this relationship in terms of db.

11-547. The pulse power output of a transmitter in decibels relative to 1 mw can be found from the relationship:

Pulse power (dbm) = Average pulse power (dbm) + duty cycle (db)

For a given radar transmitter, the two latter quantities are easily determined. The average pulse power output is measured by means of test equipment. If the value obtained from the test equipment is expressed in watts, it must be converted to dbm. The conversion can be made by means of the graph shown in figure 11-56. Some test equipments are calibrated directly in dbm, and no conversion is required. The duty-cycle figure, which depends on the duration of the transmitted pulse and on the pulse repetition rate, can be found directly from the chart shown in figure 11-56.

11-548. POWER SAMPLING TECHNIQUES.

11-549. GENERAL. The testing of radar power always requires some method of removing or inserting the power to be measured. There are three principal devices used to accomplish this. They are the test antenna, the rf probe, and the directional coupler.

11-550. TEST ANTENNA. The test or pickup antenna consists of a directional antenna array which is broadly tuned to the radar band to be used. This antenna is placed in the radiation field of the radar antenna, and picks up a certain percentage of the radiated signal. The test antenna may be made portable by mounting it on a tripod frame, or it may be fixed by means of a bracket installed as a part of the radar system. It is common practice to locate the pickup antenna at least one diameter of the radar reflector away from the radar antenna as shown in figure 11-57A, and to orient the two antennas for maximum pickup. With this procedure the space attenuation is approximately 30 db. The exact loss either will be given for the particular installation or must be measured. The test procedure for this measurement is discussed under paragraph heading 11-603. Any subsequent testing should be done with exactly the same antenna spacing. Another placement method is to clamp the pickup antenna to the edge of the radar reflector in such a manner that the pickup is directed toward the radar antenna feed array. This method is shown in figure 11-57B. With the pickup in this position, antenna leakage power is used rather than direct radiation. This procedure has the advantage of allowing operation of the test equipment at various radar antenna positions and the radar antenna does not require careful orientation. The use of a pickup antenna has the important advantage of testing the entire radar facility including the radome, if the antenna is placed outside the radome. This enables the testing to show operating efficiency, with

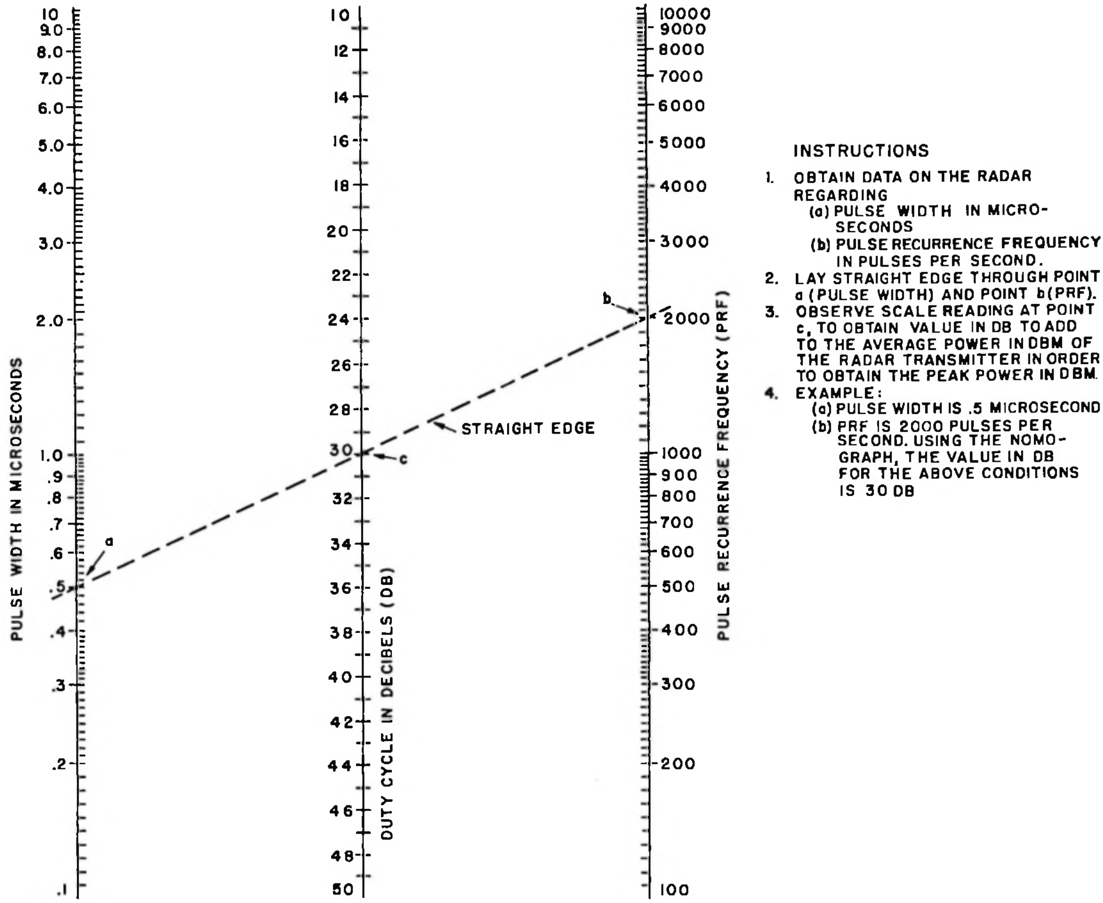


Figure 11-56. Average to Peak (Duty Cycle) Power Conversion Chart

all controllable factors included. Four primary disadvantages are associated with the pickup antenna method of sampling power. First, the placement of the antenna is critical; second, antennas are sensitive to frequency changes; third, it is difficult to make tests during radar scanning; fourth, nearby objects can modify the signal picked up by the antenna. Nearby objects, or propagation from other sources, can cause reflections and result in large errors in signal pickup. The presence of these reflections can be detected in the following manner.

While observing the signal picked up by the antenna, carefully move the pickup antenna closer to the radar antenna. A smooth increase in signal strength should be noted and, if the pickup antenna is moved farther away, a smooth decrease in signal strength should be noted. Any sudden or erratic variations or minimum points indicate that nearby objects are influencing the pickup, and another pickup position must be chosen.

11-551. RF PROBE. The rf probe consists of a small capacitive probe inserted into the

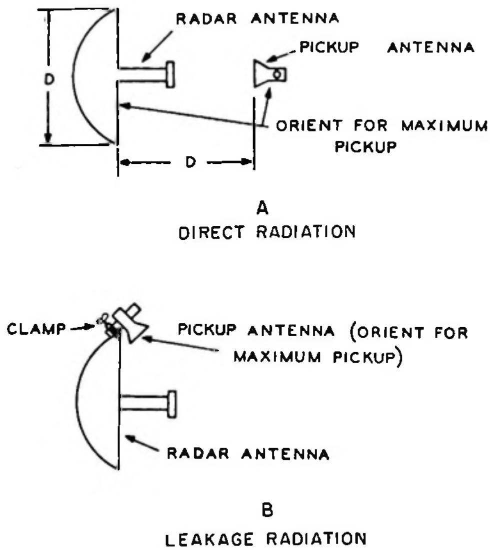


Figure 11-57. Placement of Pickup Antenna

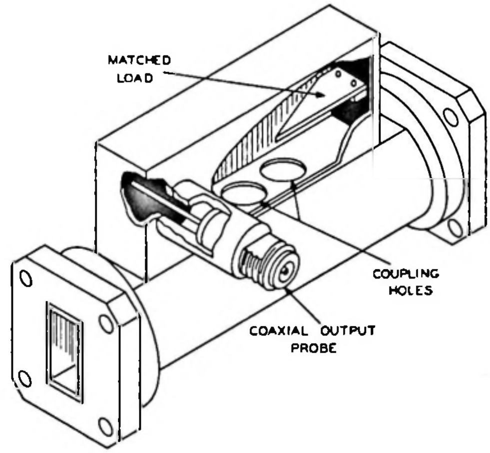


Figure 11-58. Directional Coupler, Cutaway View

electrostatic field in the rf transmission line. The greater the penetration of the probe, the greater the power pickup. The penetration of most rf probes is sufficient to provide 20 db or more attenuation between the main line and the probe output. The probe is fitted with a coaxial connector to facilitate connection to test equipment. In older radar facilities the rf probe was used extensively, but it is now considered obsolete since the development of the directional coupler. The rf probe allows normal radar operation during test, but has some disadvantages. First, reflections from nearby objects and in the rf line have a great effect on the attenuation figure; second, probe penetration is very critical; third, the probe is quite sensitive to frequency; fourth, the attenuation figure depends upon the load connected to the probe.

11-552. DIRECTIONAL COUPLER. The directional coupler, as the name implies, couples, or samples, energy only from a wave traveling in one particular direction in the waveguide. By the proper use of one or more directional couplers, reflected signal power can be prevented from affecting the accuracy of power measurements. Figure 11-58 shows a common type of directional coupler, which consists of a short section of waveguide coupled to the main-line waveguide by means of two small holes, and containing a matched load in one end and a coaxial transition in the other end. The degree of coupling between the main-line waveguide and the auxiliary is determined by the size of the two holes.

11-553. The action of this waveguide is explained by the diagrams in figures 11-59 and 11-60. In figure 11-59, power is shown flowing from left to right, and two small samples are coupled out at points C and D. Since the two paths, represented by C-D-F and C-E-F, to the coaxial probe are the same length, the two samples arrive at point F in phase and are picked up by the coaxial probe. With regard to the paths to the matched load, however, path C-D-F-E is one-half wavelength longer than path C-E, because the two holes are one-quarter wavelength apart.

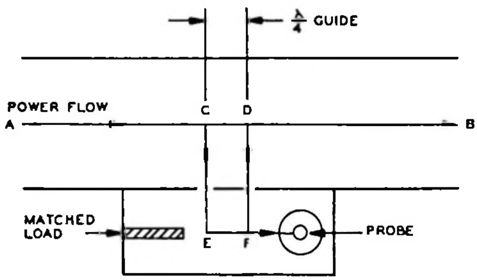


Figure 11-59. Directional Coupler, Direct Power Flow

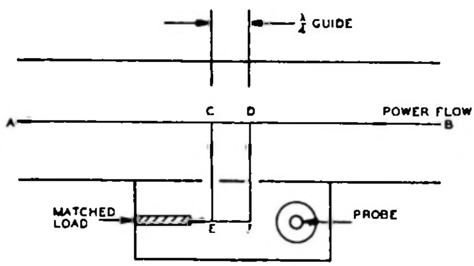


Figure 11-60. Directional Coupler, Reversed Power Flow

Therefore, the two samples arrive at point E 180 degrees out of phase, producing cancellation, and the load receives no power. Figure 11-60 shows the same coupler with power flowing in the reverse direction. Again samples are removed at points C and D. The two paths D-F-E and D-C-E are the same length, and the two samples arrive at point E in phase, and are absorbed by the load. However, path D-C-E-F is a half wavelength longer than path D-F and the resulting 180-degree phase shift causes cancellation at point E. The result is that the coaxial probe receives power only from a wave traveling from left to right in the main line, and any reflections causing power to flow from right to left have no effect upon the coupled signal. In practice, the attenuation between the coaxial output and the main line for power flowing from left to right is usually adjusted to be over 20 db and is called the nominal attenuation, or simply the attenuation, or the coupling factor. The ability to reject power in the reverse direction is called the directivity attenuation, or simply the directivity, and is usually greater than 20 db. If a certain coupler has a nominal attenuation of 20 db and a directivity of 20 db, the forward attenuation is 20 db and the reverse attenuation is 40 db. If the main line carries a 50-kw pulse the forward output is 500 watts pulse power and the reverse output is 5 watts pulse power. Five watts compared with 500 watts is too small to have any great effect.

11-554. BROAD-BAND COUPLER. Forward, or nominal, attenuation does not vary rapidly with frequency, but the directivity does. The rate of variation can be reduced by the use of other designs, so that the directional coupler can be operated over a broad band of frequencies. One type of broad-band coupler is the 3-hole coupler. If two directional couplers one-quarter wavelength apart are used, a broader bandwidth is obtained. Since the holes are one-quarter wavelength apart, two couplers have one hole common to both. This means that the three-hole coupler uses the action of two directional couplers, and the center hole serves as a common coupling to the two end holes. Another type of broad-band directional coupler is shown in figure 11-61. In this unit, the coupling holes are one-quarter wavelength apart and elongated. In addition, the two holes are in opposite halves of the main waveguide, which has the effect of causing a 180-degree phase shift between the coupled signals. This phase shift reverses the direction of coupling, so that when power enters a point A, the two signals arrive in phase at the essential output, and when power enters at point B, the two signals arrive in phase at the load and are absorbed. The result is that the coupler in figure 11-61 operates in reverse manner to that shown in figure 11-58. In

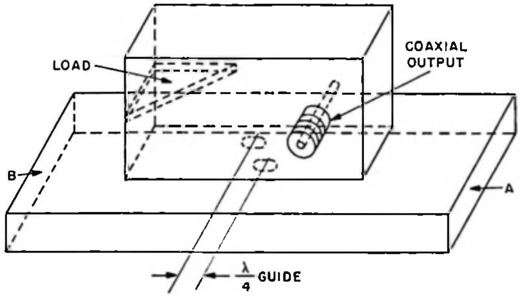


Figure 11-61. Reverse Directional Coupler

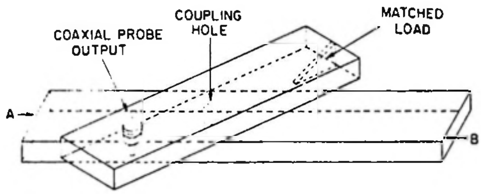


Figure 11-62. Single-Hole Directional Coupler

this case the directivity is relatively independent of frequency, but the coupling factor varies rapidly with frequency.

11-555. SINGLE-HOLE COUPLER. A third type of directional coupler, shown in figure 11-62, uses a single hole as the coupling element. This is called the Bethe-hole coupler. Through a single hole, waves are excited in the auxiliary guide because of the electric field and the magnetic field in the main guide. Because of the phase relations involved in the coupling process, the waves generated by the two types of coupling cancel in the forward direction, but reinforce in the reverse direction. Therefore, in figure 11-62, power entering at point A is coupled to the coaxial output, while power entering at point B is absorbed in the dummy load. If the two waveguides were parallel, the magnetic component would be coupled to a greater degree than the electrostatic and the directivity would be poor. By placing the auxiliary waveguide at the proper angle, the amplitude of the magnetically excited wave is made equal to that of the electrostatically excited wave (without changing the latter), and good directivity is obtained. The angle required depends upon the frequency of operation.

11-556. Directional couplers serve as stable, accurate, and relatively broad band coupling devices, which can be inserted into a transmission line so as to sample either incident or reflected power. In most cases, however, a directional coupler is made a part of the radar facility and is connected so as to sample the transmitted rf signal. Thus any undesired reflection from nearby objects, a mismatch between the line and antenna, or line discontinuities are virtually eliminated as a source of error in power measurements. Directional couplers are also made for use with coaxial transmission lines, and operate in a manner very similar to the two-hole coupler.

11-557. BIDIRECTIONAL COUPLER. In many cases it is desirable to measure the power reflected from the antenna as well as the direct power from the transmitter. A bidirectional coupler provides a convenient method to measure direct and reflected power. As shown in figure 11-63, it consists of a straight section of waveguide, with an enclosed section attached to each side, along its narrow dimension. Each enclosed section contains an rf pickup probe at one end and an impedance termination at the other end. The impedance termination in this case is in the form of a resistance card. The sections are supplied with energy from within the main waveguide through three openings spaced one-quarter wavelength apart. The rf probe farthest away from the transmitter is used to measure direct power, and the one nearest the transmitter is used to measure reflected power.

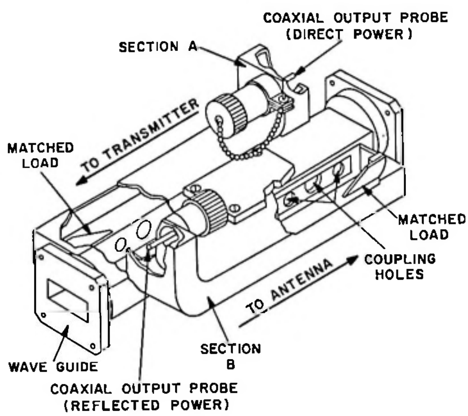


Figure 11-63. Bidirectional Coupler

11-558. Energy from the transmitter going toward the antenna enters the enclosed sections through the three openings on each side. Because the openings in each section are spaced one-quarter wavelength apart, the energy travels a quarter wavelength between each of the three openings. The energy coupled into the enclosed sections is attenuated a predetermined amount below that in the waveguide, by the coupling medium. As shown in figure 11-63, the center opening in each of the enclosed sections is larger than each of the holes to either side, thus allowing twice as much energy to enter through that opening.

11-559. The energy entering the enclosed section farthest away from the transmitter (section (A) of figure 11-63) is considered first. Part of the transmitted energy enters this enclosed section, and the rest of the energy goes to the antenna. This energy, because of the location and dimensions of the openings in Section A, enters the enclosed section and combines in phase, and is measured by the direct power probe. A power meter connected to the probe gives a direct indication of the transmitted power in the waveguide. The transmitted energy which entered the other enclosed section (section (B) of figure 11-63) is zero, because of the phase displacement of the three openings at that enclosed section. The energy passing the first opening is 180 degrees out of phase with that of the center opening, and is in phase with the energy at the third opening. Since the center opening is sufficiently large to supply twice the magnitude of energy as that supplied by either of the two openings, the energy is cancelled. The end of the enclosed section is terminated as described in the first paragraph.

11-560. To measure reflected energy caused by either standing waves or energy received from targets, section (B) acts in exactly the same manner as section (A) when making direct power measurements. Since the direction of energy flow is reversed, the energy now appears at the reflected power probe in section (B). This reflected energy, upon entering section (A), is cancelled out in the same manner that section B cancelled the transmitted power when performing direct power measurements. Typical nominal attenuation of a bidirectional coupler is 40 db for both the direct and reflected outputs. This attenuation must be considered when making power measurements with a bidirectional coupler. The use of the bidirectional coupler for making standing-wave measurements is discussed in paragraph 11-713.

11-561. ATTENUATORS. Attenuators in present use are classified as dissipative or non-dissipative. The cut-off waveguide section is a good example of a non-dissipative attenuator, in that the attenuator merely rejects signals instead of converting them into heat. Dissipative coaxial attenuators are usually short coaxial sections which use resistive material for a center conductor. One such attenuator uses a glass rod, upon which a thin deposit of metal has been sprayed, for a center conductor. Aquadag is sometimes used in place of the metal film.

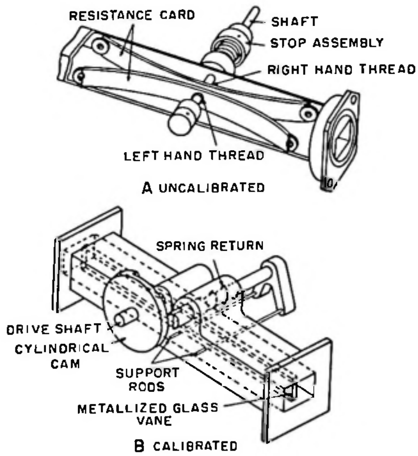


Figure 11-64. Waveguide Attenuators, Showing Construction

11-562. Short, fixed attenuator sections, called pads, come in a large variety of forms and loss values. For example, the CN-42/UP is a 10-db 50-ohm coaxial attenuator and the CN-43/UP is a 16-db, 50-ohm coaxial attenuator. The coaxial attenuator can be made variable by constructing the resistive section in two telescoping sections, so that the length, and therefore the attenuation, can be varied.

11-563. Dissipative waveguide attenuators consist of strips of resistive material placed inside the waveguide, parallel to the electrostatic field. Where the exact value of attenuation need not be known, the strips are made of bakelite or fiber, with an aquadag coating on one side, as shown in part (A) of figure 11-64. Calibrated attenuators are usually of the metalized glass variety, an example of which is shown in part (B) of figure 11-64. As the movable resistive element approaches the center of the waveguide, the power loss is greater. The resistive element is driven by a dial-and-cam arrangement that is calibrated in db; the cam surface can be shaped to give any desired spacing of the calibration marks. The ends of the resistive element are tapered to produce as little reflection as possible over a wide band of frequencies.

11-564. Most waveguide attenuators do not have sufficient range to cover all values required in normal use. To overcome this limitation, it is common practice to use two attenuators in cascade, so that the total attenuation is the sum of the individual readings. In some cases, both attenuators are continuously variable. In other cases only one attenuator is continuously variable and the other is adjusted in steps. One modern attenuator unit has one attenuator with an attenuation range of 7 to 45 db and the other with a fixed attenuation of either 0 or 35 db. Thus the combination provides two ranges of attenuation, from 7 to 45 db and 42 to 80 db. The over-all range is said to be 7 to 80 db in two overlapping steps.

11-565. Since dissipative attenuators are easily damaged by application of too much rf power, you must be careful to keep the applied power below the maximum rating of the equipment in use. Power overloads cause the resistive element to blister and peel away from the supporting section, and this condition in turn causes the attenuation value to change, and produces excessive power reflection. Once an attenuator is damaged, it should be discarded. Damage can be detected by inspecting the surface of the attenuator material; it should be smooth and of even color. The use of a directional coupler practically eliminates the possibility of power burnout, since a nominal loss of 20 db normally reduces the radar output to a level below the maximum safe value. The following equation will provide you with the power into the attenuator:

$$P_i = P_o \text{ (dbm)} - \text{duty factor (db)} - \text{directional coupler loss (db)}$$

where:

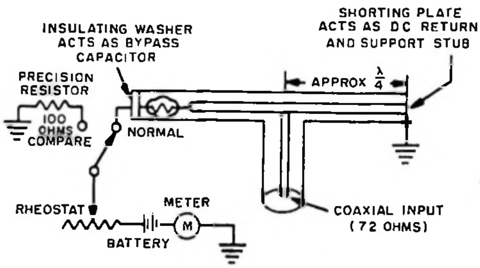


Figure 11-65. Simple Thermistor Power Meter

P_i = maximum possible power into attenuator

P_o = transmitter power output (figure 11-54)

Duty factor can be determined from figure 11-56.

11-566. SIMPLE THERMISTOR POWER METER METHOD.

11-567. GENERAL. An early type of power meter using a thermistor is shown in figure 11-65. In this meter, the thermistor is mounted at the end of a coaxial line; the center conductor of the line is supported by means of a quarter-wave stub, which also serves as a dc path for the thermistor current. This stub restricts the frequency range of the assembly, and is broad-banded and adjusted at the factory for the desired frequency range. The other end of the thermistor is brought out through an insulating washer that acts as a short circuit to rf but allows dc to flow through the meter circuit. The assembly is so constructed so that when the thermistor resistance is 100 ohms, an impedance match is achieved. A precision 100-ohm resistor is included to provide a comparison resistance so that the thermistor circuit can be set for the proper 100-ohm resistance.

11-568. TEST PROCEDURE. The following steps describe the technique you follow in making this test:

a. Adjust the rheostat until the meter indicates the same current for the thermistor as for the 100-ohm resistor. To check on the adjustment, operate the switch to the compare position and observe the current; when you return the switch to the normal position, the meter reading should not change.

NOTE

This adjustment allows for different operating temperatures. When the ambient temperature is high, less dc power is required to bring the resistance down to 100 ohms.

b. Note the current reading (the reading will vary with ambient temperature), and refer to the chart furnished with the meter to determine the current reduction factor for the particular current reading you obtained.

c. Adjust the rheostat to give the reduced current value indicated on the chart. (The reduction in current causes the thermistor temperature to decrease by an amount that requires 6 mw of rf power to return the temperature to the previous value.)

d. The rf power to be measured is supplied to the coaxial input. When 6 milliwatts of rf power is present, the meter again indicates the current value you noted in step b.

Thus, the instrument serves to indicate a standard reference power level.

11-569. CRYSTAL DETECTOR-SYNCHROSCOPE METHOD.

11-570. GENERAL. Figure 11-66 shows the setup and test equipment required to use this method. The test equipment consists of a pickup antenna for obtaining a sample of the transmitted rf power, a calibrated variable attenuator, a crystal detector, and a synchroscope.

11-571. PICKUP ANTENNA. The pickup antenna is provided to sample a known amount of the transmitter power. In one early equipment, the antenna consisted of a dipole and parabolic reflector enclosed in a Plexiglas housing. In use, the antenna is placed at a certain distance from the radar antenna, the two antennas are oriented to point directly at each other, and the polarization of the pickup antenna is adjusted to agree with that of the radar antenna. The pickup antenna, which is located about 10 feet above ground, has directional properties to minimize the effect of reflection from ground and nearby objects.

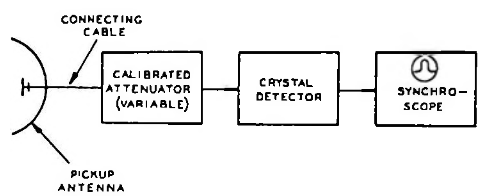


Figure 11-66. Crystal Detector Synchroscope Power Measuring Circuit

11-572. ATTENUATOR. In older microwave equipments, the most commonly used attenuator is the cut-off waveguide type. In this type, a circular pipe, too small to act as a waveguide, has an adjustable length as shown in figure 11-67. The longer the cut-off section, the greater the attenuation. In use, the sliding section is operated by means of a rack-and-pinion gear assembly, and a calibrated dial which drives the pinion gear gives the attenuation at each setting. The resistive disks in the attenuator serve to provide an impedance match over a wide range of frequencies.

11-573. SYNCHROSCOPE. The synchroscope is an adaptation of the oscilloscope. A trace is produced only when it is initiated by an input trigger, as contrasted with the continuous sawtooth sweep provided by the oscilloscope. Synchroscope circuits are similar to oscilloscope circuits except for the signal and sweep channels. Those circuits are shown in block diagram form in figure 11-68. Refer to Chapter 8 Section IV for a general discussion of these circuits.

11-574. The signal channel includes an input circuit, which is usually in the form of a 72-ohm adjustable-step attenuator. Various degrees of attenuation are available, and a dial is calibrated to indicate how much at-

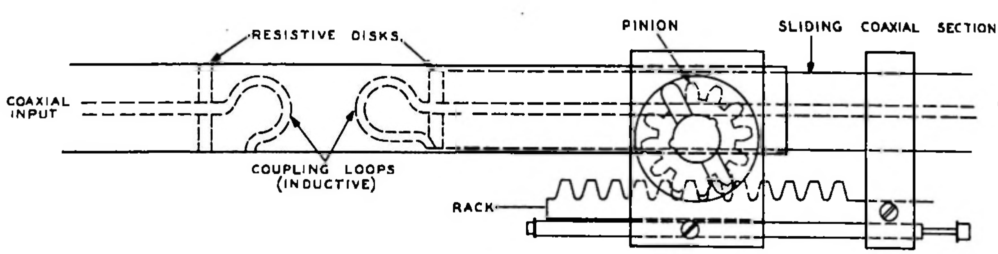


Figure 11-67. Cutoff-Waveguide-Type Attenuator

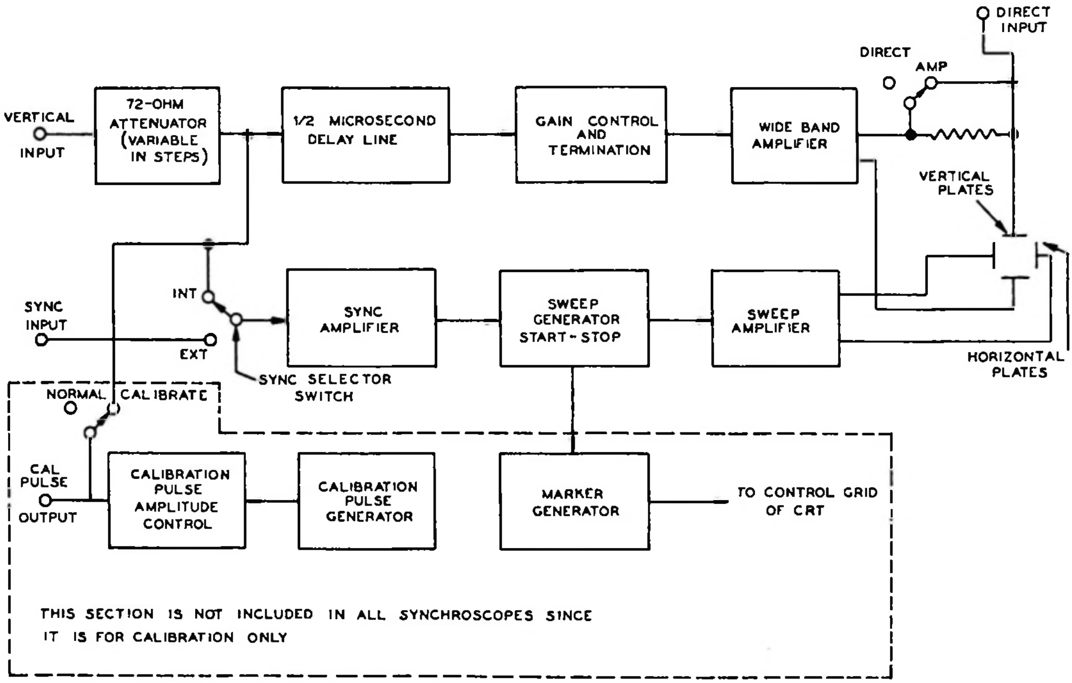


Figure 11-68. Block Diagram of a Typical Synchroscope

tenuation is present. This attenuator ensures that all signals, regardless of amplitude, produce about the same input level to the amplifier section. Following the attenuator is an artificial delay line, which is a low-pass filter with a cut-off frequency higher than the highest frequency to be passed, and which has an impedance of 72 ohms. The delay line is terminated with a 72-ohm gain control. One purpose of this delay line is to delay the signal to be observed until the sweep trace has been initiated by a portion of the input signal which is not delayed. If the delay line were not used, the initial portion of the waveform would not appear on the trace because a certain amount of time is required to start the sweep generator. With the delay line in use, the signal does not reach the amplifier until 0.5 microsecond after the trace starts; as a result, the entire pulse is seen.

11-575. The gain control feeds a wide-band or video amplifier, which is connected to the vertical deflection plates. In addition, an external connection is provided to the vertical plates. The horizontal circuit consists of a sync switch for either internal or external sync, a sync amplifier with a gain control, and a start-stop sweep generator, which does not develop a sweep voltage until a pulse of sufficient amplitude is fed in. The duration of the sweep, or sweep speed, is adjustable from a very few microseconds to about 250 microseconds. The sweep generator is followed by a conventional horizontal amplifier. Since the trace is triggered by the input signal, the synchroscope may be used to observe non-periodic pulses, such as those occurring in a radar system that has an unstable prf.

11-576. In later designs, you will commonly find provisions for calibration of input voltages and sweep time. Voltage calibration is accomplished by comparing the unknown voltage with a variable-voltage pulse of known value, generated internally. The calibrating pulse is adjusted to equal in amplitude the unknown voltage, and the value is read from the dial that controls the calibrating pulse. Sweep-time calibration is performed with the aid of marker pulses produced by accurately adjusted tuned circuits. The marker pulses appear on the trace as a series of bright dots spaced at intervals chosen by the operator. In a typical synchroscope, marker intervals of 0.2, 1, 10, 100, and 500 microseconds are selected in accordance with the time duration of the pulse under test; for greater accuracy interpolation is used.

11-577. TEST EQUIPMENT CALIBRATION PROCEDURE. In practice, the crystal detector and synchroscope must be calibrated. The usual method is to feed a 6-milliwatt pulse of rf to the crystal, and adjust the gain of the scope to give a one-half inch pulse. The controls must not be varied after calibration. One calibration procedure is as follows:

a. Tune a signal generator to the approximate transmitter frequency.

b. Feed generator output to a standard power meter.

c. Adjust signal generator for cw output, and set output level for 6 milliwatts as indicated on power meter.

d. Remove power meter, and feed generator output to crystal detector and synchroscope. Set generator for pulsed output. (The peak pulsed-generator power must be the same as the average cw power.)

e. Adjust synchroscope gain control to give one-half inch deflection on synchroscope.

11-578. ADDITIONAL CALIBRATION PROCEDURES. To simplify the calibration procedure, some early test units incorporated a meter and rheostat in series with the crystal detector. The resistance of the meter circuit is high enough to prevent loading and the rheostat is set so that the meter indicates a given current when 6 milliwatts of cw power is fed to the crystal. This meter must be checked against the standard power meter at frequent intervals to ensure calibration. When this system is used, the synchroscope calibration procedure is as follows:

a. Tune signal generator to approximate transmitter frequency.

b. Feed generator output to crystal detector.

c. Set generator for cw output, and adjust output for standard meter reading.

d. Adjust generator for pulsed output, and set synchroscope gain for one-half inch deflection.

NOTE

The calibration meter will read only during cw calibration. When pulsed output is used, the average meter current is too low to produce a readable deflection.

11-579. TEST PROCEDURE. The synchroscope and crystal are now calibrated for a 6-milliwatt level and are ready to test transmitter power. Transmitter power is measured as follows:

a. Place pickup antenna some accurately known distance from radar antenna, and orient the two antennas for maximum pickup.

b. Connect pickup antenna to calibrated attenuator by means of special cable supplied with antenna, and connect attenuator output to crystal detector and synchroscope.

c. Adjust attenuator for one-half inch pulse as indicated on scope.

d. Find total attenuation between crystal detector and radar antenna.

e. Peak power (in watts) is 0.006 multiplied by the attenuation ratio.

11-580. ATTENUATION SOURCES. In the method of testing rf power discussed above, the total attenuation arises from four sources. These sources are: first, loss between radar antenna and output of pickup antenna, called space loss; second, antenna cable loss; third, attenuator zero loss, that is, when the attenuator is set to zero, there is still a certain amount of loss inherent in the attenuator and cables; and fourth, attenuator loss.

11-581. The data on space loss is supplied in test equipment operating instructions for certain antenna placements on given radars and can be reproduced. This loss depends to a certain extent upon frequency, and correction must be made to give the loss at any specific frequency. Antenna cable loss and attenuator zero loss are measured at the factory and marked on the equipment, but the actual loss varies with age and atmospheric conditions and should be tested periodically, as well as after periods of disuse.

11-582. TEST LIMITATIONS. Measurement of power by the method discussed above offers the following advantages: First, peak power is measured directly. Second, the width of the rf pulse can be observed. Third, the shape of the rf pulse can be observed. Fourth, a check of the overall rf radiating capability of the radar facility is accomplished.

11-583. This method has the following disadvantages: First, the involved calibration procedure reduces accuracy. Second, the whole procedure is somewhat complicated. Third, frequency changes necessitate recalibration of losses.

11-584. CW RADAR POWER MEASUREMENTS.

11-585. GENERAL. Continuous wave (cw) radar equipments are used primarily to provide useful data relative to moving targets. Cw radar employs one of many adaptations of the Doppler principle to present intelligible information. Cw radar can operate down to zero range and in the presence of severe clutter. The above advantages make cw radar particularly adaptable for speed measuring.

11-586. DOPPLER EFFECT. When a fixed frequency cw carrier strikes a moving target, a reflected wave having a slightly altered frequency is returned to the receiver. This frequency shift is because of Doppler effect. A particularly notable example of this frequency shift is the behavior of sound waves. A practical analogy is that of the locomotive whistle which seems to shift suddenly to a lower frequency as the train passes. Since the motion of the train has made the sound waves pile up when approaching and stretch out when receding, the true frequency of the whistle lies midway between the two observed pitches.

11-587. A simple cw radar set employing Doppler effect for determination of aircraft speed is illustrated in figure 11-69. A fixed frequency carrier (f) is generated by a magnetron in the transmitter and conducted to the antenna. A sample of the fixed carrier

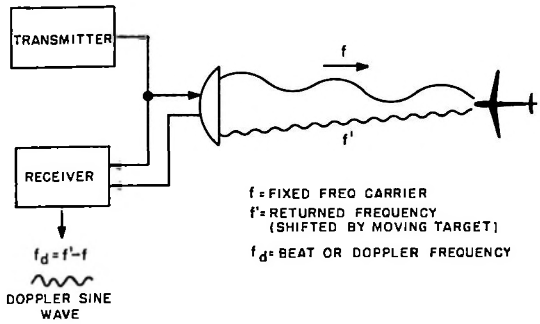


Figure 11-69. CW Radar Equipment

frequency is fed to a detector in the receiver. The antenna radiates this fixed carrier frequency to the moving target which reflects back to the antenna a wave having a slightly altered frequency (f'). The reflected wave is changed in frequency by the moving target. The reflected frequency (f') and the sample of the transmitter frequency (f) are mixed in the crystal detector in the receiver.

11-588. The mixing of frequencies f and f' results in a sum frequency and a difference (f_d). The detector cannot pass the sum frequency. Thus, the difference frequency f_d is the Doppler or beat frequency and depends upon speed as shown below:

$$f' = \frac{c + v}{c - v} f$$

in which c = velocity of light

v = velocity of the target

The beat frequency is then

$$f_d = f' - f = \frac{2v}{c - v} f$$

Since the target velocity is very small compared to the velocity of light c, this can be written:

$$f_d = \frac{2v}{c} f = \frac{2v}{\lambda}$$

For v in miles per hour and λ in centimeters, this becomes:

$$f_d = \frac{89v}{\lambda}$$

11-589. Most of the direct power meters described in Secion II of Chapter 8 are suitable for measuring average power and permit a direct determination of cw power. Power measurements of cw radar equipments are straightforward and the methods of power measurement given in Section II of this chapter are fundamentally suitable for this type equipment. You must keep in mind, however, that most cw radar equipments operate in the microwave region, and only those measurements relating to this band are applicable.

11-590. THERMISTOR BRIDGE METHOD.

11-591. GENERAL. Modern methods of measuring power make use of the arrangement shown in figure 11-70 which incorporates an attenuator and thermistor bridge. The coupling device is usually a directional coupler, but a pickup antenna can be used if desired. The thermistor bridge and attenuator method of power measurement provides an indirect indication and the simplicity of operation makes the method very desirable for radar maintenance work.

11-592. The complicated procedures found necessary in the early thermistor power meters were a result of the effect of ambient temperature changes which caused changes in thermistor resistance. This difficulty has been corrected by the development of the compensated thermistor bridge circuit, shown in figure 11-71. This circuit

Figure 11-70. Thermistor Bridge Method of Power Measurement

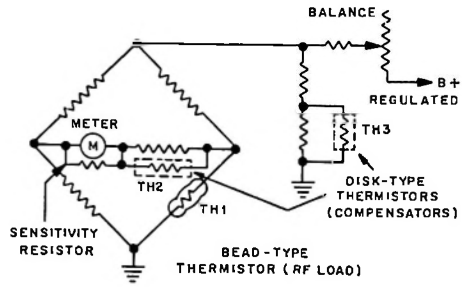


Figure 11-71. Compensated Thermistor Bridge Circuit

incorporates a Wheatstone bridge circuit, which is made up of three resistors and a bead-type thermistor. The bead thermistor acts as a matched load for the rf line when the bridge is balanced. Two disk-type thermistors are used for temperature compensation, and are in thermal contact with the section of the rf line containing the bead thermistor. Figure 11-72 shows the construction of two typical thermistor mounts, the coaxial and the waveguide types.

11-593. The resistance of the bead thermistor is varied by means of the balanced rheostat. The bridge is balanced by electrically adjusting the circuit to a point where one milliwatt of rf power produces zero meter current. A zero-centered meter is used, and the value of the sensitivity resistor is factory adjusted so that with full scale meter reading one milliwatt of rf power is required to restore the reading to midscale.

11-594. BRIDGE TEMPERATURE COMPENSATION. Temperature compensation of the bridge is necessary for two reasons: first, bridge balance must be maintained, and, second, sensitivity must remain constant under varying temperature conditions. In figure 11-71, thermistor TH3, with its associated resistors, compensates for any unbalance caused by temperature variation, as follows: As the ambient temperature rises, the resistance of the bead thermistor drops, so that if no compensation were provided, an unbalanced condition would occur. At the same time, however, the resistance of thermistor TH3 decreases, causing a reduction in the dc voltage applied to the bridge. The resulting reduction in dc bridge power allows the resistance of the bead thermistor to return to normal, with the result that the bridge balance is maintained. Since rf is applied only to the bead thermistor, compensation does not depend upon rf power.

11-595. At high ambient temperatures, the value of dc applied to the bridge is low. This condition results in reduced bridge sensitivity and, if uncompensated, results in errors in measurements. Compensation is provided by thermistor TH2 which is effectively in series with the indicating meter. At high temperatures, where bridge sensitivity is reduced, this thermistor presents a lower series resistance to the meter and, therefore, increases the meter sensitivity in the same proportion as the loss of bridge sensitivity. Thus, the over-all sensitivity is maintained essentially constant at different temperatures. In practice most bridge circuits are designed to give exact readings at temperatures of 0, 30, and 60 degrees centigrade. At other temperatures, a slight error exists but is too small to be considered in radar maintenance.

11-596. BRIDGE OPERATION. Use of the thermistor bridge greatly simplifies the measurement of rf power. With the equipment turned off, the meter is mechanically zero adjusted to center scale. A standard power meter scale is shown in figure 11-73A.

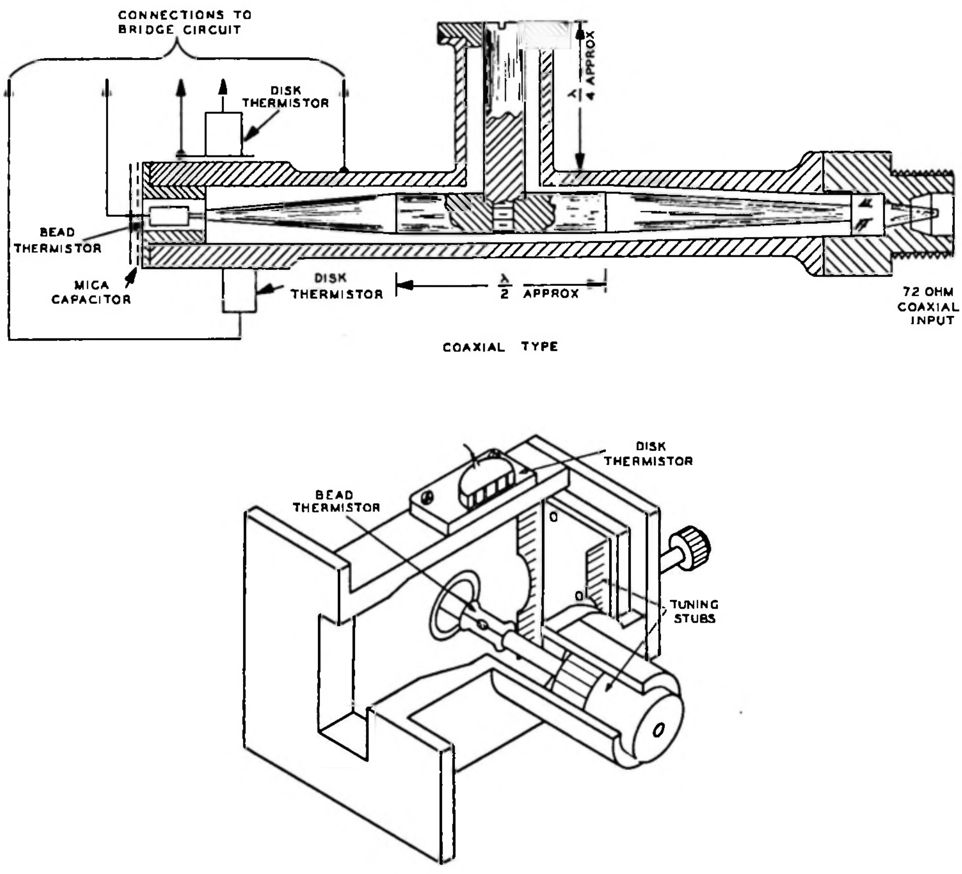


Figure 11-72. Typical Thermistor Mounts, Showing Construction

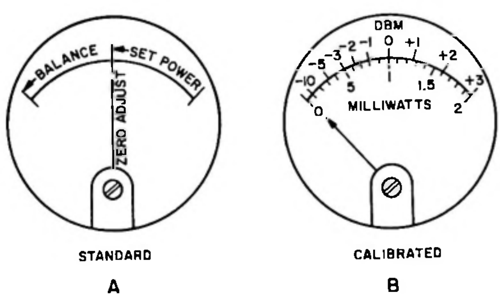


Figure 11-73. Typical Power-Meter Scales

Then the equipment is turned on, and the balance control is operated to produce deflection to the balance point on the meter. The meter is now ready to indicate power. A midscale reading indicates 1 mw, and a full-scale reading indicates 2 mw. The calibration is nearly linear, but at any point except midscale, the bead thermistor presents a slight rf mismatch, causing errors in the indication. In some equipments the meter scale is marked directly in power or dbm; for example, figure 11-73B shows meter calibration in milliwatts and dbm.

11-597. Thermistor bridges measure only the average power level because of an effect called thermal time lag. The resistance of a thermistor changes only as rapidly as the bead heats or cools. Since this heat transfer is a relatively slow process, the time required for an appreciable change in resistance is great compared to the pulse period of the average radar. In practice, the time-lag period includes many radar pulses, with the result that a true average-power indication is obtained. One obvious disadvantage of the thermistor bridge lies in the fact that erratic or sudden variations in power level often occur too rapidly to be detected with this kind of power test equipment.

11-598. TEST PROCEDURE. The test setup for this method is shown in figure 11-70. The signal from the coupling device is attenuated until the thermistor bridge shows one milliwatt of power. The input power is then calculated by finding the total attenuation between the radar and bridge circuit, using the formula:

Power (dbm) = Coupling loss (db) + cable losses (db) + attenuator reading (db)

If peak power indications are desired, the duty cycle, in db, must be added to the other losses. For example, if a given radar has a pulse width of 0.5 microsecond and a prf of 2000, corresponding to a duty cycle of 0.001 (30 db), and if the average power is 100 watts (50 dbm), the peak power is 50 dbm plus 30 db, or 80 dbm, which corresponds to 100 kw.

11-599. POWER TEST EQUIPMENT CALIBRATION.

11-600. GENERAL. In order to provide an accurate periodic check of power testing equipment, it is desirable to set up a calibration standard. Each individual test instrument is checked against the standard at periodic intervals (such as quarterly), and any calibration errors are noted and taken into account during testing. If a certain test set suddenly shows a large calibration error, it should be overhauled before being returned to service.

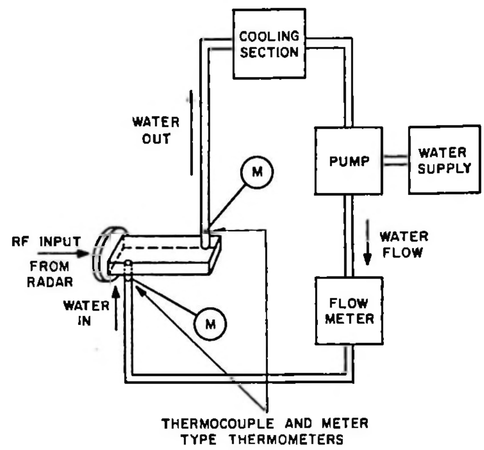


Figure 11-74. Water-Load Power-Test Setup Used for Calibration

11-601. WATER LOAD POWER METER. One calibration standard is known as the water-load power meter, which is illustrated in figure 11-74. The water load provides a nonreflecting termination for the rf line, and the rf output, which is completely absorbed by the water, produces a rise in the water temperature. The water is circulated through the load by means of a pump, and the temperature of the water going into the load is compared with the temperature of the water at the outlet. If the rate of water flow and the difference in temperature at the input and output are known, the absolute power in watts is easily calculated.

11-602. TEST PROCEDURE. The procedure for checking a thermistor bridge and attenuator is as follows:

a. Connect water-load test equipment to the rf output of a radar set, as shown by the setup in figure 11-74.

b. Connect thermistor bridge and attenuator to the same rf line by means of a directional coupler.

c. Start water flow with radar set off, and establish uniform rate of flow. (Care must be taken to eliminate air bubbles.) The two thermocouple meters should read the same temperature.

d. Turn the radar transmitter on. A rise in output water temperature should be noted. Allow operation to continue until there is no further change in the temperature difference between water input and output. This is called the equilibrium condition.

e. Note the temperature difference and the rate of water flow.

f. Measure the rf power with the thermistor bridge and attenuator test set.

g. Convert readings taken in step a to power in watts, using the following formula:

$$\text{Power (in watts)} = 4.18\ m\ C_p\ \Delta T$$

where:

m = water flow in grams per second

C_p = specific heat of water in calories per gram per degree centigrade

= 1

ΔT = temperature difference in degrees centigrade

h. Compare result obtained in step g with the power reading obtained in step f. The difference between the two values is the error in the thermistor bridge and attenuator test set.

11-603. ATTENUATION CHECKS.

11-604. GENERAL. The directional coupler is the only coupling unit for which the coupling loss is accurately predetermined. Each coupler is accompanied by a tag or stamped nameplate which gives the coupling loss and, in some cases, the midband directivity. Several variable factors are inherent in the pickup antenna method of coupling. The attenuation loss (or space attenuation) between two antennas is given by the formula:

$$\text{Attenuation(db)} = 10 \log \frac{158d^2}{g_1 g_2}$$

where:

d = spacing between antennas in centimeters

g_1 = gain of radar antenna in terms of power ratio

g_2 = gain of pickup antenna in terms of power ratio

The formula is accurate only when the antennas are spaced more than one reflector diameter apart. You can readily see from this formula that, for any given pickup antenna, the attenuation is determined by the gain of the radar antenna and the spacing between antennas. Furthermore, the gain of the pickup antenna is dependent upon the operating frequency. For convenience in future measurements, it is desirable to establish a standard spacing for the pickup antenna and measure the attenuation at several points in the operating-frequency band.

11-605. Connecting cables present another source of test error, because their attenuation value increases with age, and may vary depending upon atmospheric conditions and handling. Flexing of the cable can cause considerable variation in attenuation, and should be avoided, if possible. Most cables are factory calibrated, and are marked to indicate the attenuation in db. Since cable loss is subject to such wide variation, all connecting cables should be checked periodically at the operating frequency, under normal atmospheric conditions.

11-606. POWER TESTING TECHNIQUES. The performance testing of radar facilities is based upon a method that consists of establishing a standard reference power level and, by means of attenuation, comparing all signals to the reference level. For example, a transmitter power test is made by measuring the attenuation necessary to reduce the power to 1 milliwatt, and receiver sensitivity is checked by measuring the attenuation necessary to reduce 1 mw to the minimum-discernible-signal (mds) level.

11-607. Temperature-compensated thermistor bridges make it possible to provide a very accurate check of the reference power level, and broadbanding techniques can make reference measurements accurate over an entire radar band. For example, a modern thermistor mount is accurate over the frequency range of 8500 to 9600 mc. Broadband attenuators are now developed to a point where reliable attenuation indications make accurate power measurements possible. The metallic-film-on-glass attenuator provides an accurate and stable source of attenuation.

11-608. CALIBRATION STANDARDS. If reliable attenuation figures are to be attained, some type of calibration standard must be employed. The calibration standard generally used is the cutoff calibrated attenuator, which has the advantage of possessing linear dial markings; that is, the degree of dial rotation required to change the attenuation from 20 to 23 db is the same as for a change from 50 to 53 db. This linear characteristic makes the attenuator very useful for calibration purposes in cases where the exact attenuation is not important but the change in attenuation must be known.

11-609. Under the conditions where resistive coaxial or waveguide attenuators are used, you must necessarily rely upon calibration figures supplied in the maintenance literature accompanying the attenuators. Correct results are obtained as long as the unit is not subjected to mishandling or overload. Once a calibrated resistive attenuator is damaged, either mechanically or electrically, it should be discarded, because the calibration is no longer reliable, and the swr may be excessive. In the waveguide attenuator which employs a metalized glass vane as the resistive element, the glass vane is easily cracked or broken by mechanical shock, and it must be used with care.

11-610. CALIBRATION ACCURACY. The calibration procedures described below are usually performed when a particular device is suspected of being in error or is not marked. The accuracy of such measurements is dependent upon the accuracy of the calibrated attenuators and directional couplers used. Errors can result from operational errors caused primarily by carelessness. The precautions listed below should always be observed.

a. Allow all associated equipment to reach operating temperature. A 1-hour warm-up is preferable; however, a half hour can be considered a minimum time.

b. Subsequent testing should be done under as nearly identical conditions as possible, so as to minimize corrections otherwise made necessary by variations in temperature, pressure, relative humidity, and other operating conditions.

c. If conditions do change, apply corrections as specified in instructions or maintenance literature.

d. Make sure that all cable connections and fittings are tight.

e. Repeat those measurements considered critical several times and strike an average of the readings obtained.

11-611. PICKUP-ANTENNA CALIBRATION. The attenuation between the radar and pickup antennas is calibrated by any one of several methods. Three methods are given in the text to follow. The first and third methods rely upon the accuracy of a calibrated directional coupler, while the second, or alternate test procedure, does not.

11-612. The test setup for the first method is shown in figure 11-75. This method is accomplished by performing the steps listed below:

a. Install pickup antenna in a position at least one radar-dish diameter away from the radar antenna or farther. Make sure that this position can be duplicated in any subsequent tests.

b. Orient both antennas for maximum pickup.

c. If radar equipment has no built-in directional coupler, install one temporarily. The calibration of this coupler must be reliable.

d. Use the thermistor bridge and attenuator type of power meter to monitor the power.

e. With power meter connected to directional coupler, set calibrated attenuator for a 1-mw power reading. Record attenuator reading.

f. Transfer the connecting cable to pickup antenna, reset attenuator for 1 mw, and record new attenuator reading.

g. The space attenuation between the antennas is a combination of the directional-coupler attenuation and the difference in dial readings in steps e and f.

NOTE

1. If dial readings in steps e and f are the same, the space attenuation is equal to the directional-coupler attenuation.

2. If reading in step e is greater than that in step f, space attenuation = coupler attenuation + step e reading — step f reading.

3. If reading in step e is less than that in step f, space attenuation = coupler attenuation — step f reading + step e reading.

11-613. The second method, shown in figure 11-76, permits testing at different frequencies, and therefore is more flexible. This procedure is described in the steps given below:

a. Operate radar with the transmitter off to prevent damage to adapter, cables, or calibrated attenuator, and employ manual tuning of the receiver.

b. Use the same leakage test as for the measurement of mds described in paragraph 11-635. The signal generator may be either the fm or pulsed type.

c. Synchronize A-scope sweep with pulsing of signal generator. (The method of synchronization depends upon the radar used.) A synchroscope may be used for the

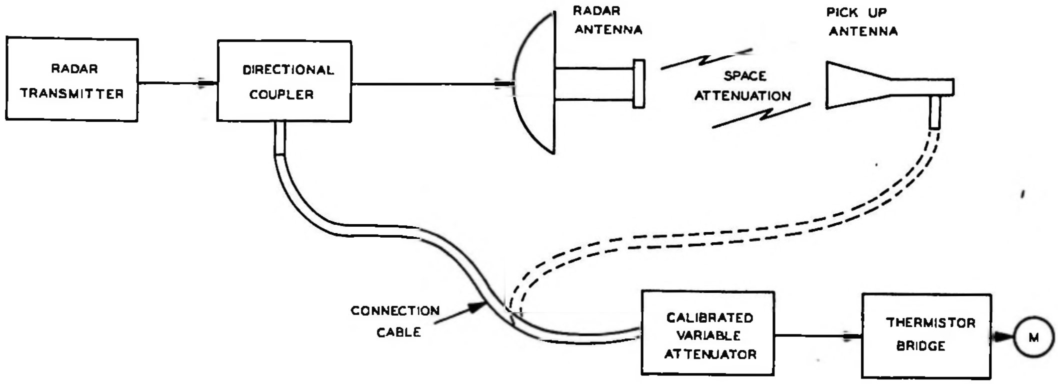


Figure 11-75. Test Setup for Pickup-Antenna Calibration—First Method

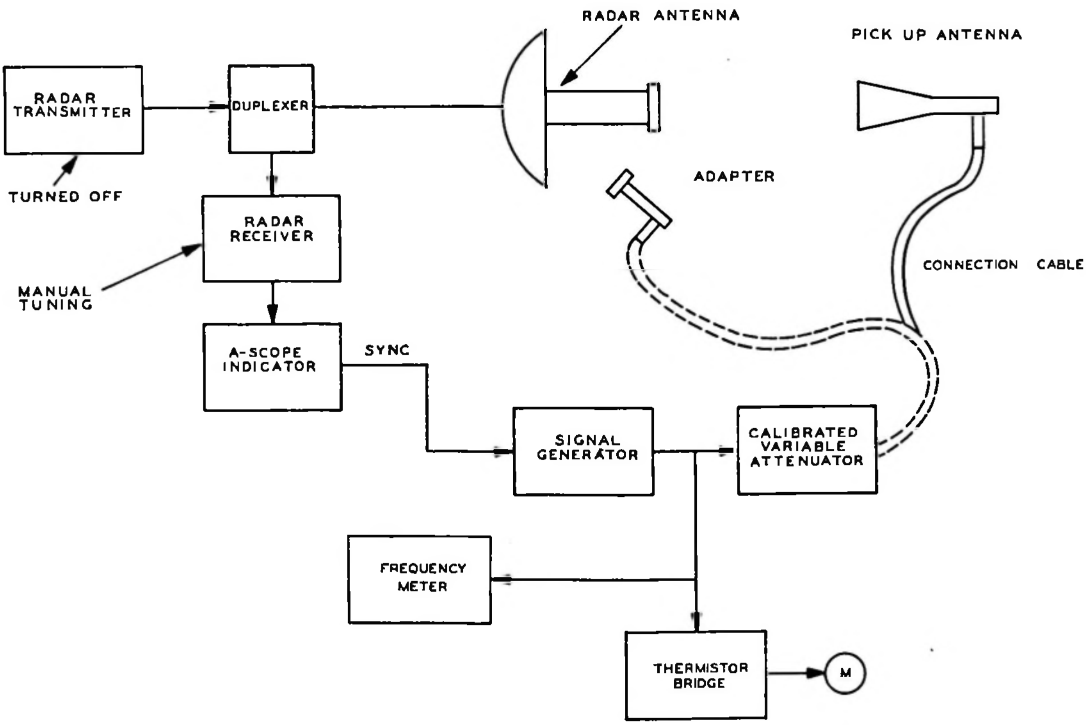


Figure 11-76. Test Setup for Pickup-Antenna Calibration—Second Method

A scope, in which case, the signal generator is synchronized by the same pulse used to trigger the synchroscope.

d. Install pickup antenna as described in step a of the first method.

e. Orient the two antennas so that maximum pickup is obtained.

f. Tune signal generator to desired frequency, establish a 1-mw cw level, and

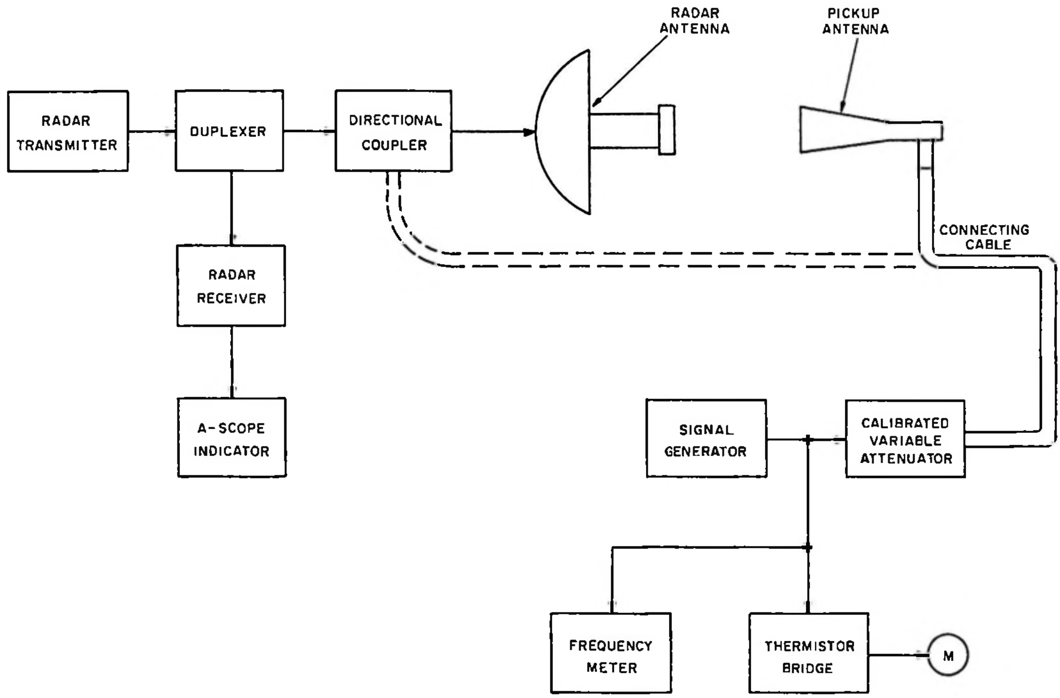


Figure 11-77. Test Setup for Pickup-Antenna Calibration—Third Method

operate generator for a pulsed output, or an fm output, if an fm generator is used.

g. Set receiver gain for about one eighth inch of noise on A scope.

h. Tune radar receiver to signal-generator frequency as indicated by maximum amplitude of received pulse. Check frequency with frequency meter.

i. Adjust attenuator to give about a half inch pulse. Be careful to check for saturation by reducing the attenuation 1 db. If the receiver is not saturating, a rise of about 12 percent can be expected.

j. Determine exact pulse height in either scale divisions or inches. Do not touch the receiver controls after this point.

k. Remove radar-antenna feed assembly, and substitute the adapter. Connect adapter to the connecting cable.

l. Increase attenuator setting until pulse amplitude obtained in step j is reached.

m. The change in attenuator setting is the space attenuation between the two antennas.

n. Repeat the measurement at several points in the operating-frequency band, and plot a graph of space attenuation versus frequency.

11-614. The following method shown in figure 11-77 is used when it is impractical to remove the radar antenna feed assembly. This method has other advantages in that the test is made with the radar transmitter turned on and no external sync is required to trigger the signal generator. The procedure is given in the steps below.

a. Install pickup antenna as described in the first method.

b. Orient both antennas for maximum pickup.

c. If the radar has no built-in directional coupler, install one temporarily. It is necessary that the calibration of this coupler be reliable.

d. Use the leakage test for mds measurement as described in paragraph 11-635. Either the fm or pulsed signal generator may be used.

e. With test equipment connected to directional coupler, set receiver gain to provide one eighth inch noise level on A scope.

f. Tune signal generator to desired frequency, and establish a 1-mw cw output.

g. Operate signal generator for pulsed or fm output, as the case may be, and tune receiver to frequency of signal generator. Check frequency by observing absorption response of frequency meter on A scope.

h. Adjust attenuator to provide about one half inch pulse, and check for saturation. Note attenuator reading.

i. Transfer the connecting cable to pickup antenna, and adjust attenuator to give same pulse amplitude as in step h. Note attenuator reading.

j. The difference between the two attenuator readings, combined with the directional-coupler loss, is the space loss.

NOTE

1. If readings in steps h and i are equal, the space attenuation is equal to the directional-coupler attenuation.

2. If reading in step h is greater than that in step i, space attenuation = directional-coupler attenuation + (step h reading — step i reading).

3. If reading in step h is less than that in step i, space attenuation = directional-coupler attenuation — (step i reading — step h reading).

k. Repeat the measurement at several points in the operating-frequency band and plot a graph of space attenuation versus frequency.

11-615. CABLE-ATTENUATION CALIBRATION. During a transmitter power test, it is a simple matter to calibrate the attenuation of connecting cables. The method outlined below requires the use of an additional connecting cable, whose attenuation figure need not be known, together with a coupling adapter. The procedure for the test is given in the following steps.

a. Use a radar with supplied directional coupler, if possible. If a pickup antenna is used, be careful not to disturb its position while the following steps are performed.

b. Connect thermistor bridge and calibrated-attenuator power meter to directional coupler (or pickup antenna), by means of a connecting cable. The attenuation figure for this cable need not be known.

c. Set calibrated attenuator to give 1-mw

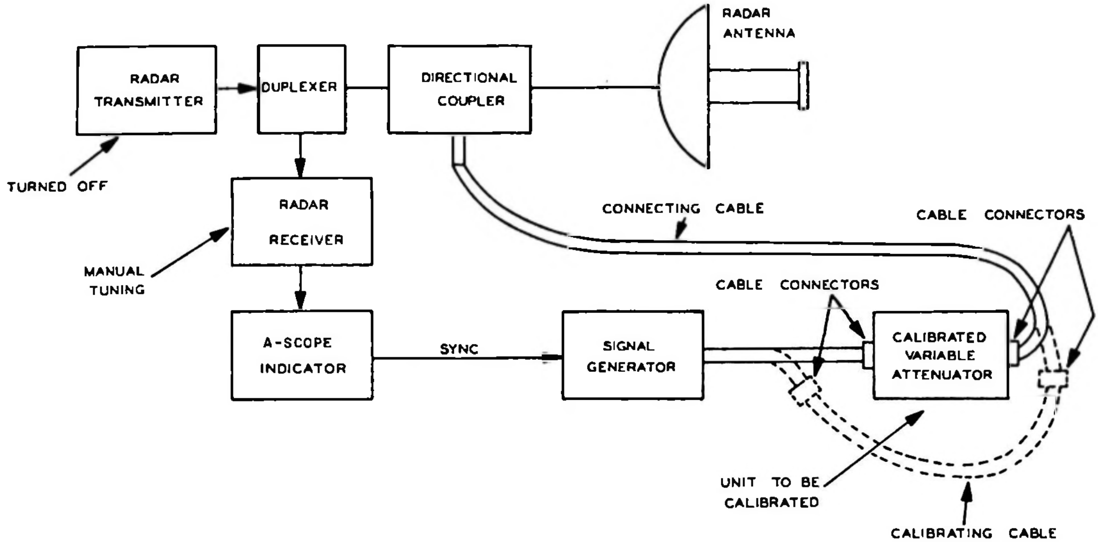


Figure 11-78. Test Setup for Attenuator-Zero-Loss Calibration

reference power level on power meter. Note the reading.

d. Using a coupling adapter, connect cable under test in series with the cable used in step b.

e. Check all connections for tightness.

f. Decrease reading on calibrated attenuator to produce the reference level on the meter. Note the new reading.

g. The difference in the readings obtained in steps c and f is the total attenuation of the coupling adapter and the cable under test. Since the loss of the coupling adapter is usually less than 0.1 db, it may be ignored. Therefore, the cable attenuation is the difference in the readings obtained in steps c and f.

11-616. ATTENUATOR CALIBRATION. The method of calibrating cable attenuation just described can also be used to calibrate attenuators of both the fixed and variable types. The use of this method requires that the attenuator under calibration have cable-type fittings, or that low-loss adapters be available. Since a cutoff-type attenuator is very nonlinear at low attenuation levels, a stop is provided to prevent operation in the nonlinear region. Therefore, when the attenuator dial reads zero, there is still a fixed loss, called "attenuator zero loss." Attenuators usually are furnished with calibrated charts that give zero loss versus frequency. When an attenuator is used, the dial readings must be increased by the value indicated on the chart for the particular frequency in use. Zero loss can be calibrated by turning the dial to zero and measuring the attenuation in the same manner as in cable attenuation. Each time a different frequency is used, the zero loss should be rechecked. Zero loss can be checked, by the test setup shown in figure 11-78, without using a second calibrated attenuator, by performing the steps given below:

a. To calibrate a cable, which is to be used as reference, use the test method described in paragraph 11-615. The loss in this cable should be greater than the attenuator zero loss. If necessary, lossy cable such as RG-21/U may be used.

b. Use either an fm or pulsed-type signal generator to provide the necessary signals.

c. With the transmitter not operating, synchronize the signal generator with the same pulse used to trigger the A scope.

d. If possible, use a directional coupler to feed the signals into the radar facility. An adapter or a pickup antenna may be used, but these are not so reliable.

e. Connect test setup as shown by the broken lines in figure 11-78.

f. Tune radar receiver and signal generator to the desired frequency.

g. Adjust the receiver gain to produce a pulse just a little over one half inch high on the A scope. Reduce signal-generator output until pulse is one half inch high, to prevent the possibility of limiting in the receiver.

h. Remove calibrating cable, and restore connections as shown by the solid lines in figure 11-78.

i. Adjust attenuator for half-inch pulse amplitude on A scope.

j. The attenuator zero loss is the difference between the cable loss and the attenuator dial reading.

k. Repeat the above steps using at least four different frequencies in the operating band, and make a graph showing attenuation versus frequency. It can be stated generally that attenuation increases with frequency.

11-617. RECEIVER PERFORMANCE TESTING.

11-618. The performance of a radar receiver is determined by a number of factors, most of which are involved and established in the design engineering of the equipment. In the text to follow, only those factors which are concerned with maintenance will be considered. The most important factors, which will be discussed in detail, are: receiver sensitivity, which includes noise figure determination and minimum-discernible-signal measurement; tr recovery time; receiver recovery time; and receiver bandwidth.

11-619. Many radar facilities have circuits which are included to serve a special function. Four of these special circuits commonly encountered are moving target indication (mti), instantaneous automatic gain control (iagc), sensitivity time control (stc), and fast time constant (ftc). These circuits may be found in combination or singly depending upon the purpose of the radar. In the test methods and procedures about to be described, the special functions should be disabled. If an automatic-frequency-control (afc) circuit is included in the radar equipment, it may be permitted to operate during receiver tests. A good check on afc is to make the tests specified for manual tuning, then switch to afc. If the afc circuit is normal, the signal indications should not change.

11-620. TESTING RECEIVER SENSITIVITY.

11-621. GENERAL. Inefficient range performance of a radar facility can result from troubles in the radar receiver. Loss of receiver sensitivity has the same effect on range as a decrease of transmitter power. For example, a 6-db loss of receiver sensitivity shortens the effective range of a radar just as much as a 6-db decrease in transmitter power. Such a drop in transmitter power is very evident in meter indications and, therefore, is easy to detect. On the other hand, a loss in receiver sensitivity, which can easily result from a slight misadjustment in the receiver, is very difficult to detect unless accurate measurements are made.

11-622. The sensitivity of the receiver determines the ability of the radar facility to pick up weak signals. Greater sensitivity then indicates that the receiver can pick up weaker signals. Sensitivity of a radar receiver is measured by determining the power level of the minimum discernible signal (mds). Mds is defined as the weakest signal that produces a visible receiver output, and its value is determined by the receiver output noise level, which tends to obscure weak signals. It follows, therefore, that an mds measurement is dependent upon the receiver noise level, and that measuring either one of these quantities will give an indication of receiver sensitivity.

11-623. NOISE ANALYSIS. In any conductor, there is a certain amount of random electron motion resulting from thermal agitation. This motion produces a voltage within the conductor that varies in a random manner. Since this voltage is a pure noise voltage, it produces signals that contain frequencies randomly distributed throughout the rf spectrum. The signals that occur in the portion of the rf spectrum covered by a given receiver are picked up and appear at the receiver as noise. The input power, in watts, represented by this form of noise is given by the formulas:

$$\text{Noise power} = KT\Delta F$$

$$\text{Open circuit noise voltage in series with the resistor} = 4KT\Delta FR$$

where:

K = Boltzmann's constant (1.37×10^{-23} watt-seconds (joules) per degree Kelvin)

T = Temperature in degrees absolute (Kelvin scale)

= (C^{o} + 273)

ΔF = Range of frequencies involved (bandwidth) in cycles per second

R = Resistance

11-624. The formula given above shows that thermal agitation noise is determined by bandwidth and temperature. The constant merely serves to convert the noise units into units of power. A decrease of temperature causes less random electron motion, and, at absolute zero, all motion and noise theoretically cease. Since noise covers all frequencies, it is apparent that a greater bandwidth encompasses a greater range of signals, and means more noise power. In a theoretically perfect receiver, this noise could be considered as a voltage across the antenna terminals and the power represented could be calculated on the basis of temperature and bandwidth. In practice, the actual noise developed in a receiver is greater than the calculated value because of the generation of other types of noise within the receiver circuits. For example, a carbon-type resistor, which is made up of fine particles of carbon, generates a noise signal when current flows through the resistor, because of small changes in the contact area of the particles. Various resistors have widely varying noise levels, and those that are used in the input circuits of a radar receiver must be chosen so as to have as low a noise level as possible. Electron tubes also generate noise signals, because of random variations in electron emission from the cathode, random variations in the current division between the plate and screen grid, etc. Since electron tubes produce noise in proportion to the number of electrodes employed, it follows then that triode tubes are generally used where noise limitation is an important consideration. Further information regarding noise analysis is given under paragraph heading 11-34.

11-625. NOISE FIGURE DETERMINATION. The term noise figure (nf) as applied to a radar receiver, indicates the amount of noise that is to be expected. Nf is defined as the ratio of measured noise to calculated noise, and may be expressed as a power ratio or in db. Therefore, nf can be said to be the input signal-to-noise ratio in comparison to the output signal-to-noise ratio and since the input ratio is larger than the output ratio, nf is always greater than 1. The input should be at a temperature of 290 degrees Kelvin.

11-626. In the microwave range of operation, virtually all of the noise originates within the receiver. Atmospheric and man-made noise or static is normally too small to be considered. The three main sources of noise in a radar receiver are the crystal mixer, the i-f preamplifier (usually the first two i-f stages), and the local oscillator.

11-627. If the noise figure of a certain radar, when compared to its normal performance standard, is too high, you should take corrective measures. The general procedure is given in the following discussion. First, the crystal in the radar is replaced with another one and the noise figure is rechecked. In practice a large number of crystals may be checked by substitution, and the one with the lowest nf is used. The same procedure is then applied to the i-f preamplifier tubes. If the nf is still too high, the local oscillator tube is replaced. It is interesting to note that the output noise of a reflex klystron is much greater than normal when the tube is tuned off the center of a mode.

11-628. Early radar receivers had noise figures in excess of 20 db while modern receivers have noise figures of only 6 to 18 db. In general, lower receiver frequencies result in lower noise figures. The noise figure of a radar receiver can be determined by the use of either a noise generator or a cw signal generator.

11-629. TEST METHOD USING NOISE GENERATOR. A noise generator produces a random noise signal which covers a frequency range in excess of the radar bandwidth. One such instrument uses a temperature-limited diode, operated at saturation, as the noise-signal source. When a diode is operated under these conditions, the noise produced is proportional to the dc plate current, the generator frequency range must be adequate, and the diode cannot be used for over 1000 mc. Therefore, the dc input power can easily be converted to obtain the true noise power. The procedure for determining the noise figure of a radar receiver, using a noise generator, is given in the following steps:

a. Connect a milliammeter (0—1 ma) in series with the diode load of the second detector of the receiver.

b. Terminate the receiver input in an impedance equal to the normal source impedance. Adjust receiver gain to produce a 0.5-ma reading. This reading results from noise alone. Remove receiver input from termination and attach generator to the receiver input. Decrease the generator noise excess to zero.

c. Connect noise generator to receiver input.

d. Adjust output of noise generator until meter reads 0.707 ma (1.4 x 0.5). At this point make certain that a further increase in noise causes a corresponding increase in meter reading. If this does not happen, then the receiver is limiting and the readings will not be accurate. In this case, use a lower value of current in step b; for example, start with 0.3 ma and increase this value to 0.42 ma (0.3 x 1.4) in step d (detector is assumed to be linear).

e. The noise generator power output is now equal to the receiver noise power. Note the dial reading. A chart is usually furnished with the instrument for converting the dial reading to power.

f. Calculate the noise figure by using the following formula:

$$NF\ (db) = 10 \log \frac{P\ \text{measured}}{P\ \text{calculated}}$$

where "P measured" is the amount of noise being fed to the receiver by the noise generator. This figure must be in picowatts. "P calculated" is the figure arrived at by using the following formula:

$$\text{Noise power} = 4KT\Delta F \text{ or } 20\ I_{dc}R_s$$

where:

K = Boltzmann's constant (1.37×10^{-23} watt-seconds (joules) per degree Kelvin)

T = Ambient temperature of the equipment under test measured in degrees Kelvin ($^oC + 273$)

ΔF = Receiver bandwidth (in cycles per second)

I_{dc} = Dc plate current of the noise diode

R_s = Source resistance of the antenna

As an example, to calculate the noise figure of a receiver having a bandwidth of 4 megacycles (4×10^6), and ambient temperature of the equipment in degrees centigrade at 20 (20 + 273):

$$NP = 4KT\Delta F$$

$$= 4 \times 1.37 \times 10^{-23} \times (20 + 273) \times (4 \times 10^6)$$

$$= 6422.56 \times 10^{-17}$$

$$= 0.06423 \text{ picowatts}$$

The measured noise power being fed to the receiver under test is 1.018 picowatts. Using the formula below, the noise figure of the receiver is arrived at as follows:

$$NF\ (db) = 10 \log \frac{P\ \text{measured}}{P\ \text{calculated}}$$

$$NF\ (db) = 10 \log \frac{1.018}{0.0642}$$

$$= 10 \log 15.85$$

$$= 10 \times 1.2$$

$$= 12 \text{ db}$$

The noise figure for the receiver under test is calculated as 12 db, this being a normal noise figure for present-day radar receivers.

11-630. A gaseous discharge tube may be used as a noise source. The noise figure (nf) is doubled when the noise generated is turned on and is given by:

$$NF_{(db)} = 15.4 - \text{Attenuation (db)}$$

Figure 11-79 illustrates this procedure when the noise figure is greater than 15.4 or when you are using a calibrated output power meter.

11-631. TEST METHODS USING CW SIGNAL GENERATOR. The cw method of measuring

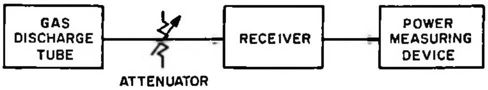


Figure 11-79. Gas Discharge Tube Noise Source

the noise figure uses a calibrated signal generator in the same manner as the noise generator. This method is not so accurate as the noise-generator method because the detector characteristics of the receiver under test may affect the ratio of signal power to noise power. Further procedures and considerations concerning these methods are included in paragraph 11-42.

11-632. MINIMUM DISCERNIBLE SIGNAL MEASUREMENT. The measurement of a minimum discernible signal (mds) consists of measuring the power of a pulse whose level is just sufficient to produce a visible receiver output. It follows that if a radar receiver has the specified mds level, the noise figure should be correct also. Therefore, measurement of the mds is a satisfactory substitute for a noise-figure determination, and is less complicated. Correct pulse length must be used, and when readings are taken periodically for comparison purposes, the identical pulse length must be used each time.

11-633. RF LEAKAGE DETERMINATION. In the measurement of mds, a very high degree of attenuation (approximately 98 db for the average radar) and a very low power level (about one picowatt) are involved. Because of these factors, very little rf leakage from the signal generator can be tolerated, or the amount of leakage signal picked up by the receiver will be appreciable compared to the signal fed through the attenuator. Since leakage signals are independent of attenuator setting, very inaccurate mds readings can be obtained when leakage is present. If the leakage signal reaches the receiver in phase with the signal through the attenuator, the mds reading will be low, and thus will indicate that the receiver sensitivity is much better than it actually is. In such a case there is a good possibility that a defective receiver may appear to be normal. On the other hand, if the leakage signal reaches the receiver out of phase with the signal through the attenuator, the mds reading will be high, and thus will indicate that the receiver sensitivity is poorer than it actually is. In the construction of a signal generator, special attention is given to the problem of minimizing rf leakage. The rf oscillator is carefully shielded, and then it and the attenuator assembly are enclosed in a second shield which serves as the case of the instrument. In addition, all connecting cables and couplings are provided with shields and close-fitting connectors. In spite of these precautions, however, a small amount of leakage exists, even in the most modern equipment.

11-634. The presence of leakage makes it imperative to locate all equipment associated with mds tests outside the radar-antenna radiation field. In addition, the equipment should never be operated outside its case, or with loose cable connections. Also, on early signal generators where a door is provided on the front panel for access to the oscillator adjustments, the door must be kept closed during measurements. These precautions must be observed; otherwise erroneous results will be obtained.

11-635. LEAKAGE DETECTION METHOD. The presence of rf leakage is detected by the following method. Determine the mds level, then move the test set to another position and determine the mds level at that point. Observe whether the first mds reading differs from the second; it it does, leakage is present in one of the two positions. When leakage is found to be present, locate the test set as far from the radar antenna and receiver as possible. Find that posi-

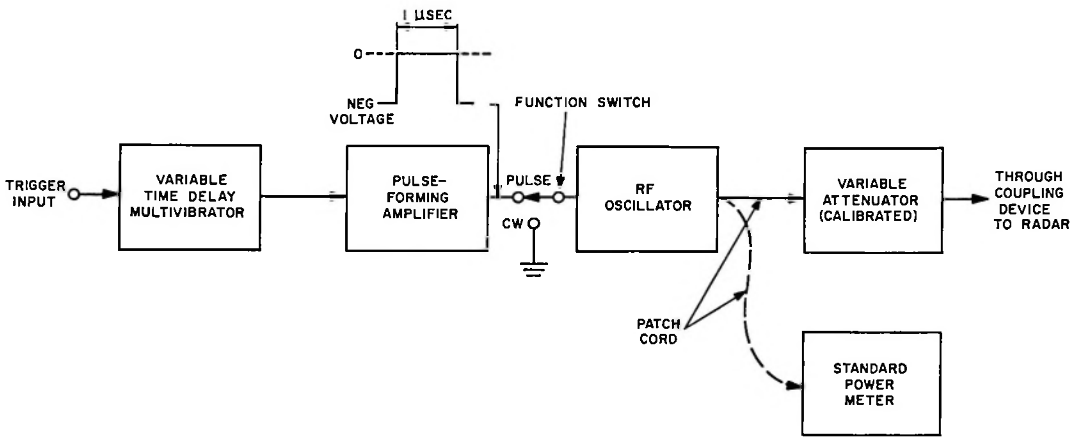


Figure 11-80. Early Type MDS Measurement Using Pulsed RF Signal Generator

tion where movement of the test set does not affect the mds. In general, if rotation of the test set does not change the mds level, the rf leakage can be considered negligible.

11-636. MDS MEASUREMENT USING PULSED-SIGNAL GENERATOR, EARLY METHOD. Early mds test equipment consisted of a pulsed-type rf signal generator, which is shown in block diagram form in figure 11-80. In this generator, the radar transmitter trigger pulse is used to trigger a start-stop multivibrator, the period of which can be adjusted to any value from about 0.5 to 50 microseconds. The trailing edge of the multivibrator output pulse triggers a pulse-forming amplifier, which produces a 1-microsecond pulse to modulate the rf oscillator when the function switch is in pulse position. This modulating pulse is a positive-going pulse with a peak at zero voltage level. For cw operation, the function switch grounds the rf oscillator. This arrangement ensures that the peak pulse power is the same as the cw power. The rf oscillator is usually a reflex klystron and is coupled to either an attenuator or a power meter, for calibration, by means of a flexible patch cord. Early pulsed-type signal generators were coupled to the radar with a pickup antenna, but a directional coupler is more convenient and accurate. With a directional coupler, all of the power inserted into the transmission line is directed toward the receiver and away from the antenna. This is not true with an rf probe. When a probe is used, half of the power is lost to the antenna.

11-637. When an early type equipment is used, the procedure for making an mds measurement is given in the following steps:

a. Connect radar trigger-pulse output to trigger input of signal generator.

b. Connect output of attenuator to pickup antenna, which should be placed as described in paragraph 11-571.

c. Set function switch to cw position, and tune the rf oscillator to the frequency of the radar receiver using techniques discussed under paragraph heading 11-508.

d. Feed rf oscillator output to power meter, and adjust oscillator coupling to pro-

A
0-DB ATTENUATION

B
10-DB ATTENUATION

C
15-DB ATTENUATION

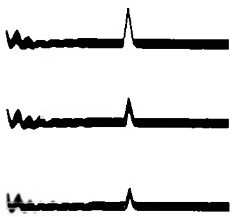


Figure 11-81. Artificial Echo Presentation on A Scope

duce standard 6-mw deflection on power meter.

e. Return patch cord to variable attenuator and set function switch to pulse.

f. Adjust receiver gain to produce a fairly large noise output (quarter inch on A scope).

g. Using a low attenuator setting, observe artificial echo pulse on radar A scope. If the radar has no A scope, connect a synchroscope to the receiver output and trigger it with the same pulse applied to the trigger input.

h. If artificial echo coincides with a target, adjust variable time delay to position the echo at a range where no targets are present.

i. Tune radar receiver for maximum echo.

j. Increase attenuation until artificial echo is just barely visible in the noise. Refer to figure 11-81. Part (A) of this figure represents a strong signal, part (B) represents the same signal attenuated 10 db, and part (C) represents the same signal attenuated an additional 5 db and approaching the mds level. The echo pulse can be distinguished from the noise more easily during the final adjustment of the attenuator if the time delay is varied slightly.

k. Find the total attenuation by adding the attenuator reading (db), the space loss (db), the attenuator zero loss (db), and the cable loss (db).

The mds figure is in db below 6 mw and, if desired, it may be converted to power by the use of the conversion chart of figure 11-54. The space loss and attenuator zero loss will be the same as for a power measurement.

11-638. MDS MEASUREMENT USING PULSED-SIGNAL GENERATOR, MODERN METHOD. A typical modern, pulsed-signal generator is shown in block diagram form in figure 11-82. The pulser circuit is similar to that of the older type which is shown in figure 11-80, except for the internal sync circuit. To make internal sync possible, a small part of the rf output pulse is applied to a crystal detector, and the resulting rectified output signal is made available as a trigger to the pulser circuit. Thus the radar pulse can produce sync automatically, and, if desired, external sync may be used. The rf oscillator may operate in either cw or pulsed condition, and it may be turned off when desired. With the oscillator off, the equipment may be used to measure transmitter power and frequency. The purpose of the uncalibrated variable attenuator, sometimes called the "power set," is to drop the output of the rf oscillator to the standard 1-mw level used in modern test sets. The thermistor bridge monitors the power entering the calibrated attenuator. Calibration of this attenuator is such that the dial reading includes the zero loss.

11-639. An mds measurement, using a modern pulsed signal generator, is given in the following steps.

a. If internal sync is not to be used, connect radar trigger pulse to trigger input.

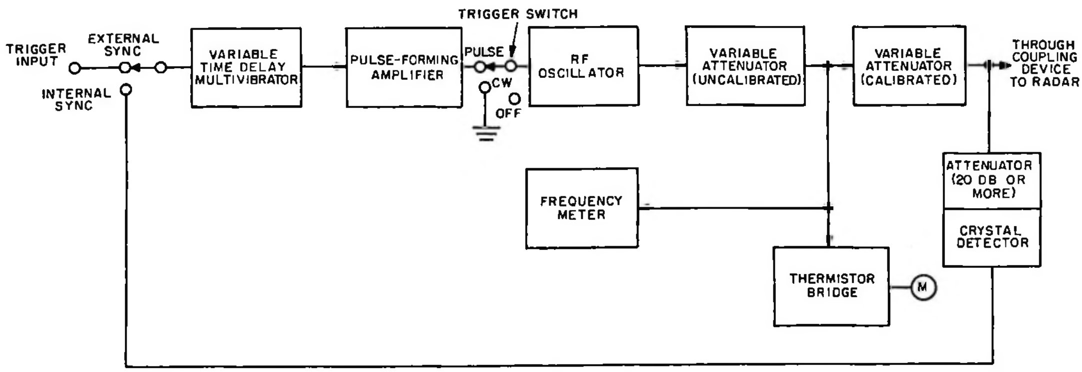


Figure 11-82. Modern-Type MDS Measurement Using Pulsed RF Signal Generator

b. Connect attenuator output to coupling device. A directional coupler is usually employed, but a pickup antenna may be used.

c. Set function switch to the off position and zero the thermistor bridge.

d. Turn function switch to cw and adjust uncalibrated attenuator for midscale reading on thermistor bridge.

e. Using the frequency meter, adjust the rf oscillator to the radar frequency. The frequency meter should be kept detuned, except during frequency checks.

f. Set the uncalibrated attenuator accurately for a 1-mw bridge reading.

g. Set function switch to the pulse position.

h. Adjust the calibrated attenuator for the mds indication previously described in the early test method. The variable time delay must be set so that the artificial echo does not occur at the same range as a radar target.

i. Find the total attenuation. The value obtained is the receiver mds in terms of —dbm. Total attenuation = coupling loss + cable loss + attenuator reading (all in db). If desired the —dbm value may be converted into terms of power by utilizing the chart shown in figure 11-54, but the dbm reading conveys more information to the technician.

11-640. MDS MEASUREMENT USING FM SIGNAL GENERATOR. When a pulsed-signal generator is used for the measurement of mds, one major disadvantage presents itself; that is, the accuracy of the results depends upon how accurately the signal generator is tuned to the radar frequency. This difficulty has been overcome in the design of an fm signal generator incorporated in a test set as shown in figure 11-83. The rf section of this signal generator is very similar to that of the pulsed generator shown in figure 11-82. Since the rf oscillator is a reflex klystron, the oscillator frequency may be varied by means of a sawtooth voltage fed to the repeller plate of the tube. This voltage is developed by a sawtooth generator, which is activated by a trigger pulse. The sawtooth voltage rise is nearly linear and lasts for about 50 microseconds. This voltage is fed to a circuit shown in basic form in figure 11-84, which contains two controls. The signal-width control determines the amplitude of the sawtooth fed to

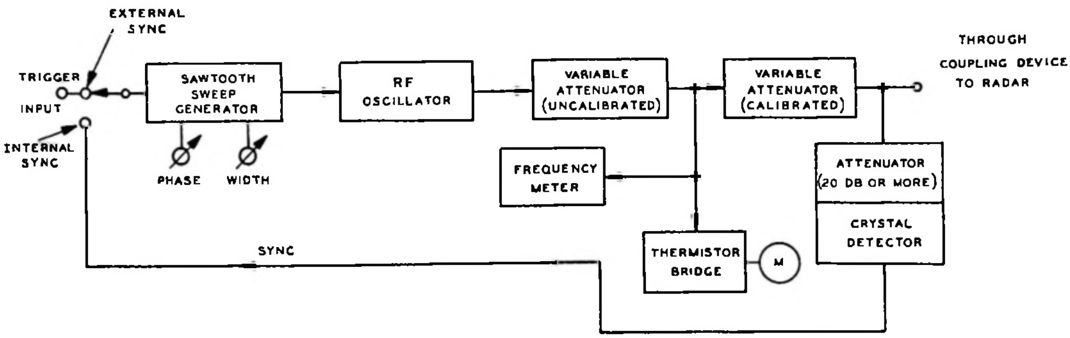


Figure 11-83. MDS Measurement Using FM Signal Generator

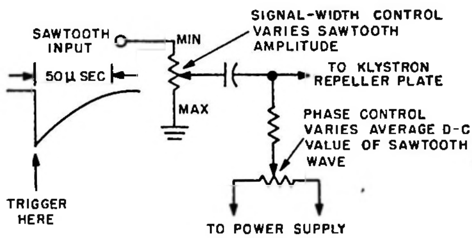


Figure 11-84. Signal-Width and Phase-Control Circuit

the klystron, and the phase control determines the level of the sawtooth by supplying a variable negative dc voltage to fix the operating point of the klystron repeller. When the signal-width control is in the maximum position, no sawtooth voltage is applied to the repeller and the signal generator supplies a cw output. Hence, with the signal-width control in this position, the klystron mode pattern can be explored by manually varying the repeller voltage with the phase control. In operation, the fm generator creates an artificial echo pulse by rapidly sweeping an rf signal through the receiver pass band at a given time after the transmitter fires. This action is shown in figure 11-85. The transmitter pulse starts the sawtooth sweep which varies the signal frequency. As the generator frequency is swept through the receiver pass band, a plot of frequency versus output is obtained, with the result that a pulse, very similar to an echo pulse, is reproduced on the radar A scope. The width of the pulse obtained depends upon the receiver bandwidth and the rate at which the frequency is swept. For example, if the receiver bandwidth is 4 mc and the klystron frequency is swept at a rate of 4 mc per microsecond, the signal will be within the frequency range of the receiver pass band for only one microsecond; hence, the pulse seen on the A scope will be one microsecond wide. The effect of decreasing the amplitude of the sawtooth voltage is shown in figure 11-85. A decrease in amplitude allows the frequency to be swept at a lower rate and thus results in a wider pulse on the A scope. By varying the signal-width control, the desired pulse width can be obtained. When measuring mds, the pulse width is usually made equal to the transmitter pulse width; however, this is not a critical factor.

11-641. The effect of varying the phase control is shown in figure 11-86. The sawtooth waveform shown in part (B) of this figure represents a more negative average voltage. Since the sawtooth rises in a positive direction, it will take longer for the voltage level to reach the point at which the echo pulse occurs. Therefore, the average dc voltage level determines how much time elapses between the trigger pulse and the echo pulse.

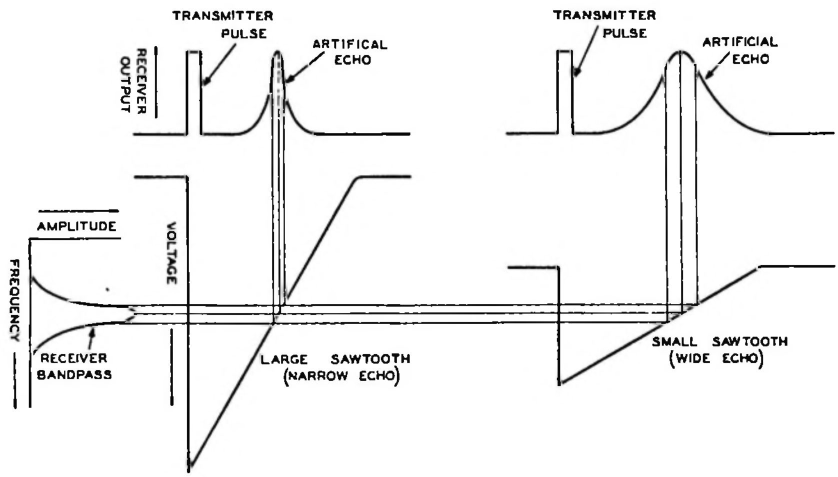


Figure 11-85. Effect of Sawtooth Amplitude on Presentation of Artificial Echo

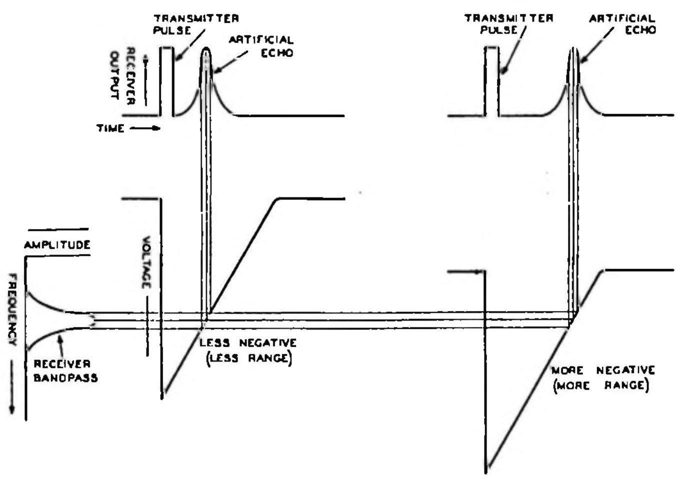


Figure 11-86. Effect of Sawtooth Level on Presentation of Artificial Echo

Hence, the phase control is really a range control, although it is seldom labeled as such. The higher the average negative voltage, the greater the echo range. You must remember that each klystron mode produces an echo indication. Therefore, by increasing the amplitude of the sawtooth voltage sufficiently with the signal-width control, it is possible to get a series of echoes, one for each mode, with the one for the most negative mode having the shortest range on the radar A scope. However, for normal operation, only one mode is used.

11-642. When an fm signal generator is used, the procedure for measuring mds is given in the steps below:

a. Connect radar trigger-pulse output to trigger input jack on fm signal generator. Omit this step if internal sync is to be used.

b. Connect rf input through coupling device to radar. (A directional coupler is preferred.)

c. Set signal-width control to maximum.

d. Adjust phase control for maximum klystron output as indicated on thermistor bridge.

e. Tune klystron (cavity) to approximate radar frequency by adjusting the frequency meter to the frequency of the radar transmitter and tuning the klystron for a dip in the thermistor-bridge meter reading. Since the klystron mode is fairly broad, extreme accuracy in tuning the klystron is not necessary. For example, the width of the flat portion of an X-band klystron is about 10 mc; therefore, the tuning accuracy required is $\pm$ 5 mc.

f. If necessary, adjust the phase control again for maximum output. If a sizable adjustment is required, repeat step e.

g. Adjust uncalibrated attenuator for a 1-mw indication.

h. Set receiver gain control for a quarter inch noise level on the A scope.

i. Reduce signal-width control setting until the pulse is seen on the A scope. This will have to be done at a low setting on the calibrated-attenuator control.

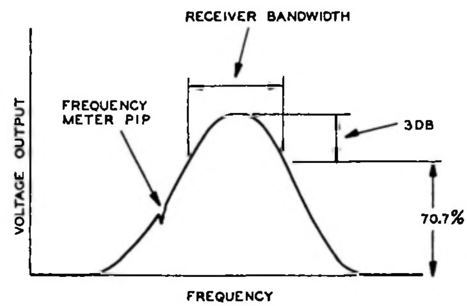


Figure 11-87. Receiver Response Curve

j. If necessary, adjust phase control to position echo pulse in a target-free area.

k. Adjust signal-width control for desired echo-pulse width.

l. Set calibrated attenuator for the mds previously described. Rock the phase control during this step to distinguish the echo more easily.

m. Find the total attenuation in db. The value obtained is the mds in db below 1 mw (—dbm). Total attenuation = coupling loss + cable loss + attenuator reading (all in db).

11-643. TESTING RECEIVER BANDWIDTH.

11-644. GENERAL. Receiver bandwidth is defined as the frequency spread between the half-power points on the receiver response curve. Receiver bandwidth is specified for each radar, but wide variations are often tolerated. If either the bandwidth or the shape of the receiver response curve is not correct, you can remember that a considerable change in the value of circuit components is required to alter the response materially. It is suggested that the receiver response be checked after an extensive repair to any i-f amplifier. Figure 11-87 shows a typical response curve of a radar receiver. The half-power points are shown as 3 db below maximum (mid-frequency) response. Since the curve is plotted in terms of voltage, these points are also represented by the 70.7-percent voltage points ($\sqrt{1/2} = 0.707$) as shown in the figure.

11-645. INTEGRATED RECEIVER METHOD. The bandwidth test procedure given below is used when the receiver is operating as an integral part of the radar facility, and can very easily be made after checking the mds. When the radar receiver is to be tested as an individual component, another test procedure is used which will be described immediately following this method.

11-646. This method, which is considered superior to other methods, makes use of the test setup for measuring mds, using an fm signal generator as previously described in paragraph 11-640. The procedure is given in the following steps:

a. With the equipment connected in the same manner as for an mds measurement, turn the signal-width control to obtain a response curve about one-half inch wide.

b. Reduce receiver gain so that the noise amplitude is just barely visible.

c. Adjust calibrated attenuator to produce a pulse amplitude below receiver saturation level.

d. Tune frequency meter until response curve shows an absorption pip at one of the half-power points. Read the frequency, then repeat for the other half-power point. The difference between these two frequencies is the receiver bandwidth.

11-647. When the foregoing procedure is used, the half-power points may be located very easily as outlined in the following steps:

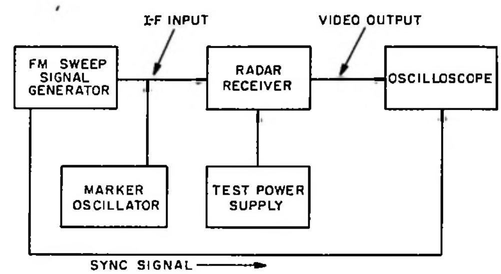


Figure 11-88. Test Setup for Checking Receiver Response

a. Note the attenuator dial reading following step c given above.

b. Increase the attenuator reading 3 db and mark the level at the top of the response curve.

c. Return the attenuator to the previous setting.

d. The half-power points are at the level marked in step b.

11-648. DETACHED RECEIVER METHOD. This method is employed when the radar receiver is to be tested as an individual component rather than part of the radar facility. Figure 11-88 shows the test setup for checking a receiver which is detached from the radar facility. A sweep generator produces a variable-frequency signal that is fed into the receiver i-f input.

NOTE

The sweep width and the center frequency of an fm signal generator is usually adjustable to cover any standard radar intermediate frequency.

The receiver video output is fed to the vertical-deflection circuit of an oscilloscope.

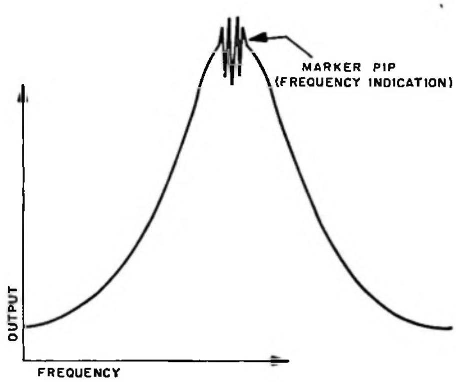


Figure 11-89. Response Curve, Showing Marker Pip at Mid-Frequency Point

In addition, a sync voltage is supplied by the sweep generator to maintain horizontal motion of the electron beam in synchronism with the frequency sweep. The oscilloscope, therefore, indicates frequency horizontally and receiver output vertically. A second signal generator, called the marker oscillator, produces an accurately calibrated cw signal which is mixed with the sweep generator output. When the varying sweep passes the marker-oscillator frequency, a beat signal results, producing a marker pip on the response curve as shown in figure 11-89. The marker-oscillator dial indicates the frequency at which the pip occurs.

11-649. To check receiver bandwidth using the test arrangement discussed above, the marker pip is positioned until it rests at the 70.7-percent point on the curve and the frequency dial is read. The frequency at the other half-power point is determined in the same manner. The spread between these two points, expressed in frequency, is the measured bandwidth.

11-650. TESTING TR RECOVERY TIME.

11-651. GENERAL. The time required to permit tr recovery is determined by the time it takes the tr switch to deionize after each transmitter pulse. It is usually defined as the time required for the receiver to return to within 6 db of normal sensitivity after the end of the transmitter pulse. However, some manufacturers use the time required for the sensitivity to return to within 3 db of normal or to full sensitivity. Tr recovery time is one of the factors that limit the minimum range of a radar because the radar receiver is unable to receive signals until the tr switch deionizes. The recovery time may vary from about 3 to 20 microseconds, depending upon the radar set.

11-652. The primary function of the tr section is to protect the crystal detector from the powerful transmitter pulse. Even the best tr switches allow some power to leak through, but, when the switch is functioning properly, the leakage power is so small that it does not damage the crystal. It has been found, however, that the useful life of a tr tube is limited because the amount of leakage and the recovery time increase with use. To ensure efficient performance, some technicians make it a policy to replace the tube after a given number of hours. A better practice is to measure the tr recovery time at frequent intervals, as called for in preventive maintenance procedures, and make a graph or chart, which will immediately disclose any change in performance. Figure 11-90 shows in an approximate manner how recovery time is correlated with leakage power. Note that the end of the useful life of the tr tube is indicated by an increase in recovery time. This method of checking the condition of a tr tube is reliable, because the recovery time increases before the leakage power becomes excessive. In practice, the tr tube is replaced when any sharp increases in recovery time become apparent. Ambient temperature has an effect on recovery time. The colder a tr tube, the greater is its recovery time. For example, the tube type 721A recovers in about 7 microseconds at 28 degrees C; how-

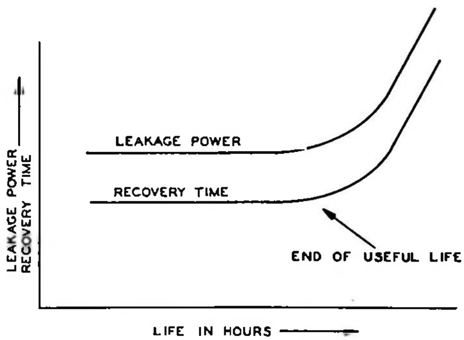


Figure 11-90. Graphic Comparison of TR Recovery Time and Leakage Power

ever, at —186 degrees C, the recovery time is about 100 microseconds, and at —20 degrees C, the recovery time is about 14 microseconds. When tests are conducted under widely varying temperature conditions, this effect must be considered.

11-653. CURRENT AND VOLTAGE CHECKS. One method used in testing a tr tube is to measure the keep-alive current. This current maintains the tr tube partially ionized to make the firing more reliable, and thus helps protect the crystal. The current is usually about 100 microamperes, and falls off as the end of the tr tube life approaches. Another method is to measure the keep-alive voltage between the plate and ground of the tr tube when the voltage is known to be good, and to record this voltage for use as reference for future checks. However, these checks are not so reliable as a recovery-time test.

11-654. PULSE OR FM SIGNAL GENERATOR METHOD. Tr recovery time can be tested by means of a setup that uses either an fm or a pulsed signal generator. When an fm or pulsed-signal generator is used, the tr recovery time test is conducted by performing the steps given below:

a. The same test setup used for the measurement of mds is utilized for this test (figure 11-82).

b. Set receiver gain to indicate about an eighth inch noise on A scope.

c. Adjust calibrated attenuator to give a pulse amplitude about halfway between the noise level and saturation. Note the attenuator reading.

d. Reduce attenuator setting by 6 db.

e. Rotate phase control (time delay) to position the pulse closer to the transmitter pulse. Continue rotation of control until the pulse amplitude drops to the level established in step c.

f. Read the range at which the pulse is now located. The value obtained is the "6 db recovery time" or the recovery time to within 6 db of normal receiver sensitivity. Recovery time may be indicated in either microseconds, miles, or yards as long as subsequent readings are in the same units.

11-655. If you desire, the test procedure given above can be modified so that the "full-sensitivity tr recovery time" is measured. Steps d and e are modified as follows:

d. Omit.

e. Using the phase control, move pulse toward the transmitter pulse until the pulse amplitude just starts to decrease. The use of this modified procedure will result in a longer recovery time, but when the results of a series of measurements are plotted, the curve obtained will be similar to the one shown in figure 11-90. Consistency is the most important factor; therefore, the type of recovery measurement used should always be noted in the maintenance records.

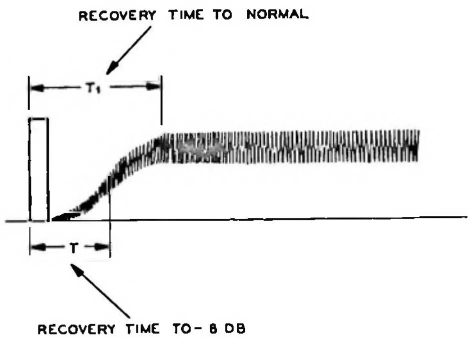


Figure 11-91. TR Recovery Test Indication Using a CW Signal

11-656. CW SIGNAL GENERATOR METHOD. It is also possible to test tr recovery time by using a cw signal generator. The test is conducted by performing the steps listed below:

a. Connect signal generator to coupling device.

b. Adjust radar-receiver gain to indicate about one quarter-inch of noise on A scope.

c. Tune signal generator to the frequency of the radar receiver. Proper tuning will be evidenced by a rise in the A scope trace.

d. Adjust output of signal generator (calibrated attenuator) to a point just below receiver saturation. The indication should now appear similar to that shown in figure 11-91.

e. Measure the range between the transmitter pulse and the point on the A scope where the noise amplitude is one-half of the maximum noise level.

11-657. The procedure described above gives the tr recovery time to within 6 db of normal. If the full sensitivity tr recovery time is desired, the time is measured to the point where the noise just barely reaches full value. In many cases, nearby targets will interfere with the testing of recovery time. When this occurs, position the radar antenna so as to point it at free space. In case this still does not eliminate the interference, the use of an absorption screen or a dummy rf load is recommended.

11-658. TESTING RECEIVER RECOVERY TIME.

11-659. Radar-receiver recovery time is defined as the time required for the receiver sensitivity to return to normal after a saturation echo is received. This time is determined in the original radar design and is of very short duration. Receiver recovery is not discussed in terms of minimum range since tr recovery is much longer. The receiver recovers from a transmitter pulse long before the tr tube recovers. Figure 11-92 illustrates the effect of receiver recovery. Note that immediately following the echo pulse the noise is at reduced amplitude, and that the recovery time is the period of reduced noise level. No absolute measurement of receiver recovery is necessary. A noticeable time interval, however, usually indicates trouble, which is quite often caused by an open grid-leak circuit.

11-660. CHECK OF RECEIVER OR AFC CRYSTALS.

11-661. In most modern microwave radar facilities, two silicon crystal mixers are provided, one for signal mixing and one for afc mixing. This practice has universally supplanted the earlier effort to use a common mixer for both functions. In a mixer application, the efficiency of the crystal is related to the conversion loss and the noise that the crystal introduces into the receiver. In addition, certain radar equipments contain these types of crystals in other circuits. When used as a mixer, the silicon crystal diode is one of the primary factors in estab-

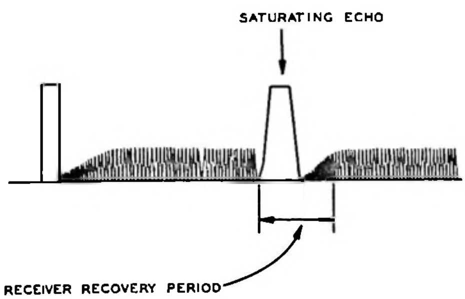


Figure 11-92. Receiver Recovery Indication

lishing the sensitivity or the minimum determinable signal strength of the radar receiver.

11-662. Conversion loss and noise figure are functions of the rectified crystal current and are determined by the local oscillator signal level. Noise figure increases and conversion loss decreases as the rectified crystal current increases. These crystal characteristics have been correlated to the front to back resistance ratio of the crystal and the back crystal current at one volt. Therefore, a measurement of this ratio and of the back current become indications of the crystal condition. In most cases, the most expedient and successful check of a mixer crystal is to replace it with a new or known good one. However, specific checks and adjustments can be made of the individual crystal and these checks are given in detail under paragraph heading 10-676, Volume VIII.

11-663. CW (DOPPLER)-TYPE RECEIVER CONSIDERATIONS.

11-664. Since the maximum Doppler frequency (difference frequency) is dependent upon the maximum relative velocity between target and receiver, the bandwidth of the receiver need only be wide enough to pass the entire expected band. Ordinarily the range of the Doppler frequency (for presently used equipment) is from about 1 to 10 kilocycles. The receiver operates in the microwave region and a filter gate (selective filter) generally follows the video amplifiers which effectively reduces the bandwidth and allows only the range of Doppler effected frequencies to pass to increase the signal-to-noise ratio.

11-665. In this type radar, the same antenna is generally used simultaneously for transmitting and receiving so that the receiver must be isolated from the transmitter. The transmitter energy is supplied to the waveguide thruplexer plane-polarized so that the electric vector (plane of polarization) is parallel to the vertical wall of the thruplexer; the receiving crystal detector is sensitive only to energy reflected in the horizontal plane. Receiver tests of this type equipment therefore require that energy being fed into the receiver be properly polarized.

11-666. The output of the receiver is limited before being detected so that the dc output will be directly proportional to the frequency deviation. The dc voltage can then be sent to a counting-type indicating device.

11-667. This equipment is rarely used for detecting distant targets; therefore, it is seldom necessary to amplify exceedingly weak signals. However, since the gain of the receiver must be sufficient to cause limiting of all signals received for accurate operation, the sensitivity of the receiver should be checked periodically. The sensitivity of this receiver can be checked with an fm signal generator set for an appropriate deviation modulation. When possible, the signal is fed through the antenna so that the sensitivity of the entire radar set can be checked. The output indication is best obtained from the grid of the limiter. If an fm signal generator is not available, a standard generator may be used, but the receiver sensitivity should be checked at several points throughout the Doppler range. Procedures for checking Doppler receiver sensitivity are the same as for communications receivers and are included under paragraph heading 11-78.

11-668. Cw radar equipment usually operates in the microwave region and most of the noise is inherent in the receiver. Therefore, a much better method of determining the performance of the receiver is to take a noise figure measurement, since noise is the limiting factor in determining maximum sensitivity. Besides the considerations and procedures regarding noise-figure measurements that are given under paragraph headings 11-34 and 11-620, you should note that the bandwidth of a cw radar receiver should be taken to include the filter gate (selective filter) since a much higher figure can be obtained from the resulting narrowing of the bandwidth. The output measurement is obtained by measuring the grid current of the first limiter tube.

11-669. TRANSMITTER PERFORMANCE TESTING.

11-670. The performance of a radar transmitter is determined by several factors, most of which are evolved and established in the design engineering of the equipment. In the text to follow, only those factors which are concerned with maintenance will be considered. The most important factors which will be discussed are measurement of the magnetic field of the magnetron magnet, pulse repetition rate measurements, pulse duration measurements, and the measurement of the modulator pulse.

11-671. MAGNETRON MAGNET FIELD STRENGTH MEASUREMENT.

11-672. GENERAL. The magnetron is the high-power transmitting tube almost universally used in modern microwave radar. As a source of high-power microwaves, the multicavity magnetron represents a very great advance over both conventional space-charge and velocity-modulated or medium power klystron-type tubes. Magnetrons produce pulse powers on the order of hundreds of kilowatts at frequencies as high as 24, 000 mc. Magnetrons are basically self-excited oscillators whose purpose is to convert dc input power into rf output power. Magnetrons are generally constructed so that they are inserted between the pole pieces of a permanent or electromagnet. With all the magnetron parameters fixed, except magnetic field strength, the output power is a direct function of the magnetic field. Because the efficiency of operation of the magnetron is directly affected by the magnet, it is often advisable to measure the field strength of the magnet. The fluxmeter is a test instrument designed for this purpose.

11-673. TEST EQUIPMENT. The usual magnetic flux measuring set is designed to permit the ready measurement of the magnetic field strength existing between the pole faces of the magnet and covers a range of about 1000 to 10, 000 gausses. The equipment incorporates two meters, one of which is used as a probe and is inserted between the pole pieces. Figure 11-93 shows the schematic diagram of a fluxmeter in current use.

11-674. The fluxmeter consists basically of a battery and two milliammeters connected in a series circuit. The milliammeter calibrated in gausses derives its field power from a self-contained permanent magnet. The second milliammeter, used as a probe, secures its field power from the magnet under test. The probe unit consists of a double-pivoted, moving coil assembly having a pointer fastened to the movement. The coil assembly rotates the pointer between black and red marks located respectively at either end of the scale. At the red mark, the moving coil system is lined up perpen-

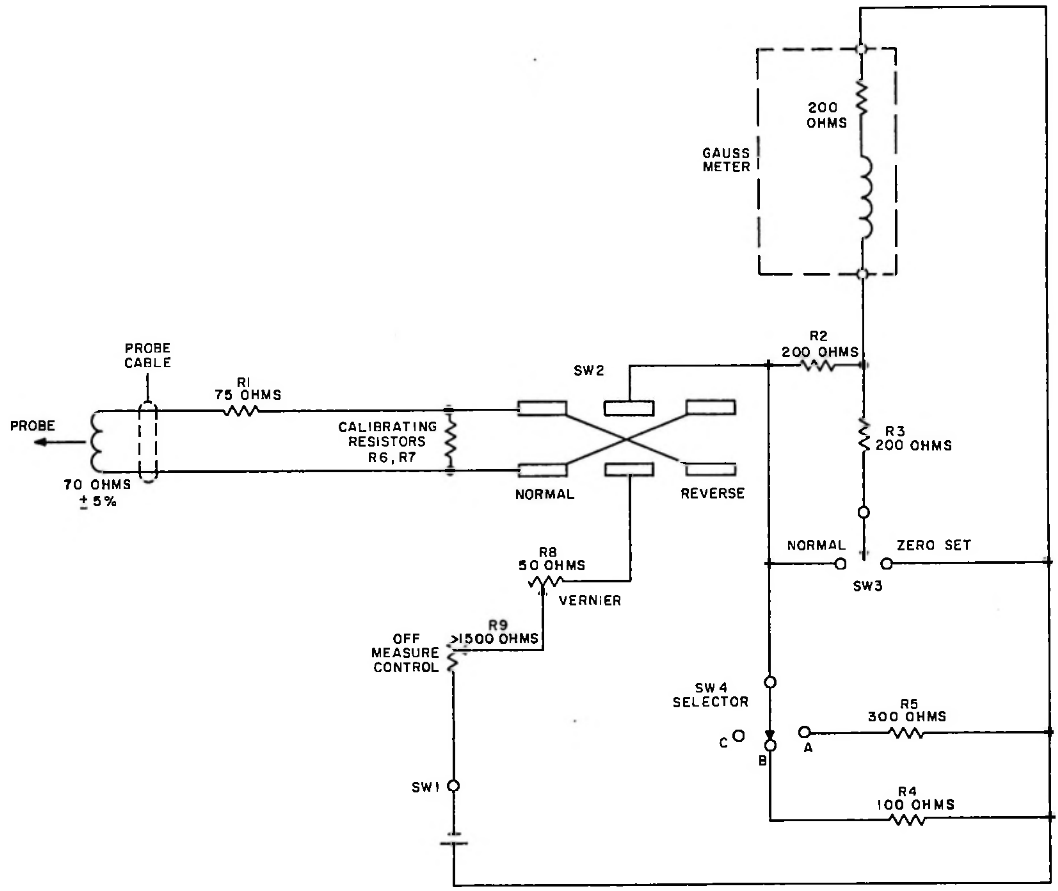


Figure 11-93. Magnetic Flux Measuring Set

dicular to the pole faces of the magnet and is the full-scale indication. The deflection of the probe is a product of the direct current through it, and the strength of the magnetic field or flux in which it is immersed. With a known field strength, the current required to set the probe pointer to the red mark can be measured. Conversely, the current required to set the probe pointer to the red mark is a measure of the field strength of the magnet in gausses. That is to say, the magnetic field is proportional to the amount of current necessary to rotate the pointer to the red mark (perpendicular to the pole faces). The gauss meter is provided with several ranges.

11-675. PROCEDURE. The measuring procedure using a typical fluxmeter is described in the following steps:

a. Remove probe unit of the fluxmeter from its recess, and place it in a level position, with probe-meter face horizontal on edge of table, or any flat surface which will clear cable.

b. Zero-set probe pointer by turning adjusting screw on probe until the pointer is aligned with the BLACK mark. (The pointer and its image in the mirror scale should coincide with the BLACK mark when the pointer is directly above it.)

CAUTION

Do not adjust probe zero set when probe is in magnet gap, or immediate vicinity.

c. Set A/B/C RANGE SELECTOR switch on "C". Adjust zero-set on gauss meter, by rotating OFF-MEASURE control, until gauss meter pointer is lined up with the full-scale mark. Hold NORMAL-ZERO SET switch to ZERO SET position. This should line up gauss meter pointer with the mark at the low end on scale. If pointer is not lined up, turn zero adjusting screw on face of gauss meter to displace pointer an equal amount on the opposite side of the low scale mark.

d. Recheck step c and repeat adjustment of zero adjusting screw if necessary.

e. With OFF-MEASURE control knob in OFF position, place probe unit between pole faces of magnet to be measured, making certain that the probe is well seated, and centered. (The face of the probe must be horizontal.)

f. Unless the approximate strength of the magnet is known, set range selector switch to the high scale. If the approximate strength of the magnet is known, set selector switch to appropriate range.

g. With the DIRECT-REVERSE switch in DIRECT position, turn OFF-MEASURE control knob clockwise from OFF position. If the probe pointer moves backward, turn OFF-MEASURE control knob to OFF position. Set DIRECT-REVERSE switch to REVERSE position and proceed.

h. Rotate OFF-MEASURE control knob until probe pointer is aligned with RED mark, or the gauss meter pointer reaches full scale. If the latter occurs, return OFF-MEASURE control knob to OFF position, set A/B/C SELECTOR switch to next lower range and repeat.

i. When probe pointer is set to RED mark (use vernier control to position pointer exactly), the gauss meter reading on the scale corresponding to the A/B/C SELECTOR switch setting gives magnetic flux directly in hundreds of gausses.

11-676. PRECAUTIONS. Certain precautions should always be taken when handling magnetron magnets. The field strength of these magnets is greatly reduced if they are jarred or hit even lightly. The magnetic field is very strong and if magnetic tools are used when working close to the magnet, the strong field may pull them sharply against the magnet.

11-677. In one test which was made to determine the effect of allowing tools to strike a magnet, it was found that only one touch with a screwdriver reduced the main magnetic field by 50 gausses. Since the magnetron used with this particular magnet was designed to operate properly within limits of 50 gausses above or below its rated 2500 gausses, two or three light taps on the magnet would seriously affect its performance. This difficulty can be avoided by the use of nonmagnetic tools. A nonmagnetic screwdriver is essential. If magnetic tools must be used, special precautions must be taken to prevent them from jumping toward the magnet. A nonmagnetic cover such as cloth or tape, wrapped around the pole pieces to a depth of three sixteenths inch, reduces the effect of touching a screwdriver to the magnet to about one-tenth of what it would be without the cover. In general, iron, nickel, and other magnetic objects should not be brought near the magnet. When drilling or filing in the vicinity of the magnetron assembly, cover it completely so as to prevent any metal filings from becoming attached to the magnet. During storage, care should be taken to prevent the interaction of the fields of two or more magnets. A safe rule is to allow not less than six inches between them. In addition, always store a magnet with the keeper in position.

11-678. PULSE REPETITION RATE MEASUREMENTS.

11-679. GENERAL. Pulse repetition rate (prf), also termed pulse recurrence frequency, is an important factor in radar performance. The pulse rate is a direct function of average transmitter power (paragraph 11-538) and also serves to set the upper limit or maximum range of a radar equipment. Short range radars (eg fire-control or ground controlled approach radars) have high prf's while long-range (search) radars have much slower rates, since the required rest time is considerably longer for the greater ranges. Pulse recurrence frequencies of 300 to 400 pulses per second are typical for radar equipments operating at 250 or 300 mile ranges. Fire control and ground approach radars have pulse repetition rates that are as high as several thousand pulses per second. Regardless of the type of pulse radar, the pulse repetition frequency is in the low audio frequency range. There are two general methods for measuring the rate. The first method employs an audio generator and an oscilloscope; the second, only a high-speed oscilloscope or synchroscope which incorporates accurate timing-mark calibration of the sweep traces.

11-680. AUDIO OSCILLATOR AND OSCILLOSCOPE METHOD. Figure 11-94 shows a typical equipment arrangement. Procedures

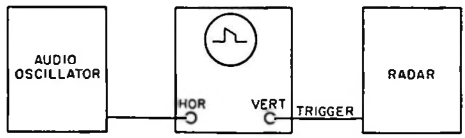


Figure 11-94. PRF Measurement, Audio Oscillator and Oscilloscope Method

for performing the measurement are as follows:

a. Connect audio oscillator, adjusted to expected prf, to the HORIZONTAL input of the oscilloscope.

b. Connect the test trigger output (or a sample of the transmitted pulse) from the radar to the VERTICAL input of the oscilloscope.

c. Adjust output frequency of the audio oscillator until a single stationary trigger pulse appears on the oscilloscope. The audio oscillator frequency reading is the pulse repetition frequency.

11-681. OSCILLOSCOPE METHOD. The oscilloscope method employs a high-speed oscilloscope or synchroscope having an accurate internal means of sweep time calibration. Since one complete transmitting time cycle equals the sum of the pulse width and the rest time, the number of cycles in one second equals the pulse recurrence frequency or

$$PRF = \frac{1}{\text{Pulse width + rest time}}$$

Procedures for determining the pulse recurrence frequency are as follows:

a. Connect the test trigger or a sample of the transmitter output to the vertical input of the oscilloscope.

b. Adjust the sweep frequency until at least two pulses appear on the time trace.

c. Using the sweep calibration, determine the time elapsed between either the leading or trailing edges of two successive pulses.

d. Using the above formula, calculate the pulse repetition rate.

11-682. PULSE DURATION MEASUREMENT.

11-683. The following is a typical procedure for measuring pulse durations:

a. Connect the pulse from the radar to the vertical input of a synchroscope.

b. Set the SYNC selector to SIGNAL.

c. Turn MARKER INTERVAL switch to the interval which will give the most convenient time scale for the measurement to be made, making certain that a sufficient number of markers appear for accurate measurement.

d. Carry out the measurement by directly comparing the signal and markers, counting the interpolation made necessary when the limits of the signal image do not coincide with the markers. The pulse should be measured at its 50 percent amplitude points. The reference scale of the synchroscope is a convenient aid in interpolating, especially when the sweep time and horizontal position controls are adjusted so that the markers of interest are made to coincide with the horizontal scale marks, a convenient number of divisions apart.

11-684. Alternatively, pulse duration can be measured by calibrating the reference scale with markers and then by determining the duration by the scale. For a series of rough measurements, this indirect method may be convenient, but care must be exercised not to move the SWEEP TIME controls after the calibration is made.

11-685. MODULATOR PULSE MEASUREMENT.

11-686. GENERAL. Since modulator pulses in the order of thousands of volts are common in radar equipment, several different pieces of test apparatus have been developed so that these high-voltage pulses may be viewed on an oscilloscope. These pieces of apparatus all perform the basic function of dividing the modulator pulse so that a small portion of the pulse voltage, which is proportional to both the magnitude and shape of the pulse, can be applied to the oscilloscope at a voltage low enough to be within the operational range of the oscilloscope.

11-687. RESISTIVE LOAD. The resistive load is a form of dummy load capable of providing a termination for making over-all performance tests on radar modulators. The termination is comprised of high-voltage resistors in the form of a voltage divider of known ratio for the purpose of measuring and viewing the output pulse of the modulator with an oscilloscope.

11-688. A dummy load typical of a 50-ohm termination consists essentially of two resistive elements made up of one 49-ohm resistive element and a 1-ohm element connected in series, thus providing a 50-to-1 ratio voltage divider. A tap or coaxial connector, with a nomenclature such as MEASURE PULSE, is connected across the 1-ohm resistor. The electrical connection from the divider to the modulator is made with the high-voltage pulse cable of the radar set. The modulator is operated in accordance with the standard instructions of the radar set. Before any measurements are made, you must operate the modulator for a period long enough to heat the load thoroughly (often 10 minutes). THIS IS IMPORTANT since the values of the resistances are generally unstable, decreasing in temperature. The total resistance and divider ratio is determined with the resistors thoroughly heated. To measure the modulator pulse, connect an oscilloscope, capable of displaying video pulses, to the MEASURE PULSE terminal video jack. Observe the shape and measure the pulse voltage. The voltage is determined by means of the oscilloscope and a calibrating voltage. Measure the pulse by taking the voltage indicated by the oscilloscope and multiply it by the ratio factor of the resistance load (50 in the case illustrated).

11-689. VOLTAGE DIVIDER. The voltage divider is a complete portable test equipment used to measure, in conjunction with an oscilloscope, video pulses from about 200 volts to 20, 000 volts (or higher in high-impedance circuits). The voltage divider differs from the resistive load basically in that the pulse voltage division is accomplished by capacitors rather than resistors.

11-690. OVER-ALL SYSTEM TESTING.

11-691. It was pointed out in paragraph 11-480 that judging the range capability and data accuracy of a radar facility by visual observation alone has been found to be inaccurate and valueless. Numerous field tests made on radar sets have verified this statement. In fact, these tests disclosed that many field radars were performing at no more than one-half their maximum capability, although the maintenance and operating personnel considered the operation to be normal. Since any performance less than optimum reduces the tactical area protected by these radars, you can readily see that the measurement of performance is of the utmost importance, especially in time of war. Investigation of the above situation disclosed that many technicians are not completely familiar with the latest techniques of radar system testing. In the following text, the over-all system techniques of radar testing which were not previously covered are discussed. The topics included are: (1) Timing-Circuit Calibration; (2) Standing-Wave-Ratio Measurement; (3) Spectrum Analysis; and (4) Over-all System Performance.

11-692. TIMING-CIRCUIT TESTING.

11-693. GENERAL. Although pulsed radar facilities were developed primarily for detection and range-finding of various objects, certain specialized requirements have come into later applications. Each radar facility is designed for a particular job which requires the use of all the information and data supplied by the facility. Radar facilities are classified as to the particular use that is made of the data supplied. Any radar facility performs one or more of the following functions: (1) Search—location of targets with respect to the position of the radar (this may include iff applications); (2) Navigation—location of the radar with respect to targets or beacon stations; (3) Ground Control—control of aircraft and direction of air traffic (blind landings and fighter direction); (4) Fire Control—aiming of guns controlled by radar information; (5) Interception—directing fighters toward enemy and, if necessary, enabling blind firing; (6) Bombing—providing information that enables bombers to locate and destroy targets under blind conditions. Of the six functions just mentioned, all except search require a high degree of range accuracy. Therefore, in all radars except those used for search, it is most important for the timing circuits to be accurately calibrated.

11-694. RANGE MEASUREMENT. A block diagram of a typical radar ranging facility is shown in figure 11-95. The facility trigger generator supplies the pulse required to drive the modulator and to initiate the action of the range-marker generator. This generator produces an output pulse at some time after the action of the facility trigger. The

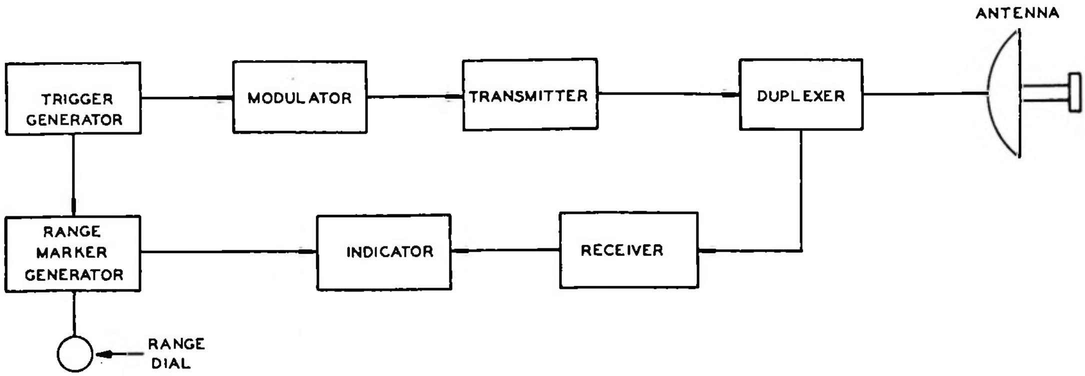


Figure 11-95. Typical Radar Ranging Method

exact time difference is dependent upon the time delay introduced. Range is determined by setting the range control to a point where the range pulse coincides with the target whose range is to be measured. The range in miles or yards is then read from the dial. In those radar facilities used for fire control or bombing, the dial may be mechanically connected to a computer.

11-695. RANGE DATA ACCURACY. Since the range of a target is determined by the time interval between the transmitted pulse and the echo return, all time delays introduced by the radar-facility components will add to the range indication, so that the range indicated will be in error by an amount equivalent to the time delays which have been introduced.

11-696. In every radar facility, there are time delays which occur within the equipment between the time the facility trigger is initiated and the time the echo pulse arrives at the indicator. The causes of these time delays are as follows: (1) the modulator output pulse occurs a short time after the trigger input pulse; (2) the rf output from the transmitter takes time to increase to the proper amplitude after application of the modulator pulse; (3) time is required for the transmitter rf energy to travel to the antenna, and for the reflected rf energy to travel from the antenna to the receiver; (4) time is required for the rf pulse to travel through the receiver; (5) time measurement errors are caused by incorrect scale factors in the ranging circuits, eg the oscillator frequency which determines the range mark position may not be correct, or the linear sawtooth slope which generates timing delay may slope too much or too little. Therefore, this is called a slope or rate error; (6) nonlinearity of the ranging circuits is caused by nonlinearity of the "linear" sawtooth waveform; (7) there is also a small delay in starting the range circuit; however, this is a negative or compensating delay as it subtracts from the other delays.

11-697. The delays listed above in steps (1) through (6), when combined, may represent a range of 150 to 350 yards. Therefore, a target at zero range can be erroneously indicated at a range of 150 to 350 yards. This zero error seriously affects the accuracy of all range measurements. For example, a target at a distance of 1000 yards may be indicated at a range of 1150 to 1350 yards. Since the zero error represents a fixed quantity for any radar, it remains the same for all ranges of that set.

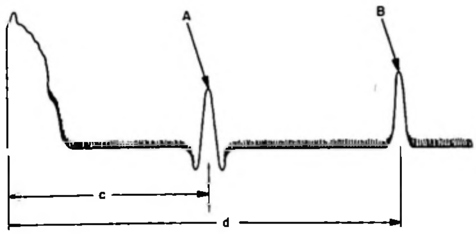


Figure 11-96. Double-Echo Range Scope Presentation

11-698. CALIBRATION. The sweep rate of the range unit must be adjusted to provide the correct distance between known range marks. You can use an external range calibrator or the fixed calibration unit built into the radar set to accomplish this adjustment. If the fixed calibration marks are at 2000 yard intervals, then the range unit should be adjusted to read 2080, 4080, 6090, 8070 yards, etc. A slight variation can be caused by nonlinearity of the ranging unit. This variation should be recorded because any large variation indicates range unit troubles. This calibration does not fix the zero error unless a previous calibration has provided the correct distance to the first range marker. Setting the range marker at the beginning of the sweep will not provide accurate results for zero error correction because of the nonlinearity at the beginning of the sweep and trace brightening, and because of other delays not considered, such as rf propagation time.

11-699. ZERO ERROR DETERMINATION. Before the effect of zero error can be corrected, it is necessary to determine how much time delay is present in a given radar set. There are four methods which are used in making this measurement. These are the fixed-target, the double-echo, the external range-calibrator, and the synchroscope methods. In some radars this determination is not necessary because a reference pulse precedes the transmitted pulse. Compensation for these time delays is therefore provided.

11-700. FIXED-TARGET METHOD. The fixed-target method of zero-error determination is the most reliable method in common use. This method involves the use of a fixed target at some accurately known distance. A natural target may be used, but a portable reflector gives more reliable results. The target range indicated by the radar facility is carefully read and compared with the known range. The range indicated by the radar should be greater than the known distance; the difference is the zero error.

11-701. DOUBLE-ECHO METHOD. The double-echo method of range correction is normally associated with fire-control radar equipments, but any radar set that is capable of receiving two echoes from the same target during one sweep on the range scope is capable of determining range accuracy by the double-echo method. An example of how the double-echo method is used is given below:

a. The radar antenna is trained on a target at a range of between 1000 and 3000 yards. A transmitted radar pulse strikes the target, is reflected, and returns to the point of transmission. A small portion of this returned radar pulse is accepted by the radar antenna and presented as target "A," (figure 11-96) on the indicator of the radar equipment. The remainder of the returned pulse is re-radiated and directed back toward the target for the second time. There the pulse is again re-radiated back to the point where the radar antenna accepts the pulse and presents it as target "B."

b. The time elapsed between the reception of target pulse "A" and target pulse "B" indicates the true range of the target from the radar set, provided, of course, that both pulses were received during the time of one sweep of the range scope. By placing the ranging device (step, notch, or marker) at target "B" and noting range "d," and then moving the ranging device to target "A" and noting range "c," the correct range may be computed. Subtract the range recorded for target "A" from the range recorded for target "B" ("d" minus "c"). This resultant figure should equal the range recorded for target "A." If it does not, then with the ranging device set at target "A," the range correction control is adjusted, or the range counter mechanically disengaged and "slipped" until the counters indicate the correct range. This procedure of noting the ranges and altering the counters is repeated as many times as necessary to obtain accurate ranging. One note of caution is observed—the measurement of ranges "c" and "d" is made as quickly as practicable, to prevent the possibility of the target range varying by any appreciable amount during the time interval between two range measurements.

11-702. EXTERNAL RANGE-CALIBRATION METHOD. Various test equipments are available for calibrating radar range units. These test equipments function basically in the same manner. They provide crystal-controlled calibration markers for checking and adjusting the calibration circuits of the unit under test. The range calibrator also provides a synchronizing pulse input to the calibration circuit under test. The timing of this triggering pulse is varied in order to synchronize the occurrence of markers on the range indicator with the marker generated by the range calibrator. The manner in which a typical range calibrator is employed is as follows:

a. Connect the 50-mile marker output of the range calibrator to the input of the first marker amplifier of the radar range indicator. Connect the trigger output of the range calibrator to the range indicator trigger input jack or terminals. Energize the radar indicator, and adjust the horizontal and vertical centering controls for proper positioning of the range sweep. Set the range indicator to operate on the 200-mile scale.

b. Adjust the range calibrator trigger delay control until the first marker aligns with the start of the range sweep. If it does not, then align the circuits using the adjustment controls provided, such as sweep rate or gate length controls.

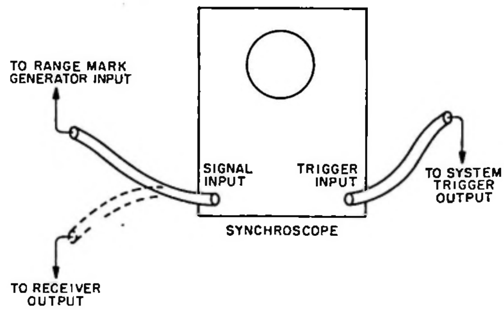


Figure 11-97. Test Setup for Synchroscope Method of Zero-Error Determination

11-703. SYNCHROSCOPE METHOD. The synchroscope method of measuring zero error does not require the use of a fixed target, and gives fairly accurate results. Figure 11-97 shows the test arrangement for this method. The synchroscope sweep is triggered by the radar facility trigger. A fast sweep (about 2 microseconds per inch) is used, and is carefully calibrated to the number of microseconds each inch of sweep represents. If the sweep is linear, any portion of the trace may be used. To determine zero error by this method, proceed as follows:

a. Use radar-facility trigger pulse to provide external sync for the synchroscope.

b. Calibrate sweep speed of the scope to represent about 2 microseconds per inch.

c. Feed the trigger pulse which starts the range-marker circuit into the vertical amplifier, and set the scope gain to provide a half-inch pulse.

d. Carefully mark the leading edge of the pulse on the scope.

e. Remove the trigger pulse and feed the radar-receiver output to the vertical amplifier. (Note that the radar local oscillator should be detuned; the transmitter pulse shock-excites the rf preamplifier sufficiently to produce an i-f signal.)

f. Adjust the scope gain to provide a one half-inch pulse when the receiver gain is set to produce about one eighth inch of noise.

g. Carefully mark the leading edge of the pulse on the scope, in the same manner as in step d.

h. Measure the distance between the two marks on the scope. Note that the pulse in step g occurs later than the pulse in step d.

i. Convert the distance between pulses to microseconds and, then, if you desire, into yards. This figure is the zero error.

NOTE

The above method of measuring zero error involves the use of a synchroscope having a fixed, internal time delay. Note that this delay is present in the readings taken in both steps d and g; therefore, both pulses are delayed by the same amount, thus eliminating the effect of the test equipment's internal delay.

11-704. ZERO-ERROR COMPENSATION. After the zero error of a radar facility is measured, compensation is made in the range-marker circuit. In most cases, calibration is carried out at two different points in the delay range. If a given radar has a 200-yard zero error and a 12, 000-yard range-marker circuit, and the calibration points are at 1000 yards and 10, 000 yards, the 1000-yd point is set at 800 yards and the 10, 000-yd point is set at 9800 yards. After compensation, the zero error is measured again, to make sure that it has been reduced to zero yards. The zero-error figure for a given radar facility is a fixed characteristic. To facilitate compensation during range-marker calibration, it is common practice to label the equipment with the last measured value of zero error.

11-705. STANDING-WAVE-RATIO MEASUREMENT.

11-706. GENERAL. In a radar facility a low swr is maintained principally for the following reasons: (1) reflections occurring in the rf line cause magnetron pulling, and can result in faulty pulsing (this effect is more pronounced when the line is long, as compared to a wavelength); (2) arc-over may occur in the rf line at maximum voltage points; (3) mechanical breakdown in the line may sometimes occur, because of the development of hot spots. To prevent magnetron pulling, the swr should be less than 1.5 to 1, which represents a reflection of less than 5 percent of the incident power. In the maintenance of radar facilities, swr measurements are useful in two ways. First, defective rf line components can be located by checking the swr of each component or by substitution. Second, radar facilities having rf tuning adjustments can be adjusted with the aid of swr test equipment.

11-707. SLOTTED-LINE METHOD. The slotted-line method of measuring swr can be used with the aid of an rf probe and a slotted line. The slotted line is a coaxial or waveguide section of transmission line, with a longitudinal slot cut into its outer conductor, which permits insertion of the rf probe. This is shown in figure 8-130. The slot is constructed at least a wavelength long, and is not wide enough to cause appreciable loss by radiation. In order to explore the voltage field existing in the line, the probe is placed in the electrostatic field through the slot and moved back and forth. The probe feeds an rf detector, and the rectified output operates a meter which indicates the swr.

11-708. Many radar facilities have slotted-line test sections that are integral parts of the rf section. In such cases, a removable protective plate usually covers the slot, and allows the line to be pressurized. For those facilities not having built-in slotted sections, a series of slotted-line sections to fit any radar set have been devised. In some cases the section is inserted into the radar set by means of coupling sections, or a test setup is developed whereby the radar units may be tested. Figure 8-132 shows a typical built-in slotted-line section.

11-709. The construction of a typical rf probe is shown in figure 8-138 along with the associated test circuit for swr measurement. The probe wire is adjustable for depth of penetration (coupling), so that the amount of rf pickup can be controlled. In practice, the coupling is maintained at a minimum, to prevent distortion of the fields inside the rf line. The rf detector shown in figure 8-138 is a type of bolometer called a barretter, and consists of a fine resistance wire, which presents a matched load to the rf probe. Refer to paragraph 8-282 for a discussion concerning the barretter. A direct current is passed through the resistance wire, and the current value is altered by the rf energy, which heats the wire, thus changing its resistance. Rf pulses cause corresponding pulses of current flow, which are amplified, rectified, and fed to an indicating meter calibrated to read swr directly. If a crystal rectifier, such as a 1N21 or 1N23, is used in place of a barretter, no direct current is necessary.

11-710. Swr is measured by performing the steps given below.

a. If the radar facility has no built-in slotted line, insert a slotted section into the radar transmission line (with the adapters provided), as close to the magnetron as possible.

b. Connect rf probe to amplifier, and adjust probe for a penetration of a few thousandths of an inch.

c. Operate radar transmitter and move probe to a maximum point.

d. Set probe penetration to provide a scale meter reading of 1.

e. Move probe to a minimum point, and read swr on meter.

f. If swr is too high (1.5 to 1 or higher) and the radar facility has tuning stubs, adjust the stubs for an swr of as near 1:1 as possible. The latter adjustment is made with the antenna pointing at free space.

11-711. DIRECTIONAL-COUPLER METHOD. The directional-coupler method of determing swr is frequently used in the field. To determine the swr of an rf assembly, the coupler is inserted into the transmission line and the incident power is measured. The coupler is then reversed, and the reflected power is measured. The swr can then be calculated. This method is not very accurate if the coupler directivity is low. For example, if a given coupler has a nominal loss of 20 db and a directivity of 20 db, the swr obtained as a result of its use is 1.6 to 1; however, the actual swr may be any value from 1.4:1 to 1.8:1. It is seen, then, that the greater the directivity, the greater the accuracy.

11-712. The above method has been largely superseded by the use of the bidirectional coupler, which greatly simplifies the application. This coupler, which consists of two directional couplers mounted on the same line and coupling in opposite directions, has been developed to a point where the accuracy of the method is quite good. The bidirectional coupler does not require reversing, is easier to use than the slotted line, is more rapid in operation, and can be made a permanent part of a pressurized rf assembly. In addition, the coupler can be used in connection with power and frequency testing and spectrum analysis, which is discussed in detail following standing-wave-ratio measurement.

11-713. BIDIRECTIONAL COUPLER. In many modern equipments, a bidirectional coupler is included in the transmission line. The operation of this coupler for use in waveguides is discussed in paragraph 11-557, and is shown in figure 11-63. Only the method for using the coupler for measurement of swr is given here. Bidirectional couplers are also designed for use in coaxial transmission lines.

11-714. The value of swr is given by the ratio of the voltage at a maximum point to that at a minimum point; that is:

$$SWR = \frac{E_{max}}{E_{min}}$$

Since E_{max} is the sum of the incident voltage E_i and the reflected voltage E_r, and E_{min} is the difference of these two voltages, the equation above can be rewritten as:

$$SWR = \frac{E_i + E_r}{E_i - E_r}$$

Making use of the relationship $P = \frac{E^2}{Z}$, the equation becomes:

$$SWR = \frac{\sqrt{Z_L}\sqrt{P_i} + \sqrt{Z_T}\sqrt{P_r}}{\sqrt{Z_L}\sqrt{P_i} - \sqrt{Z_T}\sqrt{P_r}}$$

where:

P_i = incident power from the transmitter

P_r = reflected power from the transmission line termination

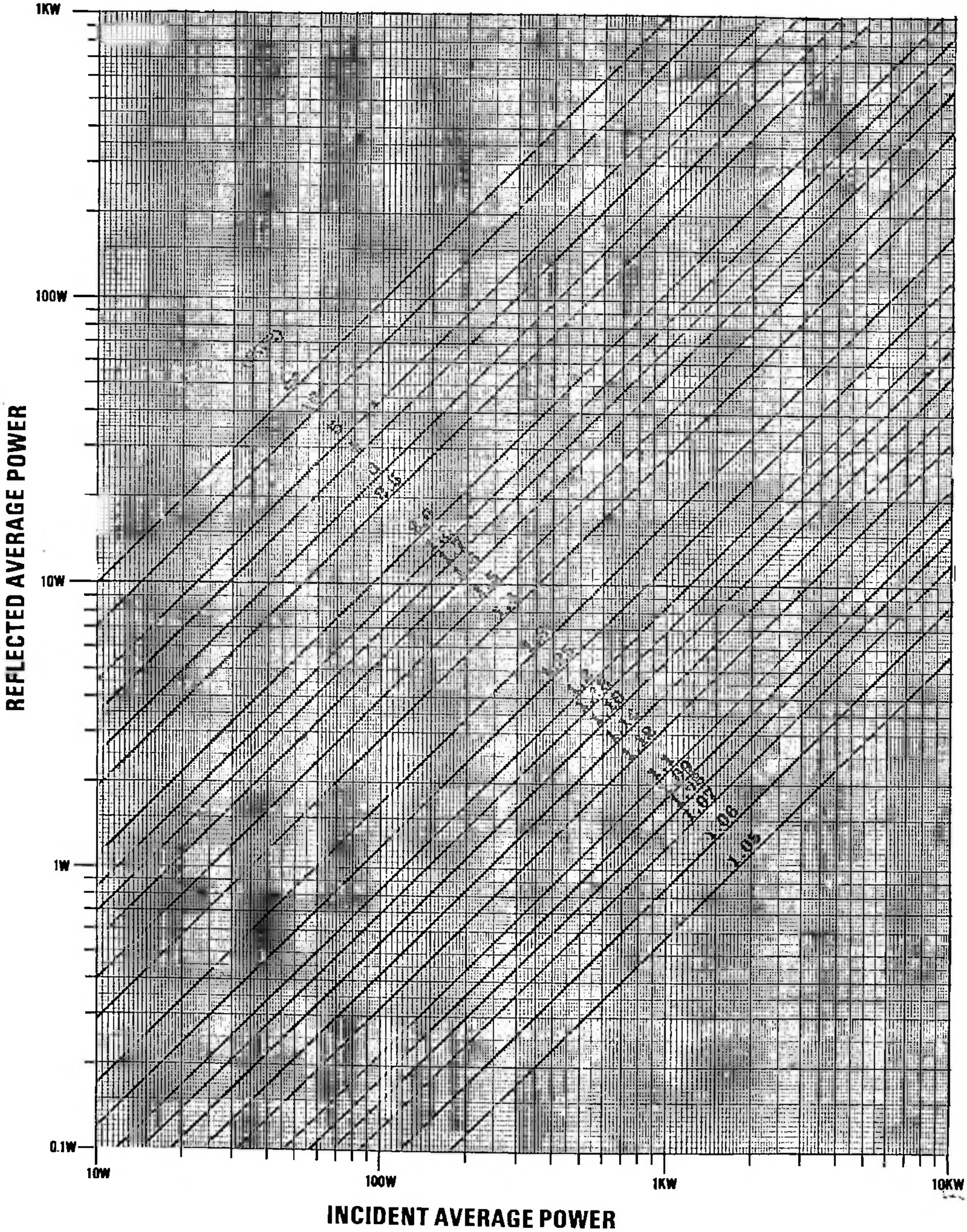
Z_L = the load impedance of the terminated transmission line as seen from the coupler

Z_T = the transmitter output impedance as seen from the coupler

In most cases, $Z_L = Z_T$ (the condition for maximum power transfer). For this condition, the equation is simplified to:

$$SWR = \frac{\sqrt{P_i} + \sqrt{P_r}}{\sqrt{P_i} - \sqrt{P_r}}$$

11-715. The chart in figure 11-98 is a plot of this equation for common values of incident and reflected power. An important item in using this method for finding swr is that the power meter used with the bidirectional coupler will not, in general, directly give actual values of incident and reflected power. The total attenuation in the coupling to the meter must be taken into account when calculating incident and reflected power. These values are then used to determine swr on the chart.



$$SWR = \frac{\sqrt{P_I} + \sqrt{P_R}}{\sqrt{P_I} - \sqrt{P_R}} \qquad P_R = \frac{(SWR - 1)^2}{(SWR + 1)^2}\, P_I$$

Figure 11-98. Graph of Relation Between Incident Power, Reflected Power, and SWR

11-716. MAGIC-TEE COUPLER METHOD. A special form of directional coupler called the "Magic-Tee," which is illustrated in figure 8-139 can also be used for swr measurements. The magic-tee coupler consists of four waveguides joined together as shown in the above named figure. The important consideration in using this type of coupler is that the polarization of arm D is at right angles to that of arm A, so that there is no direct coupling between arm A and arm D. However, note that arm A couples to both arm B and arm C, and that arm D also couples to both arm B and arm C. Therefore, power is coupled from arm A to arm D only when reflections are produced in arm B or arm C. A signal generator is connected to arm A through an isolating attenuator which provides an impedance match to this arm. A matched load, similar to that shown in figure 8-140, is connected to arm C. The device under test is connected to arm B. An rf detector is connected to arm D, and provides an impedance match to that arm. In operation, power enters arm A, and is divided between arm B and arm C. No power is coupled to arm C; the matched load absorbs all the incident power coupled in that direction. If the device under test causes no reflection, all the incident power flowing to arm B is absorbed. When this condition prevails, the detector in arm D indicates zero. However, if the device connected to arm B is not matched, some of the incident power is reflected. The amount of reflected power is a function of the degree of mismatch. The reflected power is returned to the junction of the arms, and a portion is coupled to arm D. Thus, any power in arm D indicates a mismatch at arm B, and the swr is evaluated in terms of the power in arm A, as compared to the power in arm D.

11-717. CAUSES OF STANDING WAVES. Any discontinuous change along an rf line, such as might be introduced by a change in dimensions, or a change in geometry introduced by a sharp bend or dent, or an obstacle in the line, produces reflections. Some of the most common causes of an excessively high standing-wave ratio are: (1) dirt or moisture in the rf line; (2) dented or bent line; (3) burrs or poorly soldered joints; (4) defective coupling joint; (5) defective rotating joint; and (6) mismatched antenna.

11-718. LOCATING DISCONTINUITIES. If an increase is noted in the standing-wave ratio, check the rf transmission lines for the common causes listed above, as well as for any other damage which may result from battle, storms, or normal wear. Check the antenna also, since any bending of the reflector or dipoles changes its impedance and results in an increased swr. Many rf transmission-line faults are visible and easily located; however, in some cases the trouble may be of such a nature that the defective part can be found only by making special tests.

11-719. DUMMY RF LOAD METHOD. A dummy rf load is a resistive section of transmission line which absorbs rf power without causing appreciable reflection. Figure 8-140 shows the construction of a typical waveguide-type dummy load. This particular dummy load uses a section of waveguide which is filled with a mixture of sand and squadag which serves as a resistance to absorb power. In order to minimize reflections, the front surface of the resistive element is constructed so as to present an oblique surface to the incident rf power. The exterior of the dummy load is fitted with cooling fins, and is painted a dull black for greater heat transfer. A typical X-band load gives an swr of less than 1.05 to 1, and absorbs 150 watts of power (average). If necessary, the power rating of dummy rf loads is increased by the use of forced air or water cooling. Coaxial dummy rf loads are similar in operation to the waveguide-type loads. The resistive mixture forms a tapered contact between the inner and outer conductors. Two distinct advantages are gained by use of dummy rf loads in radar maintenance: (1) When military security prohibits rf radiation, maintenance can still be carried out with the aid of the loads; and (2) the load can be used to absorb power without reflections, regardless of surroundings; it may also be used for measurement of power output.

11-720. When the swr of an rf transmission line becomes excessive, the cause of the standing waves may be located with the aid of a dummy rf load by performing the steps given below:

a. Remove antenna feed, and substitute dummy rf load. If the substitution corrects the condition, the trouble may possibly be in the antenna, or it may be the result of reflections from nearby objects.

b. If the swr is still too high, change the antenna scanning position and recheck the swr. If the swr changes, a defective rotating joint is sometimes indicated. In some cases, the swr may not change, even though a rotating joint is defective. Therefore, this test will not eliminate the possibility of a bad joint, but will locate faults caused by rotation.

c. If the rotating joints are not defective, remove the section of line next to the antenna feed section, and replace the dummy load at the open end of the rf assembly. Recheck the swr. If the section removed is defective, the swr will improve. Continue the process of removing sections until the offending section is located. The swr should be checked again after the trouble is corrected and the rf components are reassembled.

11-721. COMPONENT CHECKING. Test methods using the swr for checking rf transmission-line components are described in Section VI of Chapter 10.

11-722. SPECTRUM ANALYSIS.

11-723. GENERAL. It is possible, by means of a spectrum analyzer, to observe a selected portion of the electromagnetic spectrum on the screen of a cathode-ray tube. The display consists of vertical pulses distributed along the horizontal axis; the position of each pulse indicates the frequency of a particular signal, while the relative height of each pulse indicates the relative strength of the signal. In other words, the display viewed on the cathode-ray-tube screen is, in effect, a graph of energy plotted against frequency.

11-724. By analyzing the spectrum of a radar transmitter, a great deal of information is obtained regarding the condition of the radar. The spectrum analyzer can show the presence or absence of frequency modulation, and can also indicate the presence or absence of amplitude modulation in the signal. By means of a frequency meter, which is normally an integral part of the spectrum analyzer, it is possible to determine the bandwidth necessary to transmit each signal. The built-in frequency meter may also be used to check local-oscillator frequencies and retune receivers. If frequency pulling is present, the shift of the spectrum may be observed. To summarize, spectrum analysis provides information regarding the condition of a radar. Incorrect spectral displays can indicate magnetron pulling, magnetron pushing, magnetron double-moding, mode shifting, and mode jumping, and improper pulse width. In addition to the items mentioned above, indications of transmitter and local-oscillator frequencies are available which facilitate local-oscillator tune-up and afc testing.

11-725. SINE-WAVE SPECTRAL DISPLAY. For purposes of discussion, a pure sine wave can be considered to represent a single frequency. The spectral display of this waveform is shown graphically in part (A) of

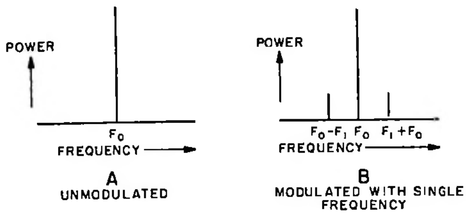


Figure 11-99. Sine-Wave Spectral Displays

figure 11-99 as a single vertical line designated as F_0. The height of the line represents the power contained in the single frequency. Part (B) of figure 11-99 shows the spectral display for a single sine-wave frequency (F_0) amplitude modulated by another sine wave (F_1). In the latter case, two sidebands are formed, one above and one below the carrier frequency represented by F_0, corresponding to the sum and difference frequencies. When additional modulating frequencies are used, two sidebands are added for each frequency.

11-726. SQUARE-WAVE ANALYSIS. A perfect square wave is considered to consist of a fundamental sine wave plus an infinite number of odd-harmonic, in-phase sine waves which progressively decrease in amplitude as the harmonic number increases. Theoretically, therefore, a 100-cps square wave contains frequencies of 100 cps, 300 cps, 500 cps, 700 cps, etc, to infinity. Since a perfect square wave represents an ideal condition, only a limited number of harmonics are involved in the usual square wave. However, a good square wave may contain frequencies up to the 100th harmonic, in actual practice.

11-727. TRANSMITTER SPECTRAL DISPLAY. When a transmitter is modulated by short rectangular pulses occurring at the prf (pulse repetition frequency) of the radar, two distinct modulating components are present. One component consists of the prf and its associated harmonics; the other consists of the fundamental and odd-harmonic frequencies that comprise the rectangular pulse, as was previously discussed. Figure 11-100 shows an ideal display of that part of the spectrum covered when an rf carrier is pulse-modulated. The vertical lines in the figure represent the modulation frequencies produced by the prf and its associated harmonics, while the lobes represent the modulation frequencies produced by the pulse frequency and its associated harmonics. The vertical lines are separated by a frequency equal to the prf. The amplitude of the main lobe falls off on either side of the carrier until it is zero at the points corresponding to the second harmonic of the fundamental pulse frequency. The first side lobe is produced by the third harmonic of the pulse frequency; the second zero point, by the fourth harmonic; and the second side lobe, by the fifth harmonic. In the ideal spectral display, each frequency above the carrier has a counterpart frequency equally spaced below the carrier, so that the pattern is symmetrical about the carrier. The amplitude of the side lobes is considered important, because in the ideal spectral display, the first side lobe represents 4.5 percent of the carrier amplitude, and the second side lobe represents 1.6 percent of the carrier amplitude. The main lobe, of course, carries the major portion of the transmitted energy.

11-728. TRANSMITTER OUTPUT VERSUS RECEIVER RESPONSE. The importance of the transmitter output characteristics as compared with the receiver response becomes readily apparent by inspection of figure 11-101, which shows an optimum receiver response curve superimposed upon an ideal pulse spectral display. The receiver bandwidth is broad enough to include all the energy between the first zero points. Note that the receiver also responds to the first side lobes, but at reduced level. Any rf energy that exists outside the limits of

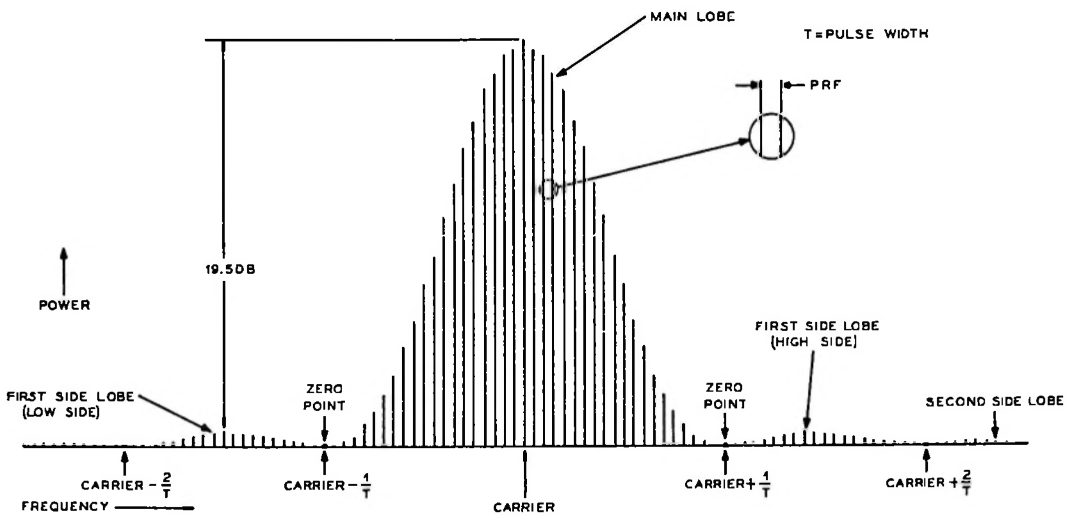


Figure 11-100. Ideal Spectral Display of a Pulse-Modulated RF Carrier

the receiver response is, of course, lost, and the effect is the same as if the transmitter power were reduced. Since practically all of the transmitted energy is within the limits of the receiver response, as shown in figure 11-101, further broadening of the receiver bandwidth results in very little increase in energy pickup. It is apparent, however, that a decrease in bandwidth causes a definite reduction in energy pickup. The spectrum side lobes contribute very little in terms of pulse amplitude, but they contribute to the steepness of the edges on the output pulses, as shown in figure 11-102. From this cursory examination, it can be seen that an ideal receiver has sufficient bandwidth to include a great many side lobes, in order to reproduce the transmitted pulse with a high degree of accuracy. However, if the above condition is obtained, the increased bandwidth allows the receiver to respond to more than the normal amount of noise and limits its sensitivity. A reduction of bandwidth, within limits, does not lessen the pulse amplitude, but reduces noise response. Too great a reduction of bandwidth

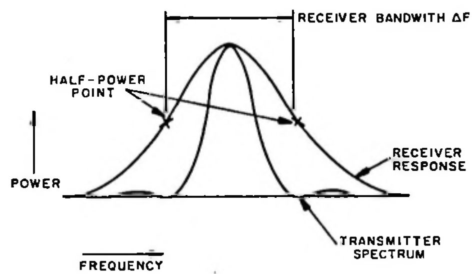


Figure 11-101. Transmitter Spectral Display Compared with Receiver Response Curve

results in decreased pulse amplitude, as shown in figure 11-102, because of the loss of some energy in the main lobe. Optimum bandwidth results in the greatest receiver sensitivity, but causes a slight distortion of pulse shape. Since accurate pulse shape is important in precision ranging and tracking operations, certain radar facilities designed for this type of service have a receiver bandwidth that is broader than optimum, to provide a sharp leading edge on the pulse.

11-729. MODULATION DISTORTION. In a

properly functioning radar transmitter, the period of the transmission interval is as specified, the oscillations during the transmission interval are of constant frequency and amplitude, and the time required for oscillation to start and stop is approximately zero. Any deviation from these conditions produces distortion which is visible on the spectral display in the form of either pulse amplitude distortion or frequency modulation, or both. Figure 11-103 shows the spectral display when fm is present. The zero points are lost, indicating the presence of new frequencies in the spectral display. This has the effect of placing more of the transmitted power in the sidebands, and therefore, results in the loss of energy outside the receiver pass band. Part (B) of the same figure shows the spectral display when the amount of frequency modulation is excessive. In part (B), the magnetron is operating at two distinct frequencies (double moding), and the receiver, if tuned throughout its range, would have two tuning points. When the above condition prevails, more than half of the transmitted power is wasted. The presence of pulse amplitude distortion in the transmitted output has the effect of producing dissymmetry in the display, as shown in figure 11-104. The zero points are still clearly defined, but the lobes on one side of the carrier are much larger than normal. In general, distortion resulting from frequency modulation is far more undesirable than distortion from amplitude modulation. Figure 11-105 shows a combination of both types of distortion, which results in a very poor spectral display.

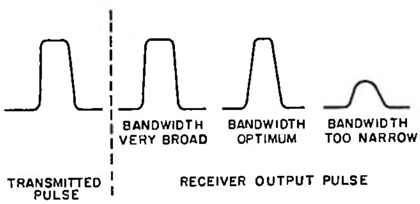


Figure 11-102. Effect of Receiver Bandwidth upon Pulse Shape

11-730. The troubles which give rise to a poor transmitter spectral display are sometimes difficult to locate. Methods used for trouble isolation are discussed under paragraph heading 11-773. Briefly, at this point it can be stated that trouble may arise from the following causes: (1) defective magnetron; (2) defective magnet; (3) mismatch in rf section (pulling); (4) improper pulse shape or amplitude (pushing); and (5) reflections from nearby objects (pulling).

11-731. METHODS. Two methods of obtaining a graph or display of the spectrum are described in the following paragraphs. The

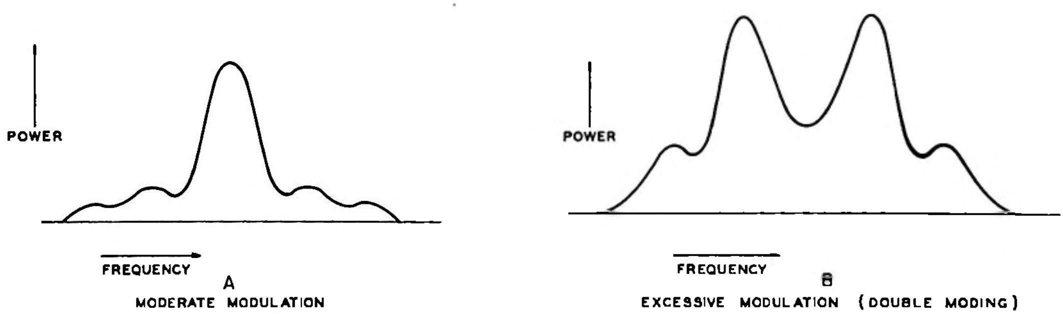


Figure 11-103. Transmitter Spectral Displays, Showing Distortion Resulting from Frequency Modulation

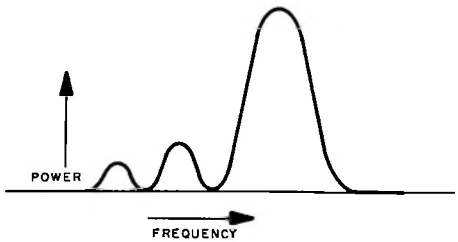


Figure 11-104. Transmitter Spectral Display, Showing Distortion Resulting from Amplitude Modulation

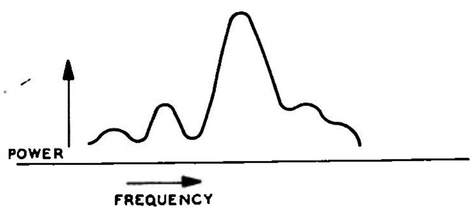


Figure 11-105. Transmitter Spectral Display, Showing Distortion Resulting from Combined Frequency and Amplitude Modulation

first method requires that a graph of power versus frequency be plotted. It is relatively slow, and demands considerable experience on the part of the technician. The second method is simplified by the use of a spectrum analyzer, which provides, as was mentioned earlier, a display on the screen of a cathode-ray tube corresponding to the graph plotted in the first method. The circuit of a typical spectrum analyzer is discussed briefly. Some spectral displays are examined and interpreted, and a method for frequency measurement is described.

11-732. FREQUENCY METER METHOD. The use of a frequency meter is a rather simple method of obtaining readings to plot a spectral graph. A high-Q, transmission-type, resonant-cavity meter, such as is found in most echo boxes is utilized, together with a rectifier-meter indicator. The test arrangement is shown in figure 11-50. Readings are taken at frequent intervals throughout the frequency range of the transmitter, and a graph is made to indicate meter readings vertically and frequency-meter indications horizontally. If the graph is very carefully plotted, a rough outline of the spectrum is obtained. After you have gained experience, you may get a good idea of the spectrum by merely noting how the meter reading varies as the frequency meter is tuned through resonance. All spectral readings must be obtained by rotating the frequency-meter dial in one direction only. If the dial is rocked into position, backlash in the dial drive mechanism will cause an appreciable error. The usual procedure is to approach each reading from the low-frequency side. A specific procedure for using the echo box is given under paragraph heading 11-745.

11-733. SPECTRUM ANALYZER METHOD. The spectrum analyzer, which is a form of panoramic receiver, provides a simplified method of analyzing spectral phenomena. A small portion of the transmitter output is coupled into the signal input circuit of the spectrum analyzer. Care must be taken to keep the input low enough to prevent burnout of the attenuator. A directional coupler provides an ideal coupling system but a pickup antenna may be used. Coupling methods are described under paragraph heading 11-548.

11-734. TEST EQUIPMENT CIRCUIT ANALYSIS. In the spectrum analyzer, a narrow-band receiver is electrically tuned through a range of frequencies, and the output, in terms of power, is displayed vertically upon an oscilloscope whose horizontal sweep is synchronized with the frequency sweep of the receiver. A block diagram of a typical spectrum analyzer is shown in figure 11-106. The receiver employed is a superheterodyne

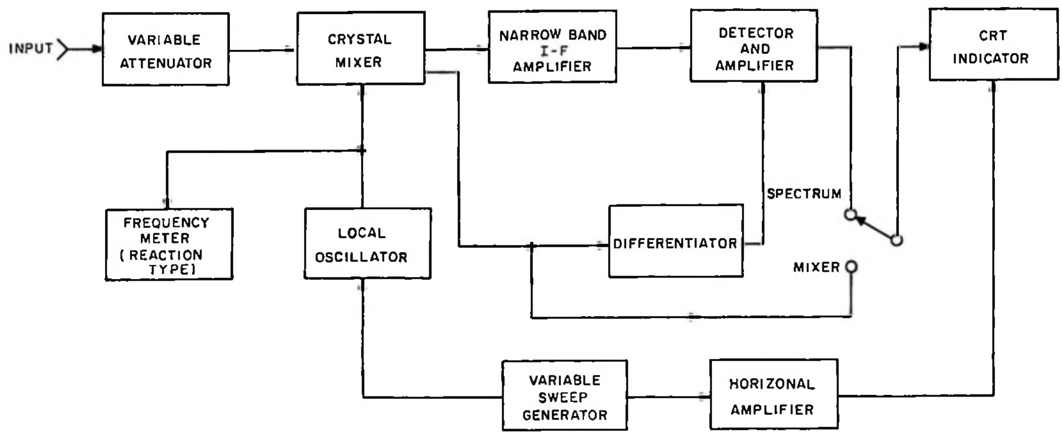


Figure 11-106. Typical Spectrum Analyzer, Block Diagram

type. The input, which usually consists of a coaxial-line termination, a broad-band attenuator, and a crystal mixer, is untuned, and therefore, responds equally well to all signals within the operating band. The local escillator is usually a reflex-klystron type. The i-f amplifier is a high-gain, narrow-band (50 kc or less) amplifier, usually operated above 20 mc. In some cases, double, or even triple, superheterodyne action is used to obtain the narrow bandwidth required. The i-f section is followed by a detector and amplifier which feed the vertical plates of a cathode-ray tube. The sweep generator produces a variable-frequency sawtooth voltage which sweeps the local oscillator repeller and, therefore, the receiver frequency and the horizontal deflection plates simultaneously. A reaction-type frequency meter is included which is designed to absorb local-oscillator power at resonance, thereby indicating the local-oscillator frequency.

11-735. On the front panel is a function switch usually labeled MIXER-SPECTRUM. In the SPECTRUM position, the indicator displays the output of the receiver. In the MIXER position, the indicator displays the crystal-mixer current, which is a function of the reflex-klystron local-oscillator output. Figure 11-107 shows a typical reflex-klystron chart. Note that the tube will oscillate only at certain voltages, and as the voltage is varied, the power output varies. Each separate voltage range of oscillation is called a mode. The modes are relatively flat on top, and each succeeding mode encountered becomes stronger as the repeller is made more negative. Within any given mode, the frequency is proportional to the negative voltage on the repeller. A frequency range of 60 mc is common in X-band tubes. However, the frequency at the top of each mode is determined by the size of the resonant cavity in the tube; therefore, all of the modes have the same center frequency.

11-736. The sweep generator produces a sawtooth voltage which is adjustable in both amplitude and average voltage value. The sawtooth amplitude control, usually called the spectrum width control, has sufficient range to cover at least one mode, and quite often, two. The average voltage control of the sawtooth, usually called the spectrum center control, allows you to choose any klystron mode you desire or to use any

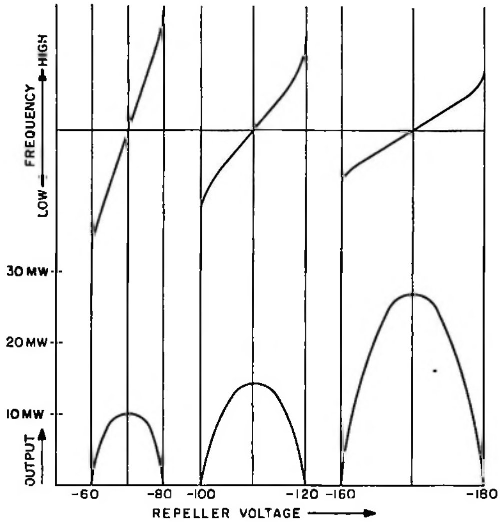


Figure 11-107. Typical Reflex Klystron Chart

range within a particular mode. In normal use, only a limited section of one mode is used.

11-737. The spectrum analyzer can be used as a klystron tube tester. When the function switch is in the MIXER position, the presentation is similar to that shown in figure 11-108, which illustrates one complete klystron mode and a part of another. The pip shown in the center of each mode is the frequency indication introduced by the reaction-type frequency meter. The mixer function of the analyzer allows the condition of the local oscillator to be checked; if you desire, the oscillator frequency can be set to any specified value. The klystron to be tested is substituted for the local oscillator in the spectrum analyzer, and the mode pattern observed. The amplitude of the mode indicates power relative to that of the regular oscillator. The tuning range is examined and any irregularities noted. Each mode should present a smooth regular curve. If desired, the tube under test is pretuned to the approximate frequency before insertion

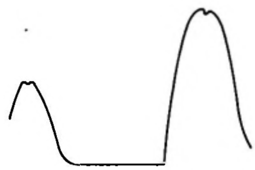


Figure 11-108. Klystron Modes as Presented on Spectrum Analyzer

into the radar, in order to simplify radar tune-up.

11-738. TRANSMITTER SPECTRAL DISPLAY ANALYSIS. As the spectrum analyzer frequency is swept, the spectral display appears upon the cathode-ray-tube indicator in the form of a series of vertical pulses. These pulses are not to be confused with the vertical lines shown in figure 11-100, which are separated by a frequency equal to the prf. If the spectrum analyzer has a bandwidth of 50 kc, a large number of prf lines are included in each pulse, because the analyzer samples a 50-kc segment of the spectrum each time the transmitter fires. Thus each pulse in the display represents the energy contained in a 50-kc band at the frequency of the analyzer at that instant. If the radar prf is 200 pulses per second and the analyzer sweep rate is 10 cps, the display consists of 20 pulses across the screen of the cathode-ray tube. Figure 11-109 shows a typical magnetron spectral display. These pulses indicate only the general outline of the display, and are much too coarse to reveal the internal structure. Figure 11-110 shows the same conditions with the spectrum width control advanced to produce a greater spread.

11-739. FREQUENCY MEASUREMENT. The measurement of frequency is greatly facilitated by the use of a differentiator. Refer to the block diagram shown in figure 11-106. A portion of the crystal-mixer current is applied to a differentiator, and the differentiated waveform is applied to the

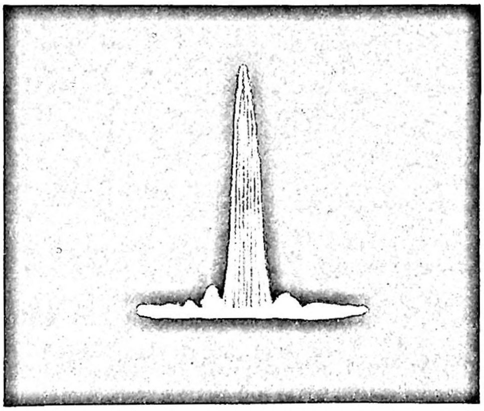


Figure 11-109. Typical Magnetron Spectral Display

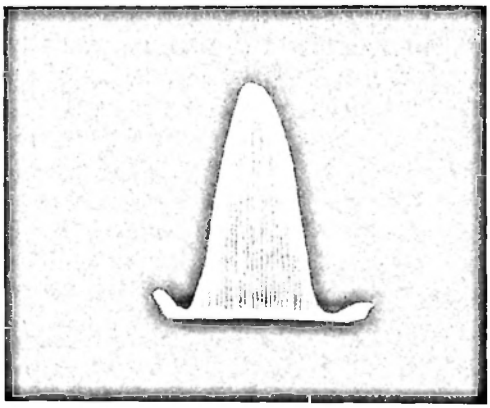


Figure 11-110. Typical Magnetron Spectral Display, Analyzer Width Control Advanced

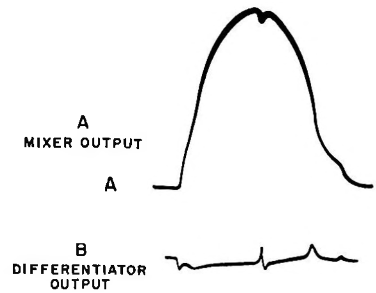


Figure 11-111. Effect of Differentiator upon Mixer Output

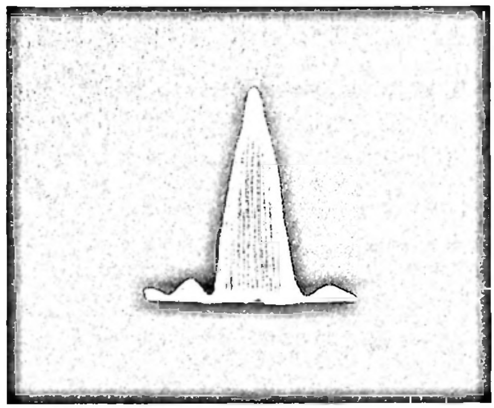


Figure 11-112. Typical Spectral Display, Showing Frequency-Meter Pip

amplifier section of the spectrum analyzer. Figure 11-111 shows the result of differentiating and amplifying the mixer signal. Part (A) of the figure shows the display with the function switch in the MIXER position, and part (B) is the display with the switch in the SPECTRUM position. Note that the frequency-meter pip now appears as an "S" curve, and that the mode ends are marked by pips. This signal is combined with the spectral display, and appears superimposed on the base line of the pattern, as shown in figure 11-112. The exact frequency is taken at the center of the "S" curve, where it crosses the base line. The pips marking the mode end limits should never be seen on the display, since no spectral indication may be obtained outside the mode limits.

11-740. FACILITY TESTING. The spectrum analyzer can also be used for some facility tests. In this case both the transmitter and local-oscillator signals can be conveniently sampled by means of a small pickup antenna placed near the base of the local-oscillator socket. In this position, the pickup antenna is in the rf leakage field, and the intensities of the two signals are approximately equal, because of the proximity of the pickup antenna to the weaker source. Because the rf section of the analyzer is untuned, image signals are also received. Thus, the signal picked up appears at two points on the analyzer tuning scale. In practice, however, an image is just as useful as the real frequency, and is often used in measurements even though the frequency scale is reversed.

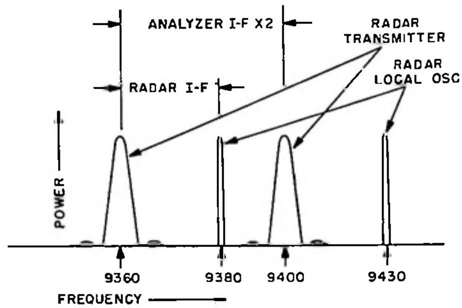


Figure 11-113. Over-all Spectral Representation of Transmitter and Local-Oscillator Output

11-741. Since the analyzer frequency meter is designed to indicate its local-oscillator frequency rather than the input-signal frequency, the most accurate frequency-measurement method is to measure the analyzer local oscillator when the oscillator is tuned above the input signal, and then measure it when the oscillator is tuned below the signal. The signal frequency is then halfway between the two readings. The frequency meter also is tuned for maximum absorption of the input signal, to obtain a direct indication of the input-signal frequency. Resonance is indicated by a slight reduction in the amplitude of the signal; however, this is difficult to observe.

11-742. Figure 11-113 shows an over-all spectral representation of a transmitter and local-oscillator of a particular radar facility. In this figure, it is assumed that the intermediate frequency of the spectrum analyzer is 25 mc, the radar transmitter frequency is 9375 mc, and the radar local-oscillator frequency is 9405 mc. This produces a radar intermediate frequency of 30 mc. If the spectrum analyzer were capable of showing the entire range of frequencies, the transmitter display would be recorded at 9350 mc and 9400 mc, and the local-oscillator display, which appears as a single frequency, would be recorded at 9380 mc and 9430 mc. (Note that the frequencies shown represent the local-oscillator frequency of the spectrum analyzer and not the signal frequency.) Most spectrum analyzers, however, cannot show the entire range of frequencies given in figure 11-113. A typical X-band analyzer, for example, is able to present a continuous range of only 50 to 60 mc. To examine various portions of the entire range, the analyzer tuning dial is turned, to vary the range of frequencies being covered. In this way, the entire band, 8500 mc to 9600 mc, may be covered by one instrument.

11-743. The spectral range can be made broad enough to display both the radar transmitter and local-oscillator frequencies simultaneously. Because it has this feature, the spectrum analyzer is recommended for use in tuning a radar local oscillator to a specified frequency. It is also recommended for use in checking afc action. The procedure for an afc check is as follows: Set the antenna scanning unit in motion, and note the pulling action on the magnetron. Any lateral motion in the display position indicates a change in frequency. With the afc in operation, any shift in the radar-transmitter frequency should be accompanied by a corresponding shift in the local-oscillator frequency. Therefore, the distance between the local oscillator and transmitter patterns should remain constant. Excessive pulling is usually evidenced by a distortion of the shape of the transmitter display.

11-744. A marginal check on afc operation to detect incipient trouble is made by turning on the afc and using a wavemeter to measure the normal oscillator frequency. Manually detune the receiver until the afc circuit fails to hold the oscillator frequency constant. The difference between the oscillator frequency when it just jumps out of tune and the normal frequency is an indication of how well the afc circuits are functioning. The larger the difference, the better the circuits are functioning.

11-745. OVER-ALL FACILITY PERFORMANCE.

11-746. GENERAL. Radar performance testing involves a series of measurements which are primarily intended to indicate the ability of the radar to detect targets. The combined results of the tests then indicate the over-all facility performance. Two distinct and separate factors are involved in the consideration of radar performance: (1) minimum-range performance; and (2) maximum-range performance. Both of these factors are discussed in detail in the text to follow.

11-747. MINIMUM-RANGE PERFORMANCE. There are certain radar facilities which are designed to detect and find the range of nearby targets. Examples of these are radar facilities used for fire control, aircraft interception, aircraft altitude indicators, and ground-controlled-approach applications. The tr recovery time and transmitter pulse width are important factors in determining the effective minimum range. If a target has a range of 200 yards, the echo is returned to the radar in about 1-1/4 microseconds after the occurrence of the transmitter pulse. If the receiver is to respond to this echo, the tr switch must recover sufficiently during this short interval to allow passage of energy to the receiver. Since a nearby target returns a strong signal, the recovery need not be complete, because the receiver responds to a strong signal, even at reduced sensitivity.

11-748. Minimum-range performance is also influenced by the transmitter pulse width. Long-range search radar facilities may use a pulse width of 2 microseconds or over, which represents a free-space range of over 320 yards. Radars designed for close-range work have pulse widths as short as one quarter microsecond, which represents only a little over 40 yards of free-space range. Furthermore, high-power radars may require the use of a pre-tr tube, to prevent transfer of harmonic energy through the tr; as a result, the recovery period may correspond to 2000 yards of range or more.

11-749. MAXIMUM PERFORMANCE RANGE. The factors which determine the maximum range of a radar facility are rather diverse; however, for the purpose of the discussion to follow, they may be divided into two general categories. The first category, which is not controllable by the maintenance activity, consists of target reflection and propagation factors. The second category is made up of facility performance factors, which are controllable to some degree by the maintenance activity.

11-750. Target reflection is a direct function of target size. Four principal factors enter into the determination of target sync or the amount of energy reflected by a radar target: (1) the material of which the target is constructed; (2) the surface area presented to the radar; (3) the configuration of the surface presented to the radar; and (4) the operating frequency of the radar. In general, it can be said that the amount of energy reflected from a target is a very complex consideration, and that the reflected-signal energy cannot be predicted, with any degree of accuracy. Therefore, when information on the reflected-signal strength is required, it is best found by direct measurement. For different target configurations, the reflected-signal strength varies considerably. Approximate target area for some aircraft is shown in table 11-4.

Table 11-4. Radar Target Area of Aircraft

OBJECT	TARGET AREA IN SQ METERS
Small aircraft	16
Large aircraft	74

11-751. PROPAGATION FACTORS. Atmospheric conditions play a very important part in radar performance. Some of the more common factors are: (1) duct formation; (2) temperature inversion and atmospheric refraction; (3) rain echoes and scattering; and (4) atmospheric absorption.

11-752. Duct formation occurs when there is a sharp discontinuity in the atmospheric conditions close to the ground. The discontinuity reflects a transmitted signal in about the same manner as a metallic surface, and thus directs the wave back to earth, where reflection occurs again. Therefore, the space between the earth and the discontinuity, in effect, acts as a waveguide, and as a result, an abnormally long radar range may be observed.

11-753. Atmospheric refraction is a phenomenon by which radar waves are bent in the earth's atmosphere. Under normal conditions the atmosphere is more dense at the surface of the earth and less dense as the altitude increases. As a result, electromagnetic energy travels more slowly at lower altitudes, and is effectively bent downward. The radar horizon is therefore extended about 15 percent beyond the calculated horizon under normal conditions. This phenomenon may be further augmented by a condition known as temperature inversion, which is caused by a warm air mass surmounting a colder air mass. The increased temperature at higher altitudes further decreases normal atmospheric density, as compared to the surface density, and the radar horizon is greatly extended. Temperature inversion is very common where warm air masses from land move over the cool air directly over a large body of water. The opposite condition can also prevail if the gradation of density is reversed, in that a colder air mass surmounts a warm air mass; refraction will then cause the radar wave to bend upward and thus greatly reduce the radar horizon.

11-754. Moisture in the atmosphere may cause microwave signals to be either scattered or reflected, depending upon the size of the droplets. If the droplets are rather large, as in a heavy rain cloud, reflection occurs and causes an echo. This effect is very noticeable at the higher microwave frequencies. Smaller droplets may cause scattering rather than reflection, causing the range to be greatly reduced.

11-755. Atmospheric gases have the property to absorb certain microwave frequencies. Each gas has its own absorption spectrum, and, of the gases studied thus far, each is unique in regard to the absorption of frequencies. For example, water vapor absorbs strongly above 10,000 mc, showing a peak at approximately 23,000 mc. Oxygen absorbs very strongly at about 60,000 mc, and ammonia gas at about 24,000 mc. The fact that the absorption characteristics of various gases differ markedly has made it possible to analyze gases by means of their absorption spectra. The absorption effect is very undesirable in radar operations, because it results in reduced range at the frequencies of maximum absorption. Fortunately, this effect is not pronounced in the X band and at lower frequencies, but it does make the K band very unreliable.

11-756. RADAR FACILITY PERFORMANCE FACTORS. Maximum range performance also depends upon the condition of the radar. Radar condition is the only factor which may be controlled to some degree by maintenance, and therefore is the most important. Radar equipment performance is dependent upon such items as transmitter power, frequency, and spectrum, receiver bandwidth and sensitivity (mds), tr recovery, and afc operation.

11-757. RADAR FACILITY SENSITIVITY. Facility sensitivity is the ratio of the transmitted power admitted by the receiver pass band to the mds power. A precise determination of facility sensitivity, therefore, involves a check of both the transmitter spectrum and the receiver pass band. It follows that if half the transmitted energy is outside of the receiver pass band, the power is effectively cut in half.

11-758. Facility sensitivity is proportional to the fourth power of the maximum range, as shown in the following expression: Range (maximum) $= \sqrt[4]{P_t/P_{mds}}$, where P_t is the transmitter peak power encompassed by the receiver pass band, and P_{mds} is the minimum discernible signal power. Note that in the above expression the fourth root is taken rather than the square root. This is explained by the fact that the inverse square law, which is used to determine the strength of a transmitted signal over a given distance, is applied twice, once for the forward path and once for the echo return.

11-759. When transmitted power and mds power are measured in watts, the sensitivity of the radar facility is calculated in terms of a power ratio by simple division. However, these two power figures are usually measured in dbm, and the facility sensitivity is more conveniently calculated in db as follows: Facility sensitivity (db) = P_t (dbm)— P_{mds} (dbm). (Note that the mds figure is a negative dbm quantity, and, as such, must be subtracted algebraically from the transmitted power.)

11-760. The sensitivity of a radar facility is normally specified in the technical order for the equipment. Any loss of sensitivity results in a corresponding decrease in the maximum range. Figure 11-114 shows graphically how a decrease in system sensitivity, given in db below optimum performance, causes a decrease in the maximum range, given in percentage loss. The shaded area shows the limits of best and worst propagation conditions. The lower curve is based on the fourth root range equation above and represents ideal propagation conditions, typical of the best air search conditions. The upper curve is based on the 16th root of the ratio P_t/P_{mds}, and represents the worst propagation conditions, typical for ground radars operating under substandard atmospheric conditions.

11-761. The dotted curve is based upon the eighth root of the ratio P_t/P_{mds}, which centers it between the extremes, and represents approximate sea search conditions. To see the effect of loss of facility sensitivity upon maximum range, consider a hypothetical radar which has operated under all three propagation conditions shown in the figure. Assume only that under the fourth root condition the maximum range with no facility sensitivity loss is 100 miles. Then the maximum ranges for the three propagation conditions, with optimum facility performance and with the performance 20 db down, are given in table 11-5. It can be seen from this table that the greatest percentage loss in range because of facility sensitivity loss (performance 20 db down) occurs when the propagation conditions are best.

11-762. RESONANCE CHAMBER (ECHO BOX). An echo box, or resonance chamber, consists basically of a resonant cavity, the dimensions of which are determined by the frequency band in which operation takes place. The resonant cavity is tuned by a plunger, which can be adjusted back and forth in the cavity. This plunger is mechanically connected to a calibrated tuning dial. Connection to the radar set is made up by a pickup dipole or a coaxial horn placed in the antenna radiation field, or by means of a

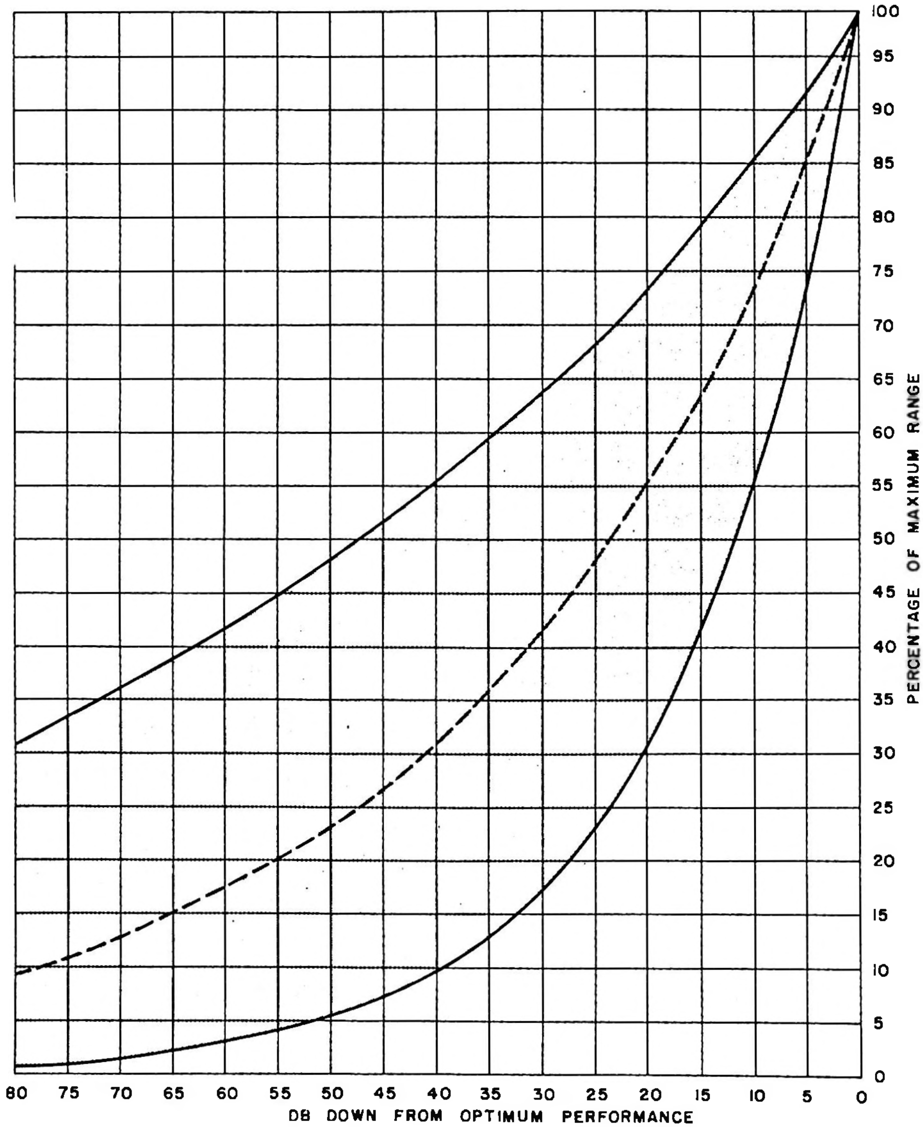


Figure 11-114. Facility Performance Versus Maximum Range

Table 11-5. Estimated Range for Different Propagation Conditions

PROPAGATION CONDITION	MAXIMUM RANGE	
	OPTIMUM PERFORMANCE	PERFORMANCE 20 DB DOWN
4	100 miles	31 miles
8	10 miles	5.6 miles
16	3.2 miles	2.4 miles

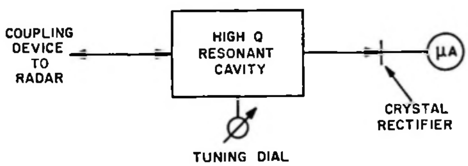


Figure 11-115. Typical Echo Box

cable which is connected to a directional coupler in the transmission line of the radar. An output power meter circuit made up of a microammeter, a crystal, a filter capacitor, and an attenuator (to prevent overloading) are usually included as a part of the echo box test equipment. The output meter indicates the relative power output of the radar transmitter. Refer to the block diagram of an echo box, figure 11-115.

11-763. Any tuned circuit can be shock-excited by a sudden application of energy. When the excitation is removed, the tuned circuit continues to oscillate (ring) for a length of time. The greater the Q of the resonant circuit, the longer the ringtime. In use, the echo box picks up rf energy from a transmitter pulse. When the cavity is tuned to the frequency of the pulse, the rf energy picked up causes oscillations to build up in the resonant cavity. These oscillations continue after the radar pulse is cut off; however, the amplitude of each succeeding oscillation decreases, because of internal losses and output-meter dissipation, and because some of the energy is coupled back to the radar set as shown in figures 11-116 and 11-117. The energy coupled back is detected by the receiver, and appears as a pattern on the radar indicator. See figure 11-118. Ringtime is measured in terms of either yards or microseconds between the start of the transmitter pulse and the point where the ringing signal reaches the noise level of the radar receiver. The value of ringtime is influenced by the following factors: (1) receiver sensitivity; (2) peak transmitter power; (3) coupling loss between echo box and radar; (4) transmitter pulse width; and (5) the Q of the echo box. It should be noted that the first two factors provide a check of system sensitivity. This check, however, is not reliable unless the other three factors are either known or kept constant.

11-764. Because of its simplicity and compactness, the echo box is a very valuable test equipment for periodic facility testing. Keep constantly in mind that an echo box presents relative information. The echo-box installation must first be calibrated with standard test equipment before the information has any practical value. In fact,

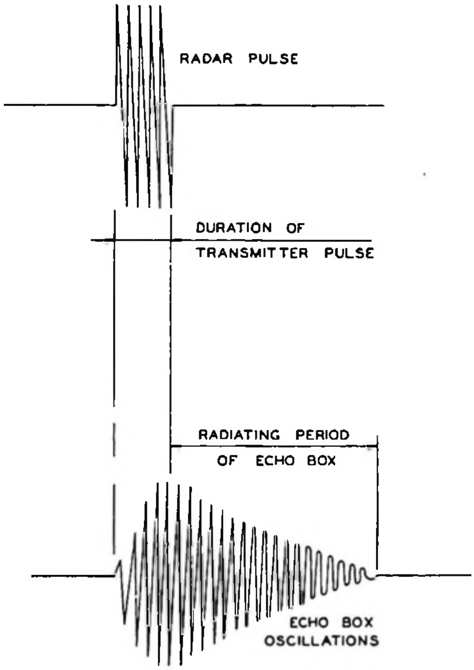


Figure 11-116. Relationship Between Transmitter Pulse and Echo-Box Ringing

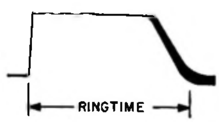


Figure 11-117. Ringtime Indication on A Scope

the echo box must be recalibrated at regular intervals, to ensure that the information gained is reliable.

11-765. The installation of an echo box is very important. If the radar has a directional coupler, the echo box is located at some convenient point, and an rf cable is used to connect the echo box to the coupler. It is important that the same cable and echo box be used for all subsequent testing. If the radar has no directional coupler, a pickup antenna is permanently installed at a point where reflections are at a minimum, and the echo box is connected to the pickup antenna with an rf cable. Again it is important that the same pickup antenna, cable, and echo box be used for any subsequent measurements.

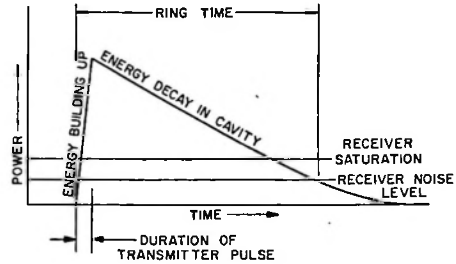


Figure 11-118. Energy Rise and Decay During Ringtime

11-766. The multiresonant type of echo box, which is used to some extent in the field, is made up of a cavity of irregular shape, and of a size corresponding to several wavelengths. Because of its construction, the multiresonant echo box effectively functions as many cavities of different sizes with overlapping response curves. Consequently it is resonant over a broad band of frequencies, and does not require tuning. In most cases a pickup antenna is built into one end of the box. The multiresonant echo box has an extremely high value of Q. However, because frequency and size are interdependent, only those frequencies of the X band or higher permit construction of reasonably sized boxes.

11-767. CALIBRATION. It is possible to calibrate the echo box so that ringtime may be correlated with facility sensitivity; future ringtime readings can then be converted into sensitivity readings. The conversion is easily made, because the change in ringtime per db change in sensitivity is specified for an individual echo box. A common figure encountered in the field is about 100 yards per db. Thus, if a radar has lost 1000 yards of ringtime, the sensitivity has decreased about 10 db. When the ringtime is found to be low, the meter reading is noted, and then compared to the calibrated reading. Since the meter measures relative transmitter output, a low reading indicates trouble in the transmitter. A normal meter reading, coupled with a low ringtime indication, however, points to trouble in the radar receiver. To calibrate an echo-box installation, proceed as follows:

a. Orient both the radar and pickup antennas for maximum pickup. Unless reflections are found to be present, this step represents the final adjustment of the pickup antenna.

b. Record all settings of radar controls that affect prf and pulse width.

c. Adjust radar receiver gain for about one quarter inch of noise on A scope. If an A scope is not used, connect a synchroscope to the receiver output.

d. Tune echo box for greatest ringtime indication on A scope.

e. Adjust coupling in echo box to give a standard meter reading of 75 percent of full scale.

f. Carefully read ringtime; at the same time note the echo-box temperature.

g. Measure facility sensitivity, using standard test equipment and procedure. Refer to paragraph heading 11-620. Ringtime is then correlated with facility sensitivity so that future ringtime readings can be converted to sensitivity readings.

11-768. FACILITY SENSITIVITY TEST. A calibrated echo box is used to measure facility sensitivity by following the steps given below:

a. Orient radar and pickup antennas.

b. Set up radar controls to correspond to the settings recorded during calibration (step b).

c. Adjust receiver gain for a quarter-inch noise amplitude on the A scope.

d. Tune echo box for greatest ringtime indication. Note that maximum meter reading should occur at the same point.

e. Read meter and ringtime. Compare these readings with the previous figures obtained, and convert any change in ringtime reading to the db change of facility sensitivity.

f. Note the temperature of the echo box. If the temperature differs from that recorded during calibration, a correction factor specified in the equipment technical order must be applied to the ringtime reading. Note that at a temperature lower than the calibration temperature, the resistance of the metal of the echo box decreases; hence, the echo box has a higher operating Q and, therefore, a greater ringtime.

11-769. TR RECOVERY TIME CHECK. Tr recovery time is checked by the use of an echo box, as follows:

a. Note the slope of the response between receiver saturation and noise level.

b. Detune echo box until this slope just starts to change.

c. Read ringtime. This reading is the tr recovery time.

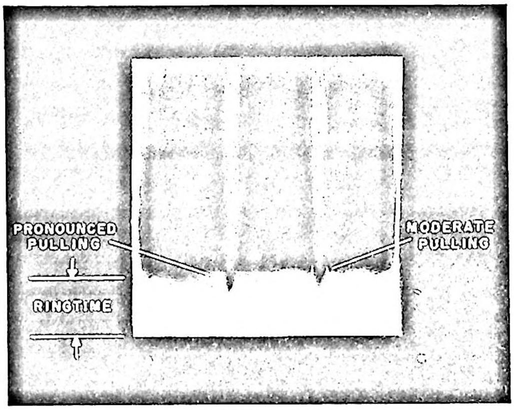


Figure 11-119. Ringtime Indication on B Scope, Showing Effect of Magnetron Pulling

11-770. SPECTRUM ANALYSIS. The spectrum of a transmitter is analyzed by the use of an echo box, as follows:

a. Detune echo box by rotating tuning dial in one direction until meter indicates zero.

b. Rotate tuning dial in the opposite direction, and record the meter readings at various dial positions. Be sure to record all maximum and minimum dial readings.

c. Plot a graph of meter readings (vertical) and dial readings. The graph then represents the transmitter spectrum.

11-771. MAGNETRON OPERATIONAL CHECK. A good means of observing magnetron pulling during scanning is available, if the echo box is coupled to the radar by means of a directional coupler. With the antenna pointing at free space, tune the echo box for maximum ringtime. Start the antenna scanning, and observe the presentation on the ppi or B scope. On a ppi scope, normal magnetron operation produces a smooth, unbroken circle of light; on a B scope, the pattern is a smooth, unbroken bar of light. Magnetron pulling is evidenced by any irregularity or breakup in the edge of the scope indication. Refer to figure 11-119. The azimuth at which pulling occurs is easily read from the indicator. Pulling is seen by watching the echo box meter as the antenna is rotated. If you desire, antenna scanning may be stopped at the pulling point, and the degree of pulling measured.

11-772. PRECAUTIONS. The following precautions should be observed when using an echo box:

a. The same echo box, cables, and pickup device should be used each time the tests are performed.

b. Make certain that the same radar test conditions are established each time the echo box is used. Record all control settings.

c. Measure ringtime very carefully; it is good practice to take several readings and average the results. Use a precision range marker if possible.

d. Keep accurate records, as specified in the technical manuals for the radar facility.

e. Detune or disconnect the echo box when it is not in use.

11-773. FACILITY TROUBLE SHOOTING.

11-774. GENERAL. A radar facility sometimes shows a gradual decline in performance and eventually reaches a point where corrective maintenance is required. On the other hand, the facility may suddenly develop a fault. The suddenly developed fault is immediately obvious. However, in the periodic testing and recording of performance, the purpose is to anticipate possible troubles. Periodic testing shows any trend as it developes, and in many cases minor corrective action at this time prevents a future major breakdown.

11-775. One of the most difficult jobs the technician meets is locating the specific cause of a certain trouble. In localizing the cause of a loss of performance, the first step is to determine whether the trouble is located in a particular unit, such as the modulator-transmitter or the receiver. This may be done by the use of the echo box and the information listed in figure 11-120, or by the use of the various test equipments and procedures described in this section. From there the specific circuits or source of power for that unit are tested. In the following text, both receiver and transmitter troubles are analyzed in detail (the power-supply troubles, being rather straightforward, are not covered).

11-776. LOW RECEIVER SENSITIVITY. Low receiver sensitivity is evidenced by a high mds test figure. The reasons for this condition may be either excessive noise generation or excessive signal loss preceding the i-f amplifier section. As long as the noise present in the receiver output is excessive, the i-f amplifier cannot contribute to the sensitivity. When defective or improperly adjusted, the following items may cause low receiver sensitivity: crystal mixer, local oscillator, i-f amplifier (first two i-f stages), and the tr and atr tubes.

11-777. AFC OPERATIONAL DIFFICULTIES. Because proper operation of the afc circuit is primarily dependent upon coupling between the afc crystal mixer, the local oscillator, and the output of the magnetron, this portion of the circuit should be checked first. A variable coupling is usually provided to adjust the amount of local-oscillator signal injection to the afc crystal mixer. Care should be taken to prevent overloading of the afc crystal mixer, and reference should be made to the applicable technical manual as to the correct crystal current for proper operation. Usually one milliampere of crystal current is the allowable limit. If the degree of coupling is found to be correct, the trouble may lie in a defective local-oscillator tube. The local oscillator must operate smoothly over the desired pull-in frequency range if normal afc operation is to take place. In addition, it is essential to make sure that the local oscillator is operating in the correct mode on the proper side of the transmitter frequency. A magnetron with an improper frequency spectrum may also cause the afc circuit to seem defective.

11-778. POOR MINIMUM-RANGE PERFORMANCE. Minimum-range performance is controlled by the recovery time of the tr tube (and, if used, the pre-tr tube). Excessively long recovery time, of course, indicates the end of the useful life of a tr tube.

11-779. INCORRECT OPERATING FREQUENCY. The trouble of incorrect operating frequency usually breaks down into two possible causes: (1) the magnetron may be defective; or (2) pulling may exist because of some fault in the rf assemblies or from strong reflections from a nearby object.

11-780. When a new magnetron is inserted to correct off-frequency operation, it is not necessarily true that the original magnetron is defective. Individual constructional differences of magnetrons may vary, causing one to be pulled more easily by external conditions than another of the same type number. You can see, then, that irresponsible replacement of apparently defective magnetrons may result in the rejection of good tubes. It is first necessary to check for the presence of pulling, to determine whether the magnetron actually is at fault. This check is made by measuring the swr of the rf assemblies, with the slotted line placed as close to the magnetron as possible, or by feeding the magnetron output into a dummy rf load and rechecking the frequen-

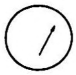
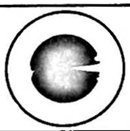
EFFECT	APPEARANCE ON		PROBABLE CAUSE
	RADAR INDICATOR	ECHO BOX METER	
RINGTIME AND TEST SET OUTPUT SATISFACTORY			RADAR PERFORMANCE SATISFACTORY.
RINGTIME LOW, OUTPUT READING SATISFACTORY.			RECEIVER TROUBLE: DETUNED MIXER OR LOCAL OSCILLATOR, BAD CRYSTAL, EXCESSIVE I-F NOISE, ADJUSTMENT OF PROBES IN MIXER CAVITY DETUNED TR BOX.
RINGTIME LOW, TEST SET OUTPUT VERY LOW.			LOW POWER OUTPUT. CHECK SPECTRUM.
RINGTIME LOW, TEST SET METER READING LOW.			TROUBLE PROBABLY IN TRANSMITTER AND RECEIVER AND/OR TROUBLE IN TRANSMISSION LINE.
RINGTIME ERRATIC, TEST SET METER READING STEADY.			TEST SET DETUNED. BAD PULSING, DOUBLE MODING TRANSMITTER, OR LOCAL OSCILLATOR POWER SUPPLY TROUBLE. CHECK SPECTRUM.
RINGTIME ERRATIC, TEST SET OUTPUT READING ERRATIC.			FAULTY TRANSMISSION LINE OR CONNECTION – CONDITION WORSE WHEN LINE IS RAPPED
END OF RINGTIME SLOPES GRADUALLY, PERHAPS EVEN EXCESSIVE RINGING. GRASS APPEARS COURSE. TEST SET OUTPUT READING STEADY AND SATISFACTORY.			OSCILLATING I-F AND/OR NARROW BAND RECEIVER.
PRONOUNCED DIP IN RING-TIME AT END OF PULSE.			FAULTY TR OR DUE TO RECEIVER GATING ACTION
RINGTIME VERY SLIGHTLY LOW, POOR OR BAD SPECTRUM.		 POOR SPECTRUM	TRANSMITTER TROUBLE.
BLANK SPACES OR ROUGH PATTERN ON PPI RINGTIME INDICATOR, TEST SET OUTPUT READING VARIES AS RADAR ANTENNA IS ROTATED.			FREQUENCY PULLING OF TRANSMITTER DUE TO BAD ROTATING JOINT OR TO REFLECTING OBJECT NEAR RADAR ANTENNA.

Figure 11-120. Echo Box Indication of Radar Trouble

cy. When off-frequency operation occurs with a low swr, the indication is that the magnetron should be replaced, unless, of course, it is of the tunable type.

11-781. POOR SPECTRUM. As was previously discussed, spectrum analysis is of considerable importance in the maintenance of a radar facility. The reason for a poor magnetron spectral display or graph can be magnetron pulling or pushing, a defective magnet, or a defective magnetron.

11-782. MAGNETRON PULLING. The test for magnetron pulling is made by means of swr measurements or by the use of a dummy antenna, as mentioned above. Magnetron pulling may cause frequency shift, but this may go unnoticed if the frequency is still within the operating band.

11-783. MAGNETRON PUSHING. A poor spectral display or graph is often evidence of magnetron pushing, and this fault is the result of improper modulator operation. When the output pulse is of improper shape or amplitude, especially at lower power levels, excessive a-m or fm may be present. The test applied to the modulator is made with the aid of a synchroscope and voltage divider. The voltage divider serves to reduce the modulator pulse to a usable amplitude. This amplitude is observed and multiplied by the appropriate factors. The pulse shape is observed and compared with available waveform charts. Under certain conditions, the magnetron causes improper loading of the modulator, and thus introduces pulse distortion. The use of a dummy load for the modulator eliminates this condition. The modulator dummy load is a resistive impedance equal to the firing impedance of the magnetron; in most cases, a voltage divider is built into the test equipment to facilitate the measurement of pulse amplitude. This load replaces the magnetron during pulse measurements. and, therefore, helps to isolate trouble definitely to the modulator.

11-784. DEFECTIVE MAGNET. A poor spectral display or graph often indicates defects in the magnetron magnet. Low magnetic strength may result from careless handling. Improper mounting may cause the magnetic field to enter the magnetron at the wrong angle. Mounting difficulties are quickly found on inspection, and magnetic field strength may be checked by using a gaussmeter. Under some conditions, reversal of the magnet may improve the spectrum.

11-785. DEFECTIVE MAGNETRON. A poor spectral display or graph may indicate a defective magnetron. A weak magnet may cause the magnetron input to exceed rated values; if so, continued operation results in a damaged unit. Missing lines in the spectral display are the result of magnetron arcing, and, if excessive, may completely destroy the shape of the spectrum. Many magnetrons display moderate arcing until seasoning is completed, and, therefore, should be allowed a sufficient breaking-in period before the spectrum is analyzed. As mentioned previously, the end of the useful life of a magnetron is characterized by an increase in arcing and general instability. When the output power is low, it usually indicates a weak magnetron or a low modulator output. This uncertain condition may be resolved by testing the modulator output pulse; normal pulse indicates that the trouble is in the magnetron.

11-786. BEAM WIDTH DETERMINATION.

11-787. The upward trend in frequency for radar and microwave communication uses has been largely the result of tremendous antenna gain which can be realized by using moderate size paraboloidal reflectors as a part of the antenna system. The reason for this is that, for a given size paraboloid, an increase in frequency produces a decrease in radiated beam width. The equation for approximating beam width at the half-power points is:

$$\theta \approx 70 \frac{\lambda}{D}$$

where:

θ = beamwidth in degrees

$$\left.\begin{array}{l}\lambda = \text{wavelength} \\ D = \text{width of paraboloid}\end{array}\right\}\text{in the same units}$$

Further information regarding determination of antenna field paths is included under paragraph heading 11-457.

11-788. RADAR MTI TESTING.

11-789. Practically all types of modern surface-to-air search radar equipment employ moving-target indication (mti) receiving equipment in addition to the normal receiver. The mti receiver output consists of a video signal in which, at least theoretically, all returns from fixed targets have been eliminated, leaving only moving targets to be seen on the mti display. The ground clutter which masks targets on the normal display is therefore removed on the mti display, enabling you to "see" moving targets which would otherwise go undetected.

11-790. The echoes from moving targets differ in two important respects from echoes from fixed targets. The elapsed time between the transmission of a pulse and the return of an echo from a target is dependent upon the target range. Since this range is constant for a fixed target (assuming the location of the radar to be fixed), the elapsed time does not change from one transmitted pulse to the next. However, since the range of a moving target is continually changing, the elapsed time between the transmission of a pulse and the reception of an echo will vary from one pulse to the next, increasing if the target is moving away from the radar and decreasing if the target is moving toward the radar.

11-791. The second important difference between the echoes from moving targets and those from fixed targets concerns the frequency of the echo signal. The frequency of the returns from a fixed target is the same as that of the transmitted pulses, while the frequency of the returns from a moving target differs from that of the transmitted pulses by an amount dependent upon the relative velocity of the target. This difference frequency is known as the Doppler frequency, and, except for a few special values, is not important in mti systems.

11-792. All modern surface-to-air search radar equipment is of the pulsed type, in which the mti receiver uses the pulse-to-pulse difference in elapsed time between the transmission of a pulse and the return echo from the target. Since this difference is often extremely small (as low as 150 picoseconds), conventional range-measuring techniques cannot be used. The mti receiver measures these minute time differences by phase comparison devices which measure the relative phase, pulse-to-pulse, of the echo signal and a cw reference signal which is synchronized by the transmitter at every pulse.

11-793. TYPICAL MTI EQUIPMENTS.

11-794. Each of the mti equipments presently in use can be divided into two major sections. In a normal radar equipment, the receiver contains rf and i-f stages in which the received signals are amplified and converted to video. The video section of the receiver then amplifies and processes the video for application to a cathode-ray tube. Similarly, the mti receiver contains rf and i-f sections which amplify a received signal

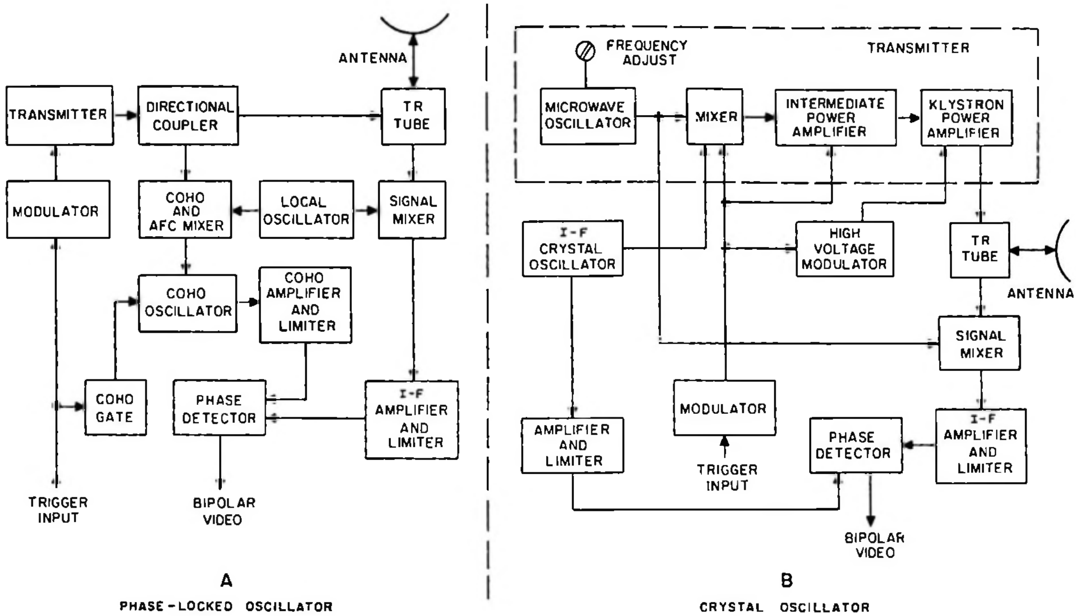


Figure 11-121. Two Common Methods of Producing MTI Bipolar Video

and convert it to video. However, since pulse-to-pulse target motion is detected by a phase comparison method, phasing circuits are also found in this section, and the output is a bipolar video in which moving targets are represented by a signal which varies in both amplitude and polarity. The video section of the mti receiver also amplifies and processes the video signal for application to a cathode-ray tube. However, since fixed targets are to be eliminated, the video section contains cancellation circuits and a bipolar rectifier.

11-795. Part A of figure 11-121 illustrates one method of producing the required video signal. The received signal is mixed with the output of the local oscillator, and the resulting i-f signal is amplified and limited. Limiting is necessary to ensure that signals fed to the phase detector do not vary in amplitude, since amplitude variation causes errors. (Weak signals are obviously not limited; however, such signals are not normally of interest at the comparatively short ranges of mti.) To provide a reference phase, a small portion of the transmitted rf signal is coupled to the coho and afc mixer and mixed with the local-oscillator output to produce an i-f signal, which lasts only for the duration of the transmitted pulse. This i-f signal is used to phase-synchronize a coho oscillator operating at the i-f frequency and turned on by the coho gate at the start of the transmitted pulse. (The term coho is derived from coherent, indicating that the oscillator is phase-locked.) The coho is a frequency-stable free-running oscillator which operates at the i-f frequency until cut off by the coho gate. It thus provides a reference phase i-f signal which is synchronized at each transmitted pulse. The phase detector compares the phases of the signal i-f and the coho i-f, and produces

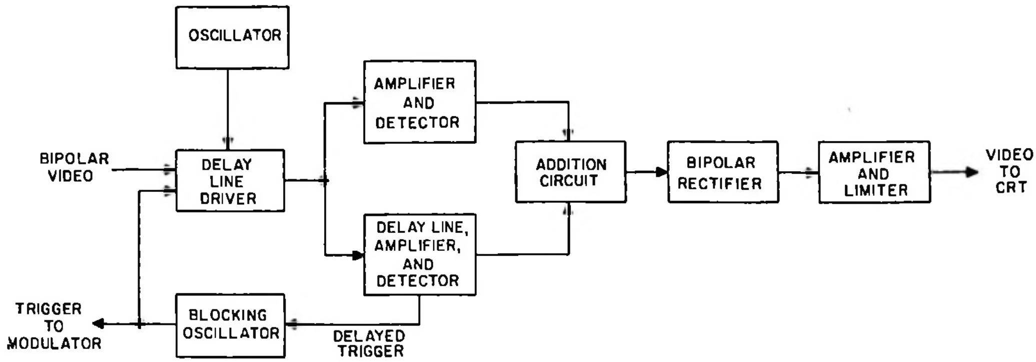


Figure 11-122. Video Section of a Typical MTI Receiver

a dc output which varies in both amplitude and polarity, the amplitude being dependent upon the amount of phase difference (maximum at 90-degree phase difference and zero at 0- or 180-degree phase difference) and the polarity upon the direction of this phase difference (0 through 90 to 180 degrees or 0 through 270 to 180 degrees). A fixed-target return does not vary in phase between any two consecutive pulses, and therefore produces an output which is fixed in both amplitude and polarity. In contrast, a moving-target return varies slightly in phase between consecutive pulses, and produces an output which varies in both amplitude and polarity.

11-796. Part B of figure 11-121 shows an alternate method of producing bipolar video which is used on a number of new types of equipment. The advantage of this method lies in the elimination of the phase-locked circuit and the use of a crystal oscillator as the coho oscillator. This circuit can be used where the transmitter consists of a microwave oscillator feeding a klystron power amplifier tube through a number of intermediate power amplifier (ipa) stages. The ipa stages and the klystron are tuned to the difference of the microwave oscillator frequency and the i-f crystal oscillator frequency. The microwave oscillator is also used as the local oscillator to produce the i-f signal. The crystal oscillator output is therefore the reference phase, and is compared (after amplification and limiting) with the amplified and limited target return in the phase detector to produce bipolar video.

11-797. Figure 11-122 is a block diagram of a typical video section of an mti receiver. The bipolar video is applied to a driver stage, where it amplitude-modulates the output of an oscillator whose frequency is on the order of 10 mc. The modulated rf is then fed to a delay-line amplifier and a conventional amplifier. These two circuits have the same gain but produce outputs of opposite polarity. In addition, the video through the delay-line amplifier is delayed by a time equal to one pulse repetition period. Thus the addition circuit adds the video of a given pulse period (through the conventional amplifier) and the delayed and inverted video from the preceding pulse repetition period (through the delay-line amplifier). Fixed-target signals, for which the amplitude and polarity remain constant, are thus cancelled, and moving-target signals, for which both amplitude and polarity vary from one pulse period to another, are not cancelled. The bipolar rectifier is similar to a full-wave rectifier, and provides an output independent of the signal polarity. This signal is amplified, limited, and applied to the cathode-ray tube.

11-798. In order to ensure that the delay line delays the video by exactly one pulse period to permit cancellation, the delay-line time controls the prf of the radar. This is accomplished by mixing the modulator trigger from the blocking oscillator with the bipolar video, taking it off through the delay-line amplifier as a delayed trigger. This delayed trigger is coupled to the blocking oscillator, causing it to produce the next trigger, which is in turn delayed, and so on. Thus, by circulating the trigger through the delay line, the pulse repetition time is made exactly equal to the delay time.

11-799. MEASUREMENTS.

11-800. CANCELLATION RATIO. The cancellation ratio is a measure of the effectiveness of cancellation of fixed targets in an mti receiver. It is obtained by measuring the amplitude of an uncancelled fixed target and the amplitude of the uncancelled residue of the same target with the cancellation circuits in operation and determining the difference in amplitude. The amplitudes are measured at the output of the mti circuits, using an rf signal generator and a synchroscope.

11-801. In making the measurements, the output from a pulsed signal source is coupled into the duplexer or, in some equipments, into a built-in input connector. Antenna rotation is stopped in a direction which allows display of a suitable fixed target. The mti is then disabled by changing the prf or by causing the coho oscillator to become noncoherent (thus effectively disabling the cancellation circuits). The signal generator pulse amplitude is then adjusted by means of the calibrated attenuator until it is equal to the amplitude of the uncancelled target return, and the attenuator dial reading noted. Upon restoration of the mti to normal operation, the amplitude of the fixed target should be greatly reduced. The calibrated attenuator on the signal generator is adjusted until the signal generator output pulse is equal in amplitude to the uncancelled residue of the fixed target, and the attenuator dial reading again noted. The cancellation ratio is the difference of the two values of attenuation, and is usually given in db. The actual value depends upon the type of equipment, but is usually on the order of 20 to 30 db.

11-802. SUB-CLUTTER VISIBILITY. The sub-clutter visibility is a measure of ability to detect a small moving target in ground clutter. The test consists of determining the difference in amplitude between a cancelled fixed target and a moving target at the same range. Ideally, any target visible in the random noise should be visible over a fixed target, but because of the fact that a fixed target leaves some uncancelled residue, this condition is not generally attained.

11-803. In performing this test, the cancellation circuits are disabled as in the previous test, and the antenna is rotated until it is centered upon a suitable fixed target. The receiver gain is then lowered until the target return is not limited, and an rf signal is introduced from the signal generator. (Since the signal generator oscillator is not phase-locked to the transmitted rf energy, the output pulse appears as a moving target.) The signal generator pulse delay is adjusted until the pulse appears next to the fixed target in range, and the calibrated attenuator is varied until the signal generator pulse is equal in amplitude to the fixed target. The dial reading is then noted.

11-804. The mti equipment is restored to normal operation, and the signal generator pulse delay is adjusted until the moving target (signal generator pulse) coincides in range with the fixed target. The calibrated attenuator is then varied until the moving target is just visible above the fixed target. This is accomplished by increasing the attenuation until the target is invisible and then slowly decreasing the attenuation until the moving target is just visible. The attenuation value is noted. The sub-clutter visibility is the difference of the two attenuator dial readings, and is usually expressed in db. The actual value depends upon the type of equipment, but is generally on the order of 15 to 20 db for older equipments and may be as high 35 to 40 db on the most modern types of equipment.

11-805. MTI TROUBLES.

11-806. GENERAL. Moving-target-indicator circuits are susceptible to the troubles commonly encountered in other electronic circuits. Since the delay line and the coherent oscillator are common only to the mti circuit, these will be discussed briefly.

11-807. EXCESSIVE DELAY LINE ATTENUATION. Excessive delay line attenuation may be caused by dirty mercury. The mercury should be drained and then replaced with clean mercury. For detailed instructions, refer to the applicable technical manual. When a delay line is refilled with mercury, or if it has been subjected to severe vibration, excessive attenuation may result. Allowing the delay line to rest for approximately 24 hours results in the mercury settling and the condition is corrected.

11-808. TOTAL FAILURE OF DELAY LINE. Total failure of the delay line is usually caused by either broken crystals or by a shorted coaxial cable connector in one of the two transducers. A broken crystal is apparent by leakage of mercury at the transducer tank. If no leakage is present, it may be assumed that the trouble is in the coaxial connectors. If a crystal must be replaced, refer to the applicable technical manual for detailed instructions.

11-809. COHERENT OSCILLATOR TUNING. The coherent oscillator has been designed to be exceptionally stable. The usual frequency tolerance of the mti circuits requires not more than a 10-cycle deviation in frequency during the interval between transmitting pulses. Because of this extremely small tolerance, the tuning of the coherent oscillator should be checked carefully and if necessary, readjusted. Since different types of mti equipments are so much different, no attempt will be made to describe any special procedure. For complete detailed instructions, refer to the applicable technical manual.

11-809A. ONE-MAN PORTABLE RADAR SETS.

11-809B. GENERAL.

11-809C. The purpose of this dissertation pertaining to one-man portable radar sets is to provide the radar repairman with a fundamental background on the repair and the testing of small compact radar units. The AN/PPS-4 Radar Set is a typical one-man portable radar set. A number of small compact radar sets ranging in weight from approximately 2 lbs to 125 lbs have been developed in recent years, but most of them follow the basic concepts of the AN/PPS-4. These small radar sets have proved to be very successful when used for military purposes. One of the sets, weighing approximately 2 lbs, when attached to a rifle, is used to detect human movement at distances up to 250 yards. In this connection, such sets have even proved effective where tall grass has afforded protective cover for enemy movement. Commercially, these radar sets have been used by police to determine the speed of vehicles. Large cities throughout the world are using, or are planning to use, radar to control traffic by detecting areas containing an excess of slow-moving vehicles and then to automatically switch traffic lights so that traffic can move more freely. Because of the growing demand for small radar sets, the following dissertation has been included in this manual.

11-809D. BASIC THEORY.

11-809E. The general functioning of Radar Set AN/PPS-4 can be explained best by dividing the set into nine basic units, each of which performs one or more specific functions. Figure 11-122A illustrates the AN/PPS-4 with the control panel removed and callouts showing the location of major components. The radar set is composed of the following units:

a. Transmitting equipment.

b. Rf unit.

c. Receiving equipment.

d. Automatic frequency control (afc) unit.

e. Ranging group.

f. Audio equipment.

g. Automatic gain control (agc) unit.

h. Power converter components.

i. Control unit.

11-809F. TRANSMITTING EQUIPMENT. The transmitting equipment generates the radio-frequency (rf) pulse which is applied to the rf unit for radiation into space. The transmitter also generates the negative blanking pulse which is sent to the receiving equipment to blank the receiving signal intermediate frequency (i-f) amplifiers during the transmitted pulse interval. The negative gate which is used to start the linear sweep in the ranging groups generated, also transmits from the ranging unit the multiar pulse whose negative overshoot terminates the negative gate. The rf pulse has peak power of 0.5 kilowatt and is transmitted at a frequency between 8,999 and 9,400 megacycles (mc). The pulse width of the rf pulse is 0.2 microsecond and the pulse repetition frequency is approximately 5,000 pulses per second (pps). The transmitting equipment is composed of the components in the transmitter unit and the magnetron. The transmitter unit is mounted on the rear side of the center section of the receiver-transmitter. The magnetron is mounted on the microwave assembly which, in turn, is housed in the front side of the center section of the receiver-transmitter.

11-809G. RF UNIT. The rf unit radiates, in a narrow beam, the rf pulse generated by the transmitting equipment and receives the target echo pulse which is applied to the receiving equipment. The rf unit also attenuates and couples a magnetron pulse to the afc unit for mixing with the local oscillator signal.

11-809H. The rf unit contains a duplexer which prevents high-power magnetron pulses from damaging the components of the receiving equipment during transmit time, and which permits the target echo to pass to the receiving equipment during the time that the transmitter is not transmitting. The rf unit also contains a microwave assembly and a parabolic reflector. The microwave assembly is mounted on the front side of the center section of the receiver-transmitter. The microwave assembly is composed of an antenna feed assembly, a hybrid waveguide assembly, a transmit-receiver (tr) tube, a directional coupler, a flexible waveguide assembly, an afc coupler, and the crystalholders in which the signal and afc mixer crystals are mounted. The parabolic reflector is attached to the inside of the radome which is mounted on the front of the center section of the receiver-transmitter.

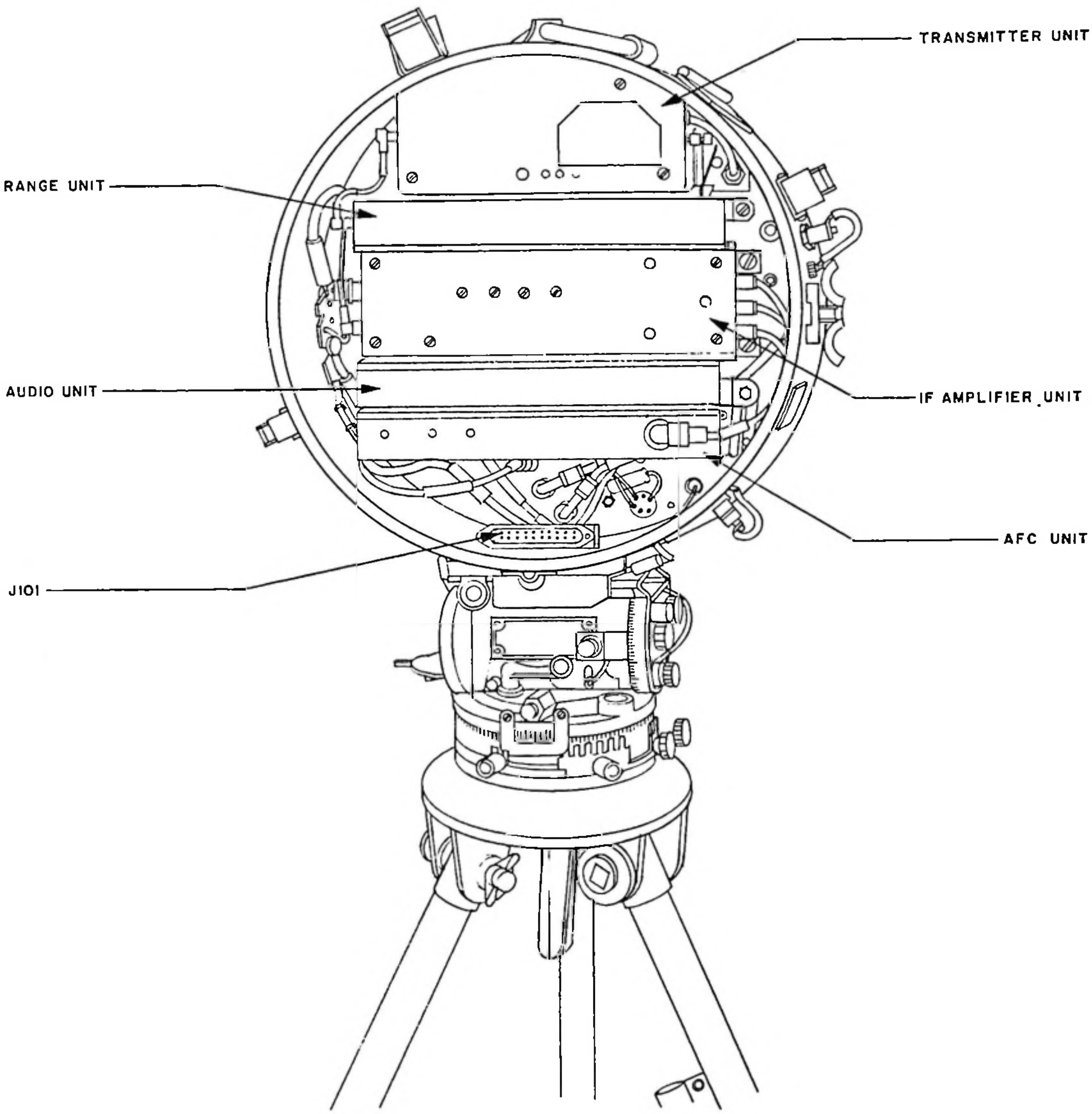


Figure 11-122A. Radar Set AN/PPS-4, Control Panel Removed

11-809I. RECEIVING EQUIPMENT. The receiving equipment generates the local oscillator signal, mixes the target echo pulse with the local oscillator signal to develop i-f signals of 30 mc, amplifies the i-f signals, detects the amplified i-f signals to produce video signals which are amplified and sent to the ranging group and receives, amplifies, and detects 30-mc rangemarks from the afc unit when the power switch is in the range position. In addition, the receiving equipment receives an agc voltage which controls the overall gain of the receiving equipment and receives the negative blanking pulse which blanks the receiving equipment signal amplifiers during the transmitted pulse interval. The receiving equipment is composed of a klystron local oscillator, signal mixer crystals and the components in the i-f amplifier unit. The klystron local oscillator and signal mixer crystals are attached to the microwave assembly which is mounted on the front side of the receiver-transmitter center section. The i-f amplifier unit is mounted on the rear side of the center section of the receiver-transmitter.

11-809J. AUTOMATIC FREQUENCY CONTROL UNIT. The automatic frequency control unit receives an attenuated magnetron pulse from the rf unit and mixes this pulse with the local oscillator signal from the receiving equipment. The afc unit then develops a direct current (dc) voltage, which controls the frequency of the local oscillator so that the i-f signal is maintained at 30 mc, and generates precision rangemarks which are used for range calibration of the ranging group. The afc unit is composed of an afc mixer crystal, the components of the afc unit, and the afc search and control circuit contained in the audio unit. The afc mixer crystal is attached to the microwave unit which is mounted on the front side of the receiver-transmitter center section. The afc unit and the audio units are mounted on the rear side of the center section of the receiver-transmitter.

11-809K. RANGING GROUP. The ranging group receives from the transmitting equipment the negative gate which starts a linear sweep, receives the video signal from the receiving equipment, and generates and sends to the audio and agc units a boxcar detector voltage whose amplitude is dependent upon the position of the range gate and the nature of the target being gated. The ranging unit also receives from the agc unit the agc voltage which is converted and displayed on the range extension meter, and provides current to deflect the range extension meter according to the mode of operation. In addition, the ranging group also generates a multiar pulse which terminates the negative gate from the transmitting equipment and receives from the control unit the range voltage which determines the range of the detected target. The ranging group is composed of the components in the range unit and the strobe unit. The range unit is mounted on the rear side of the center section of the receiver-transmitter and the strobe unit is mounted on the rear of the control panel of the receiver-transmitter.

11-809L. AUDIO EQUIPMENT. The audio equipment filters and amplifies the boxcar detector voltage from the ranging group and produces an audio signal which can be detected in your headset as a sound, indicative of a moving target. The audio equipment is composed of the audio unit, audio transformer T1, audio filter FL3, and rf filters FL103 and FL104. The audio unit is mounted on the rear side of the center section of the receiver-transmitter. Audio transformer T1 and audio filter FL3 are mounted on the front side of the center section of the receiver-transmitter. Rf filters FL103 and FL104 are mounted on the rear of the control panel of the receiver-transmitter.

11-809M. AUTOMATIC GAIN CONTROL UNIT. The automatic gain control unit receives the boxcar detector voltage from the ranging group, develops an agc voltage which controls the gain of the signal i-f amplifiers in the receiving equipment, and provides current to deflect the needle of the range extension meter when targets are detected. The agc unit is composed of two transistor stages located in the audio unit, and a potentiometer mounted on the rear of the control panel of the receiver-transmitter.

11-809N. POWER CONVERTER COMPONENTS. The power converter receives a negative dc voltage (—22.5 to —27.5) from an external power source, converts the dc input voltage to a square wave, steps the voltage up or down, as necessary, by transformer action, rectifies and filters part of the secondary output to obtain the dc voltages for use in the radar set, and provides the following dc voltage to other functional equipments in the radar set: +300, +120, —20, —165, and —550, along with the root-mean square (rms) voltages of 6.3, 7.0, and 2.7. The power converter components are composed of the power converter unit, capacitor C6, and inductor L1, all of which are mounted on the front side of the center section of the receiver-transmitter.

11-809O. CONTROL UNIT. The control unit provides a means of turning the radar set on and off, protecting the circuits in the radar set, selecting the modes of operation, and visually and aurally indicating detected targets. The control unit also indicates the range of the detected targets; visually indicating the position of the rangemarks and calibrating the ranging group. The control unit provides an arrangement which permits you to connect an external power source to the radar set, to adjust the voltage output of the power converter, to test the condition of the battery charge when a battery is used as the external power source, and to turn the set on and off. You can adjust the intensity of the lamps in the radar set, the volume of the aural indication of detected targets, and also select the strobe mode of operation. The control unit is composed basically of a range extension meter, variable resistors, switches, fuses, and receptacles, all of which are located on the rear of the control panel, the range meter indicator, and the range control handwheel. In addition, a battery test circuit on the strobe unit is considered as part of the control unit.

11-809P. TEST PROCEDURE FOR THE AN/PPS-4. The information provided herein is a guide to test the Radar Set AN/PPS-4. Generally, these test procedures will apply to all one-man portable radar sets. Perform the complete test procedures in the given order to obtain maximum results. When the use of test equipment mentioned in these test procedures is not clearly understood, refer to chapter 8 of this manual for a more detailed explanation. The following is a list of the 13 separate tests that are needed to completely test this radar set.

a. Magnetron Filament Voltage Test

b. Receiver Threshold Test

c. Afc Performance Test

d. Trigger Output Test

e. Magnetron Current and Trigger Output Pulse Test

f. Pulse Width, Repetition Rate, and Power Output Test

g. Short Strobe Test

h. Long Strobe Test

i. Range Gate Test

j. Detector Test

k. Audio Gain Test, Receiver Sensitivity Test

l. Doppler Sensitivity Test

CAUTION

Care should be exercised in the handling of all parts of the AN/PPS-4 Radar Set to prevent damage to small components, meter M101, and exposed soldered connections.

11-809Q. MAGNETRON FILAMENT VOLTAGE TEST. The magnetron used in this radar set is a variable, resonant cavity, magnetron oscillator which generates short pulses of high-power rf energy at frequencies from 8900 mc to 9400 mc. The magnetron receives a pulse of approximately 4.5 kv, having a duration of 0.2 microsecond, from the pulse transformer of pulse unit 2201 and delivers rf energy in the form of a nominal 0.5-KW pulse to the rf unit. The magnetron filament voltage is checked by measuring the voltage across pins 1 and 3 of pulse unit 2201 shown in figure 11-122B, with the power switch in the standby position. Perform the following sequence of steps to test the magnetron filament voltage.

WARNING

Perform this test with the POWER switch in the standby position or 5000 volts will be present at the magnetron filaments.

a. To gain access to pulse unit Z201, release the four trunk-type latches that fasten the control panel on the receiver-transmitter and remove the control panel.

b. Place the control panel in a support structure made from non-conductive material. The use of this non-conductive support structure will prevent shorting of the panel's exposed solder connections when power is turned on.

c. Connect the repair patchcord between plug P101 and jack J101 on the receiver-transmitter center section. The repair patchcord is furnished as part of the AN/PPS-4 Radar Set and its purpose is to permit operation of the radar set with the control panel removed. With the repair patchcord inserted between plug P101 and jack J101, the equipment is presented so that you can test the radar set with the power turned on.

d. To gain access to pins 1 and 3 of pulse unit 2201, lift off the cover of the transmitter unit after unscrewing four binding head screws.

e. Set the oscilloscope of measure ac voltage and apply the oscilloscope test probes to pins 1 and 3 of pulse unit Z201. A Tektronix Model 535A oscilloscope, or equivalent, is required for the test measurement.

f. Set the POWER switch to the STANDBY position. The oscilloscope should indicate a square wave, 14 $\pm$ 2 volts peak-to-peak in amplitude. If the indication is not as designated, the probable cause of the trouble would be in the power supply.

g. Set the POWER switch in the OFF

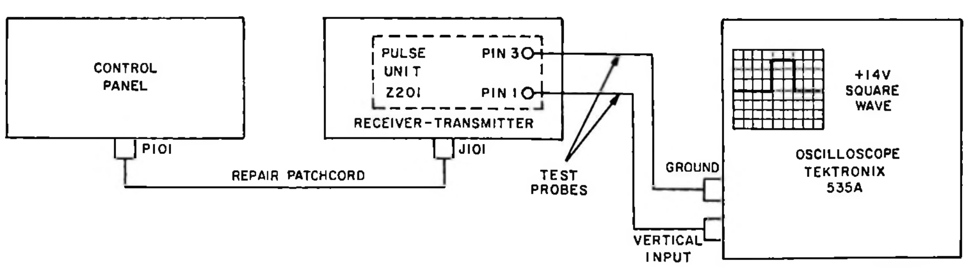


Figure 11-122B. Magnetron Filament Voltage Test Setup position. Remove the oscilloscope test probes from pins 1 and 3 of 2201 and replace the cover on the transmitter unit.

h. Remove the repair patchcord from plug P101 and jack J101. Replace the control panel on the receiver-transmitter and fasten the four trunk-type latches to secure the control panel in place.

11-809R. RECEIVER THRESHOLD TEST. In this radar set, an audio tone represents a moving target. A limiting circuit for the amplitude of the audio signal is provided as part of the i-f amplifier unit. The receiver threshold test checks the input signal level at which the radar begins to decrease the gain of the i-f amplifier unit. A block diagram of the receiver threshold test setup is shown in figure 11-122C. Perform the following sequence of steps to test the receiver threshold voltage:

a. To gain access to the test points specified in this test procedure, remove the control panel from the rear of the receiver-transmitter center section. Place the control panel in a support structure made of a non-conductive material.

b. Connect the repair patchcord between plug P101 on the control panel and jack J101 on the receiver-transmitter center section.

c. Connect a variable power supply to the 24-vdc power-input receptacle on the control panel. Adjust the variable power supply for an output of 24 volts dc. The variable power supply used in this test is a type PP-1104/G, or equivalent. The PP-1104/G power supply has a range of from 0 to 32 volts with a maximum capacity of 4 amperes.

d. Place the POWER switch in the TRANSMIT position and wait approximately 90 seconds for the thermal-delay relay to energize. The thermal delay circuit delays the application of high voltage for approximately 90 seconds in order to allow the tube filaments of the radar set to heat.

e. Using a tip jack plug, connect the probe of an oscilloscope, Tektronic Model 535A or equivalent, to jack J3 on the range unit. Use the output jack, J1, of the range unit as an external trigger for synchronizing the oscilloscope to the radar set.

f. Turn the RANGE CONTROL handwheel, located on the control panel, until a reading of greater than 3,000 meters is obtained on the range indicator. The RANGE CONTROL handwheel determines the range of the radar set by gating the output of the blocking oscillator circuit.

g. Remove the oscilloscope probe from jack J3 on the range unit and ground it to establish a ground reference amplitude on the oscilloscope.

h. Return the probe to jack J3 of the range unit, and set the oscilloscope input control for ac input.

i. Adjust the VOLUME control on the control panel until the baseline of the noise waveform on the oscilloscope is 0.2 volt above the ground reference. The VOLUME control in the control panel controls the audio signal level applied to the headset that you use.

j. Remove the cover plate from the directional coupler access hole located in the receiver-transmitter housing, and unscrew the cap which fits over the directional coupler cable assembly. Removal of this plate permits access to the directional coupler jack J908.

k. Connect the output cable of the spectrum analyzer test set to directional coupler jack J908. The spectrum analyzer test set used in this test is an AN/UPM-33 or equivalent.

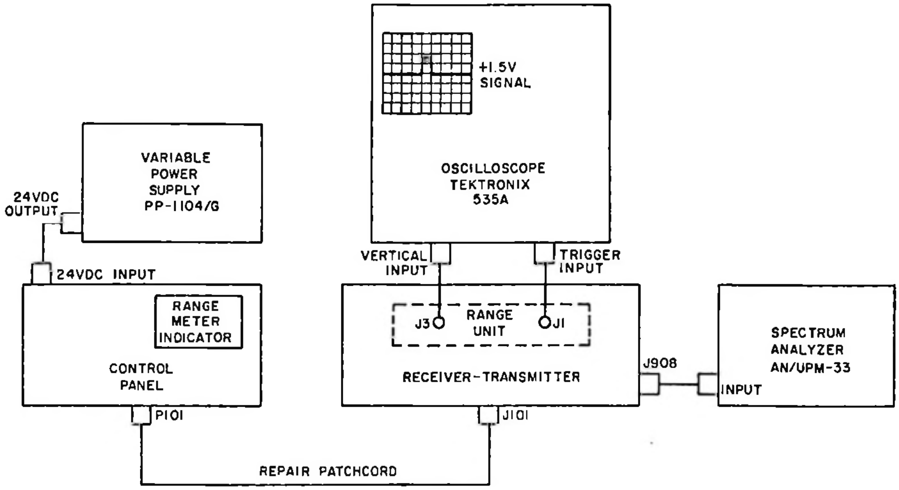


Figure 11-122C. Receiver Threshold Test Setup

l. The amplitude of the signal observed on the oscilloscope should be 1.5 ± 0.2 volts.

m. Set the POWER switch in the OFF position. Remove the test cables and replace the cover plate on the directional coupler.

n. Remove the repair patchcord between plug P101 and J101. Replace the control panel on the receiver-transmitter and fasten the four trunk-type latches to secure the control panel in place.

11-809S. AFC PERFORMANCE TEST. The automatic frequency control unit receives an attenuated magnetron pulse from the rf unit, mixes this pulse with the local oscillator signal from the receiving equipment, develops a dc voltage which controls the frequency of the local oscillator so that the i-f signal is maintained at 30 mc, and generates precision rangemarks which are used for range calibration of the ranging group. A spectrum analyzer is used to test the afc unit performance. It will not be necessary to remove the control panel during this performance check, because all measurements are performed on the accessible front of the receiver-transmitter center section. Refer to figure 11-122D for a block diagram of the performance test setup. Perform the following sequence of steps to test the afc performance:

a. Remove the cover plate from the directional coupler J908 access hole in the receiver-transmitter housing, and then remove the directional coupler cap. Removal of the cover plate will permit access to jack J908 with the pickup horn of the spectrum analyzer.

b. Place the pickup horn of the spectrum analyzer over the directional coupler jack J908 access hole. The pickup horn will transmit signals from jack J908 to the spectrum analyzer. Use a AN/UPM-33 Spectrum Analyzer, or equivalent, for this test.

c. Set the POWER switch on the control panel to the TRANSMIT position, depress the BATTERY TEST switch, and adjust VOLTAGE ADJ. switch S102 until the needle of the RANGE EXTENSION METER, M101, indicates the center of the red zone.

d. Observe the frequency spectrum displayed on the spectrum analyzer oscilloscope. The klystron frequency should be locked on at 30 mc ± 1 mc above the magnetron frequency.

NOTE

Use the higher frequency display of the two obtainable responses.

e. Set the POWER switch on the control panel to the STANDBY position, wait 5 seconds, and return the POWER switch to the TRANSMIT position. Correct lock-on should occur within 5 seconds after the POWER switch is returned to TRANSMIT.

f. Repeat step e with VOLTAGE ADJ. switch S102 advanced one position above its normal position.

g. Repeat step e with VOLTAGE ADJ. switch S102 retarded one position below its normal position.

h. The klystron frequency should be locked on at 30 mc ± 1 mc above the magnetron frequency.

i. Set the POWER switch in the OFF position. Remove the spectrum analyzer pickup horn and replace the directional coupler J908 cap and access hole cover.

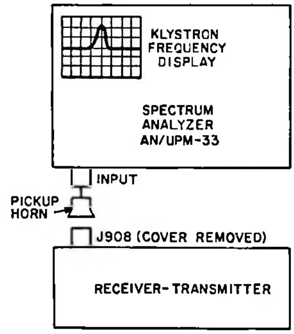


Figure 11-122D. Afc System Performance Test Setup

11-809T. TRIGGER OUTPUT TEST. The trigger provided by the pulse-forming circuit of the transmitter unit is checked during the performance of this test. The trigger is normally applied to the i-f amplifier of the receiving equipment. However, in the performance of this test, the trigger will be disconnected from the radar set and reconnected to an oscilloscope so that it can be visually observed. Refer to figure 11-122E for a block diagram of the trigger test setup. Perform the following sequence of steps to test the trigger output of the pulse-forming circuit of the transmitter.

a. To gain access to the test points used in this test, release the four trunk-type latches that fasten the control panel to the receiver-transmitter, and remove the control panel. Place the control panel in a non-

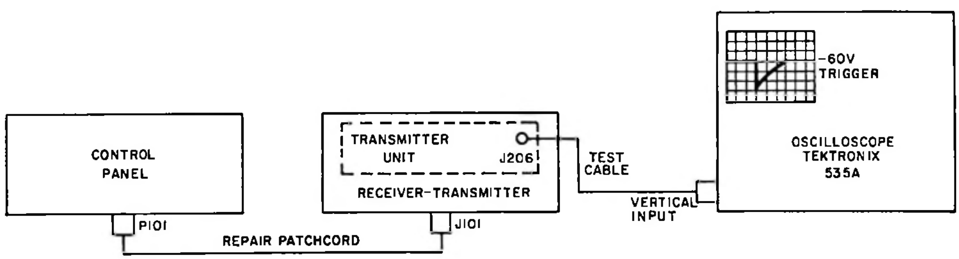


Figure 11-122E. Trigger Output Test Setup conductive support structure to prevent shorting of exposed connections.

b. Connect the repair patchcord between plug P101 on the control panel and jack J101 on the receiver-transmitter center section. With the repair patchcord in place the radar can be operated and the test points can be reached with a test probe.

c. Connect the oscilloscope test probe to jack J206 on the transmitter unit. The oscilloscope used in this test is a Tektronix Model 535A, or equivalent.

d. Set the POWER switch to the TRANSMIT position. After a warmup period of approximately 90 seconds, a negative 60 ± 6 volt waveform should appear on the oscilloscope.

e. Set the POWER switch to the OFF position. Remove the test probe from jack J206 and remove the repair patchcord between jack J101 and plug P101.

f. Replace the control panel on the receiver-transmitter center section, and fasten the four trunk-type latches to secure the control panel in place.

11-809U. MAGNETRON CURRENT AND TRIGGER OUTPUT PULSE TEST. The magnetron current is checked by measuring the voltage between jacks J4 and J5 of the transmitter unit. The trigger output pulse is checked by observing the waveform at jack J1 on the range unit. Refer to figure 11-122F which shows the test setup required to test the magnetron current and trigger output pulse. Perform the following sequence of steps to test the magnetron current and trigger output of the transmitter:

a. Release the four trunk-type latches that fasten the control panel to the receiver-transmitter center section. Place the control panel in a non-conductive support structure to prevent shorting of exposed connections.

b. Connect the repair patchcord between plug P101 on the control panel and jack J101 on the receiver-transmitter center section.

c. Set the POWER switch to the TRANS-

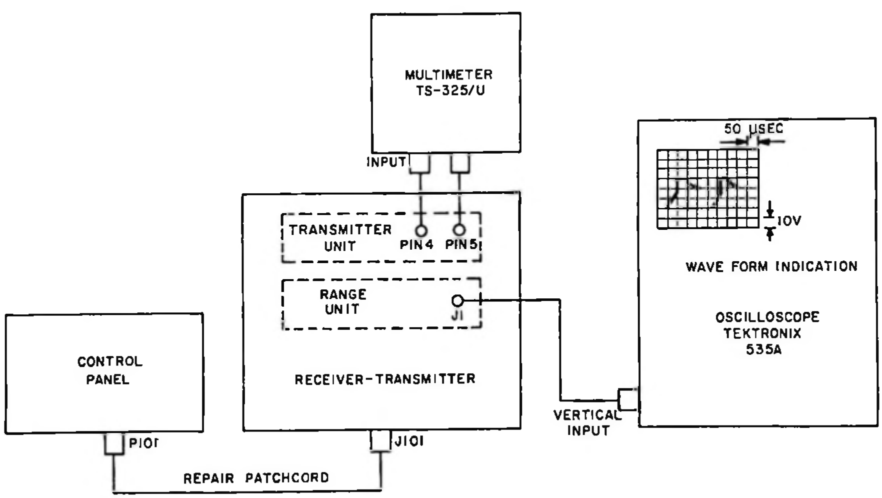


Figure 11-122F. Magnetron Current and Trigger Output Pulse Test Setup

MIT position. With a multimeter, measure the voltage between jacks J4 and J5 on the transmitter unit. The voltage should be between 0.4 and 1.7 volts dc. The multimeter used in this test procedure is a TS-325/V, or equivalent.

d. With the oscilloscope, observe the waveform present at jack J1 on the range unit. The waveform shown in figure 11-122F should be present.

e. Set the POWER switch to the OFF position. Remove the test probes from the range unit. Remove the repair patchcord between plug P101 and jack J101.

f. Replace the control panel on the receiver-transmitter center section and fasten the four trunk-type latches to secure the control panel in place.

11-809V. PULSE WIDTH, REPETITION RATE, AND POWER OUTPUT TEST. The transmitting equipment produces pulses of rf energy at fixed intervals. The rf detected pulse width (pw) is 0.2 ± 0.02 microsecond in duration and the pulse recurrence frequency (prf) is approximately 5000 pulses per second (pps). The frequency of the rf output is from 8900 mc to 9400 mc. The nominal peak-power output is 0.5 kilowatt (kw), and the average power output is 0.5 watt. To check the pulse width, repetition rate, and power output, the output magnetron pulse must be observed; refer to figure 11-122G for a block diagram of the test setup. Perform the following sequence of steps to test the pulse width, repetition rate, and power output from the transmitter:

a. Set the POWER switch on the control panel to the OFF position. Release the four trunk-type latches that fasten the control panel to the receiver-transmitter center section. Place the control panel in a non-conductive support structure.

b. Connect the repair patchcord between plug P101 on the control panel and jack J101 on the receiver-transmitter center section.

c. Release the four trunk-type latches that fasten the front section (radome) on the receiver-transmitter, and remove the radome. Within the radome is the radar set antenna. The antenna is composed of a double dipole feed assembly and a parabolic reflector. The antenna has a gain of approximately 27 db with a side lobe attenuation of approximately 19 db. The reflector has an overall diameter of 14.5 inches including the mounting structure.

d. Unscrew the cover plate from the direction coupler access hole on the receiver-transmitter housing. Unscrew the cap on the directional coupler to gain access to jack J908.

e. Connect the radar test set rf cable to

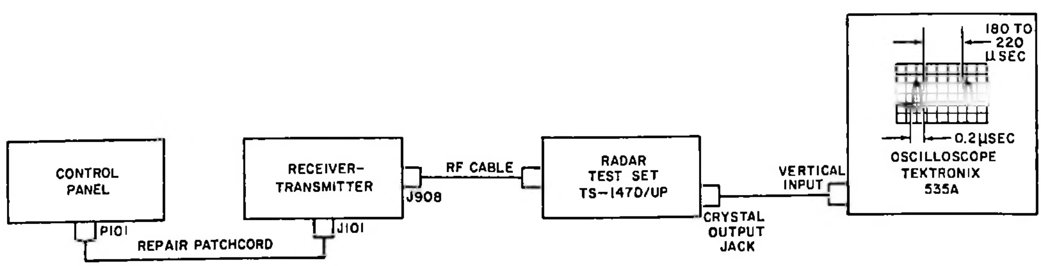


Figure 11-122G. Pulse Width, Repetition Rate, and Power Output Test Setup jack J908. Place the controls on the radar test set to the transmitter position. The TS-147D/UP Radar Test Set is used in this test procedure.

f. Connect a 50-ohm coaxial cable, terminated in its characteristic impedance, between the vertical input of the oscilloscope and the crystal output jack of the radar set.

g. Set the POWER switch on the control panel to the TRANSMIT position and observe the pulses on the oscilloscope. The oscilloscope used in this test is a Tektronix Model 535A, or equivalent.

h. Visually determine that the pulse width is $0.2 \pm 0.02\ \mu$sec at the half-power (0.707 amplitude) points, that the period between pulses on the oscilloscope is from 180 to 220 μsec, and that the power output is between 27 and 33 dbm.

i. Adjust the attenuation control until the power output is between 27 and 30 dbm. Zero the meter on the radar test set and adjust the attenuator for an indication of 1 dbm on the radar test set meter. When this meter indicates 1 dbm, the total attenuation of the direction coupler (printed next to the coupler), the radar test set rf cable, and the attenuation control is equal to the power output of the magnetron.

j. Set the POWER switch to the OFF position. Remove the 50-ohm coaxial cable and replace the cap on the directional coupler cable assembly. Replace the cover plate on the directional coupler access hole. Remove the repair patchcord from jack J101 and plug P101. Replace the control panel on the receiver-transmitter and fasten the four trunk-type latches to secure the control panel in place.

11-809W. SHORT STROBE TEST. The purpose of the short strobe is to set the time for each strobe cycle at 5 ± 1 second. To check the short strobe, the magnitude and period must be measured. Due to the simplicity of this test procedure, a block diagram will not be needed to illustrate the test setup. Perform the following sequence of steps to test the short strobe output:

a. On the control panel, set the POWER switch to the TRANSMIT position and the STROBE switch to the SHORT-RANGE position.

b. The range gate should strobe through a range of 500 ± 10 meters as indicated on RANGE EXTENSION METER M101. RANGE EXTENSION METER M101 is located on the control panel and will provide a moving target indication when the radar set receives a return signal.

c. With a watch, measure the time for each strobe cycle; it should be 5 ± 1 seconds.

d. Remove the radar set from the operational mode by placing the POWER switch in the OFF position.

11-809X. LONG STROBE TEST. The purpose of the long strobe is to increase the time interval between pulses from 5 seconds to 10 ± 0.5 seconds. Increasing the strobe time gives the radar set the capability of picking up slower-moving targets. To check the long strobe, the magnitude and period must be measured. Due to the simplicity of this test, a block diagram will not be needed to illustrate the test setup. Perform the following sequence of steps to test the long strobe output:

a. On the control panel, set the POWER switch to the TRANSMIT position and the STROBE switch to the LONG RANGE position.

b. The range gate should strobe through a range of 500 ± 10 meters as indicated on RANGE EXTENSION METER M101.

c. With a watch, measure the time for each strobe; it should be 10 ± 0.5 seconds.

d. Remove the radar set from the operational mode by placing the POWER switch in the OFF position.

11-809Y. RANGE GATE TEST. The ranging group gates the video return signals so that video from only a certain 30-meter range increment will be detected. You can select the distance, from you, of the 30-meter range increment by setting the range control handwheel located on the control panel. The output of the ranging group is obtained from peak-detecting (stretching) video signals synchronized with the range gate. This output is applied to the audio and agc unit. To check the range gate, its width, shape, and repetition rate must be observed. Refer to figure 11-122H which reveals a block diagram of the test setup. Perform the following sequence of steps to test the range gate:

a. Place the POWER switch to the OFF position. Release the four trunk-type latches that fasten the control panel to the receiver-transmitter center section, remove the control panel and place it in a non-conductive support structure.

b. Connect the repair patchcord between plug P101 on the control panel and jack J101 on the receiver-transmitter center section.

c. Remove plug P407 from jack J407 and short plug P407 to ground.

d. Connect a cable from the vertical input on the oscilloscope to jack J4 on the range unit. The oscilloscope used in this test is a Tektronix Model 535A, or equivalent.

e. On the control panel, set the POWER switch to the TRANSMIT position, the STROBE switch to the OFF position, and the RANGE METER indicator to read 3000 meters.

f. The width of the range gate observed on the oscilloscope should be 0.23 ± 0.03 microsecond at the half-power (0.707) points. The range gate should be flat within ± 0.5 volt, 0.14 microsecond minimum. The range gate repetition rate should be between 4000 and 5500 pps.

g. Set the POWER switch to the OFF position and remove the test leads. Disconnect P407 from ground and reconnect it to J407.

h. Replace the control panel on the receiver-transmitter center section and fasten the four trunk-type latches in place.

11-809Z. DETECTOR TEST. To check the detector for proper operation, the output at jack J5 on the range unit must be observed

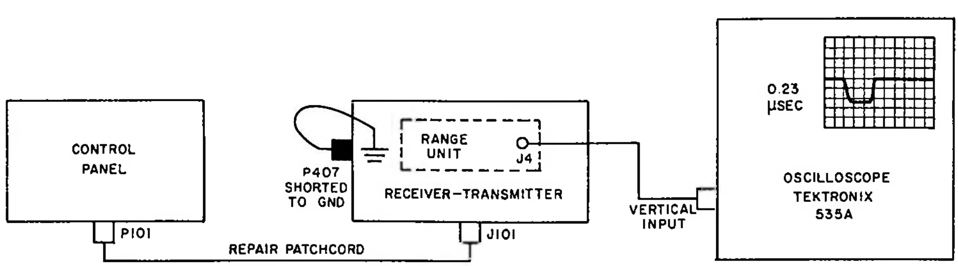


Figure 11-122H. Range Gate Test Setup

with plug P407 disconnected from jack J407 and grounded and with the output range marks being gated. Refer to figure 11-122I which reveals a block diagram of the test setup. Perform the following sequence of steps to test the detector for proper operation:

a. Place the POWER switch in the OFF position. Release the four trunk-type latches that fasten the control panel on the receiver-transmitter, remove the control panel, and place it in a non-conductive support structure.

b. Connect the repair patchcord between plug P101 on the control panel and jack J101 on the receiver-transmitter center section.

c. Set the STROBE switch to the OFF position and disconnect plug P407 from jack J407 and short plug P407 to ground.

d. Set the POWER switch to the TRANSMIT position. After 5 minutes have elapsed, observe the output at range until jack J5 with the oscilloscope. The output should be a negative-going sawtooth, not exceeding 0.15 volt in amplitude, not inclusive of spikes. The dc level at jack J5 should be a —1 volt or more negative, not inclusive of spikes.

e. Set the POWER switch to the OFF position. Remove plug P407 from ground and reconnect it to jack J407.

f. Set the POWER switch to the RANGE position and then observe the waveform at range unit jack J5 with the oscilloscope; rotate the RANGE CONTROL handwheel until a range mark is gated, as indicated on the oscilloscope. When the range mark is gated, the needle on the RANGE EXTENSION METER M101 should deflect to a minimum value.

g. Set the POWER switch to the OFF position and then remove all test cables and the repair patchcord.

h. Replace the control panel on the receiver-transmitter center section and fasten in place with the four trunk-type latches.

11-809AA. AUDIO LEVEL TEST. The audio equipment receives and amplifies the detector voltage from the ranging group. A low-pass filter in the audio amplifier circuit smoothes the detector voltage which is then amplified. The amplified audio signals are then coupled to your headset through rf filters. The rf filters prevent rf energy, generated within the radar set, from being applied to your headset leads. To check the audio level of the receiver, an audio tone must be inserted at jack J1 of the audio unit and your headsets can be used to monitor the signal at the audio output on the control panel.

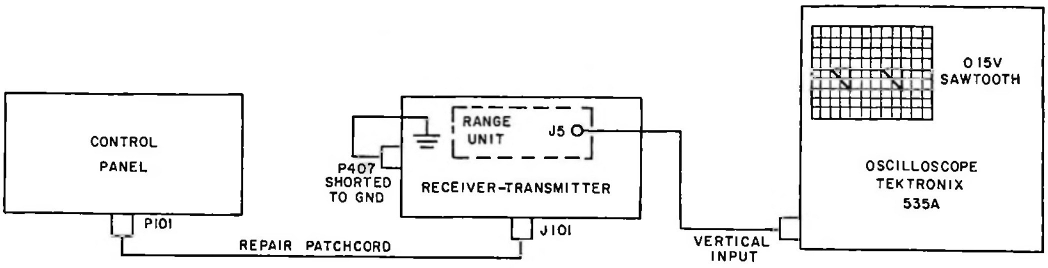


Figure 11-122I. Detector Test Setup

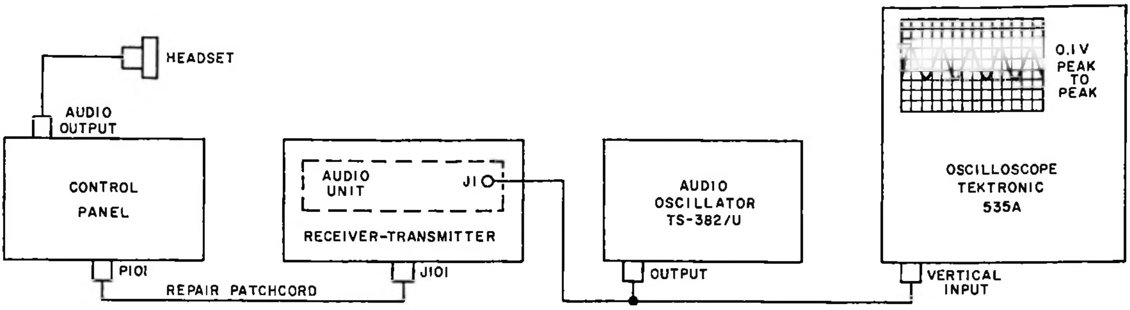


Figure 11-122J. Audio Gain Test Setup

Refer to figure 11-122J which is a block diagram of the audio level test setup. Perform the following sequence of steps to test the audio level prior to amplification:

a. Place the POWER switch to the OFF position, release the four trunk-type latches that secure the control panel to the receiver-transmitter center section, and place the control panel in a non-conductive support structure.

b. Connect the repair patchcord between jack J101 on the rear of the receiver-transmitter center section and plug P101 on the rear of the control panel.

c. Connect the power cable and the headset cable to their respective receptacles on the control panel.

d. Connect the signal output of the audio oscillator to jack J1 on the audio unit. Connect the oscilloscope across the audio oscillator output and adjust the audio oscillator to provide consecutive 80-, 400-, and 1,000-cycle signals of 0.1 volt peak-to-peak amplitude, as observed on the oscilloscope. The oscilloscope used in this test procedure is a Tektronix Model 535A or equivalent and the audio oscillator is a TS-382/U or equivalent.

e. Set the POWER switch to the TRANSMIT position and after a time delay of approximately 90 seconds, the 80-, 400-, and 1,000-cycle tones provided by the audio oscillator should be audible in your headset.

f. Set the POWER switch to the OFF position and disconnect the repair patchcord between plug P101 and jack J101. Disconnect all test leads.

g. Install the control panel on the rear of the receiver-transmitter and fasten the control panel in place by securing the four trunk-type latches.

11-809AB. RECEIVER SENSITIVITY TEST. The receiving equipment receives the target echo pulses from the antenna and produces video signals which are applied to the detector in the ranging unit. To test the receiver sensitivity, apply an rf signal from a spectrum analyzer to the directional coupler input. The video output of the receiving equipment is then displayed on an oscilloscope. Refer to figure 11-122K which illustrates a block diagram of the receiver sensitivity test setup. Perform the following sequence of steps to test the receiver sensitivity:

a. Set the POWER switch to the OFF position and then unscrew the cover plate from the directional coupler access hole on the receiver-transmitter housing. Next, unscrew the cap of the directional coupler cable assembly.

b. Connect the spectrum analyzer rf cable from the jack on the test set to J908 in the access hole. The spectrum analyzer used in this test is an AN/UPM-33 or equivalent.

c. Release the four trunk-type latches that fasten the control panel to the receiver-transmitter center section, and remove the control panel and place it in a non-conductive support fixture. Connect the repair patchcord between plug P101 on the control panel and jack J101 on the receiver-transmitter center section.

d. Connect jack J3 on the range unit to the vertical input of the oscilloscope. Connect the output of jack J1 on the range unit to the trigger input jack on the oscilloscope and synchronize the oscilloscope externally. The Tektronix Model 535A oscilloscope or an equivalent should be used in this test procedure.

e. Place the POWER switch in the TRANSMIT position. Set the spectrum analyzer for receiver operation and adjust the frequency control until targets appear on the oscilloscope.

f. Adjust the power control on the spectrum analyzer to obtain an output, with no attenuation in the spectrum analyzer, of 1 milliwatt as indicated on the meter of the spectrum analyzer.

g. Adjust the attenuators on the spectrum analyzer until the ungated target is twice the amplitude of the ungated noise as indicated on the oscilloscope. The total attenuation is equal to the sum of the indication on the attenuation meter of the spectrum analyzer plus the attenuation of the spectrum analyzer calibrated rf cable (stamped on the bracket adjacent to jack J908 on the microwave assembly). The total attenuation is also equal to the receiver sensitivity and should be 85 or more dbm. If the receiver sensitivity is less than 85 dbm, replace crystals CR901 and CR902 and repeat the measurement of the receiver sensitivity. If the sensitivity is still low, troubleshoot the i-f amplifier unit to find the trouble.

h. Set the POWER switch to the OFF position and de-energize the spectrum analyzer and oscilloscope.

i. Remove all test cables and the repair patchcord. Replace the control panel and fasten the control panel in place with the four trunk-type latches.

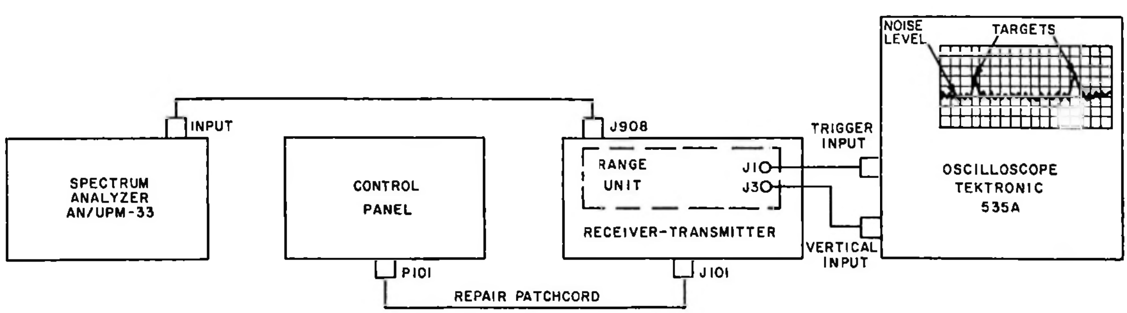


Figure 11-122K. Receiver Sensitivity Test Setup

11-809AC DOPPLER SENSITIVITY TEST. The doppler sensitivity test should be performed only when the receiver sensitivity is low; that is, the radar set does not respond well to moving targets. Refer to figure 11-122L which contains a block diagram of the test setup. Perform the following sequence of steps to test the doppler sensitivity:

a. Release the four trunk-type latches that fasten the control panel to the receiver-transmitter center section, and remove and place the control panel in a non-conductive support structure.

b. Connect the repair patchcord between plug P101 on the control panel and jack J101 on the receiver-transmitter center section.

c. Connect the test probe of an oscilloscope, Tektronix Model 535A or equivalent, to jack J3 on the range unit using a tip jack plug. Use the output at jack J1 of the range unit as an external trigger for synchronizing the oscilloscope to the radar set.

d. Turn the RANGE CONTROL handwheel until a reading of greater than 3000 meters is obtained on the range meter indicator located on the control panel.

e. Remove the oscilloscope probe from jack J3 on the range unit and ground it to establish a ground reference on the oscilloscope.

f. Return the probe to jack J3 of the range unit, and set the oscilloscope input control for ac input.

g. Connect the rf cable from the signal generator AN/UPM-11A to the directional coupler connector jack J908 and tune the signal generator to the operating frequency of the radar set until rangemarks from the signal generator appear on the oscilloscope.

h. Observe the first rangemark and make sure that it is not being gated.

i. Reduce the level of the first rangemark to approximately 3/4 of its level limit by adjusting the VOLUME control on the control panel.

j. Connect Audio Oscillator TS-382/U to the external modulator input jack on the signal generator, and adjust the audio oscillator for minimum output at a frequency of 100 cps.

k. Measure the voltage amplitude of the unmodulated first rangemark on the oscilloscope.

l. Adjust the output control of the audio oscillator until the peak-to-peak modulation on top of the rangemark is 0.1 volt. This modulation is due to the introduction of the audio oscillator signal.

m. Adjust the VOLUME control on the control panel until the noise level (baseline of the waveform) is shifted by approximately 0.3 volt.

n. Gate the first rangemark by turning the RANGE CONTROL handwheel and observing the needle of the RANGE EXTENSION METER M101 for minimum leftward deflection.

o. Listen for audio tones in the headsets while rocking the audio oscillator frequency back and forth a few cycles from the original setting.

p. Audio tones should be audible in the headsets, indicating a doppler sensitivity of 5 percent or less.

q. Repeat steps k through o using an audio oscillator frequency of 400 cps and then 1,000 cps.

r. Set the POWER switch to the OFF

position and de-energize the spectrum analyzer, audio oscillator, and oscilloscope.

s. Remove the test cables and the repair patchcord. Replace the control panel on the receiver-transmitter center section and fasten it in place with the four trunk-type latches.

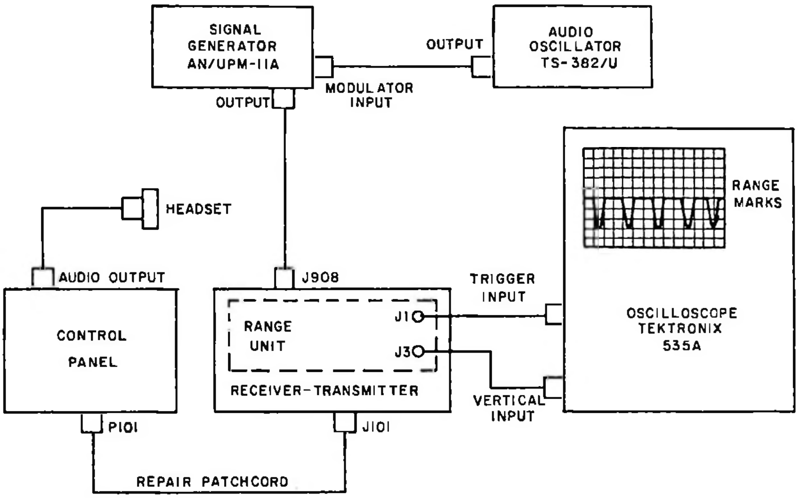


Figure 11-122L. Doppler Sensitivity Test Setup

SECTION IV

NAVIGATIONAL AIDS TESTING

11-810. GENERAL.

11-811. The importance of long range and tactical electronic navigational systems cannot be overestimated. Navigational aids make it possible for a navigator to know the position of his craft at all times. The high speeds at which modern military aircraft fly require a fast, direct method of obtaining a navigational position fix. Through the use of radio aids, the range capabilities and accuracy of observation are extended enormously beyond the possibilities of human vision.

11-812. HYPERBOLIC NAVIGATION.

11-813. GENERAL.

11-814. Hyperbolic navigation was developed during the early years of World War II as a replacement for celestial navigation. The celestial technique left too much to chance to be completely efficient for navigation for wartime use. For example, clouds could obscure the sky, so that aircraft, which need many bearings taken in a short time, because of the rapid coverage of distance, had to rely on some form of radio or radar navigation. Moreover, accurate celestial observation is hindered greatly by the motion of aircraft in flight. The hyperbolic method provides an accurate means of locating position in a shorter time than by celestial means, and is independent of aircraft flight conditions. In addition, the training required by personnel to navigate using the hyperbolic method is far less than that required for celestial navigation.

11-815. Hyperbolic methods are based upon the following premise. If two stations transmit signals at the same instant, the signals arrive at any point equidistant between them at the same time. If a number of equidistant points are connected together, they form a straight line. Now, if several points are chosen with the difference in the distance (and the difference in delays) between the stations always the same, the points when connected form a hyperbolic line. As the difference becomes greater, the curvature of the hyperbola increases. A series of hyperbolic lines, called lines of position, are plotted on a map for each pair of stations used in the hyperbolic system. By using a receiving unit which is capable of determining the difference in distance between a pair of stations in terms of time, a navigator can determine the line of position on which the aircraft is located, although the aircraft may be at any point on that line. In order to determine its exact location, a second pair of stations must be used to provide lines of position that cross those of the first pair of stations. A navigator can then determine his exact location by simply determining the two lines of position on which his aircraft is located and noting the point where these lines intersect. The point of intersection is known as a position fix.

11-816. LORAN.

11-817. The word LORAN is a contraction

of the words "LOng RAnge Navigation" and is the name applied to one of a family of facilities used in hyperbolic navigation. Loran was developed during World War II to fill the need for an accurate, long-range navigational method. By the end of the war, all theaters of operation had loran coverage for bombing missions penetrating deep into enemy territory, and all commonly used air and sea lanes were covered.

11-818. Loran differs from radar in that no transmission takes place from the using craft. Therefore, echoes are not important and the craft maintains radio silence. Loran differs from other radio direction finding equipment in that it measures the time of arrival rather than the direction or arrival of radio waves. Loran uses pulses, microsecond timing, and cathode-ray-tube techniques to form a navigation method which is capable of furnishing more reliable and accurate positioning information over greater distances than is possible by other methods of radio and radar navigation.

11-819. To accomplish this, a series of master and slave stations transmit powerful and accurately timed pulses from which the navigator, using specially designed loran receiving equipment, can obtain an accurate position fix. By measuring the difference in the time of arrival of the signals from any two of the transmitting stations, a line of position on the earth's surface is located. This is a line because there are any number of points on the earth's surface where this time difference is constant: the locus of these points is a hyperbola. A similar measurement using another pair of transmitting stations determines another similar line of position, and the intersection of these two lines determines the position of the craft. This "loran fix" is plotted on a specially designed loran chart or map in order to determine the true physical position of the craft.

11-820. In order to provide long-range navigation, the loran facility is designed for pulse transmission at a relatively low frequency. The frequencies between 1.7 and 2 mc are the optimum. These frequencies not only provide good ground-wave transmission, but since they travel skyward, they are reflected from the ionosphere and provide relatively stable sky-wave transmission. The usable range of loran during the day is 700 nautical miles over water; because of the attenuating effect of ground, the range over land is limited to 1000 miles for high-flying aircraft and to 250 miles at the earth's surface.

11-821. TYPES OF FACILITIES. In an attempt to gain longer-range navigation, two different varieties of the standard loran system have been developed. These are the SS (sky-wave-synchronized) loran and low-frequency loran, both of which operate basically in the manner of standard loran.

11-822. In the SS loran facility, the slave station of a pair is synchronized by a sky-wave pulse reflected from the E layer, rather than by the ground wave as in standard loran. This allows the master and slave stations to be separated by as much as 1000 to 1200 miles. The loran charts are calibrated in terms of sky waves, instead of ground waves, so that correction factors are unnecessary when sky waves are used. A disadvantage of this method is encountered when the indicator is located close to either or both stations, since erratic reception may result when the angle of reflection of the sky wave from the E layer approaches the critical angle. As the critical angle is approached the radio waves exhibit increasing penetrating power and may go part way or entirely through the E layer.

11-823. LOW-FREQUENCY LORAN. Low-frequency loran is an attempt at long-range navigation over land and water without the use of erratic sky waves. It is essentially the same as standard loran except for the carrier frequencies and the width of the transmitted pulse. The carrier frequency is in the 180-kc band and the pulse is wider. Ranges of 1200 nautical miles over water and 1000 nautical miles over land are not uncommon with this system. The difficult task in establishing low-frequency loran is the construction of sufficiently high transmitting antennas. In an experimental attempt, a barrage balloon was used to support a 1300-foot vertical wire; however, it is supposed that a tower approximately 600 feet high supporting a top-loaded antenna can be used.

11-824. LORAN TRANSMITTERS. Loran transmitters are generally located in coastal areas, since loran is primarily used today in transoceanic navigation. These transmitters have peak pulse power outputs of 100 kilowatts, to ensure adequate long-range coverage. Pulse transmission requires a much greater channel width than does continuous amplitude-modulated transmission; however, by assigning a different recurrence rate to each transmitter, several transmitters can use the same transmitting frequency. Each pulse transmitted by a loran transmitter is 40 μsec in duration. The pulses radiated by a pair of stations which determine a set of hyperbolic lines, must be identical with respect to shape and repetition rate. It is necessary, therefore, that careful control exist between the two stations of a pair. In each pair of transmitters, one is known as the master station and the other as the slave station. Basically, coincidence between master and slave stations is established as follows.

11-825. The master station transmits uniformly spaced pulses at a specific recurrence rate. These rf pulses are radiated in all directions at the speed of light and when they reach the slave station they are used to synchronize the recurrence rate and shape of the pulses produced by the slave-station transmitter. After a specified time delay, the slave station produces and radiates pulses which are identical to those received.

11-826. In the general discussion of hyperbolic methods and in the example used to establish hyperbolic lines of position, it was assumed that the two stations transmitted signals at the same instant. Actually in a loran facility the two stations transmit at different times. This is done to eliminate the possibility of obtaining a false line of position, which could readily be established if both stations transmitted at the same instant. After the master-station pulses are received at the slave station, a controlled amount of delay is added before the slave station transmits. The delay consists of a coding delay, which can be changed from day to day to provide security against enemy use during wartime, and a delay equal to one-half the recurrence interval of the transmitted pulses. Therefore, before the slave station transmits, these two delay periods, together known as the absolute delay, will have elapsed.

11-827. In order that the distances between the master and slave stations can be determined by the receiving unit and then be applied to the loran navigation charts, distance is defined in terms of microseconds. Radio waves require approximately 5.37 μsec to travel 1 statute mile, approximately 6.18 μsec to travel 1 nautical mile, and approximately 3.34 μsec to travel 1 kilometer. The loran charts may be calibrated in terms of statute or nautical miles, or kilometers; however, the hyperbolic lines are labeled in μsec. Figure 11-123 shows a set of hyperbolic lines plotted about a hypothetical master and slave station pair. The smallest time differential, 1000 μsec, is on the base line extension from the slave station; the greatest, 1850 μsec, is on the base line extension from the master station.

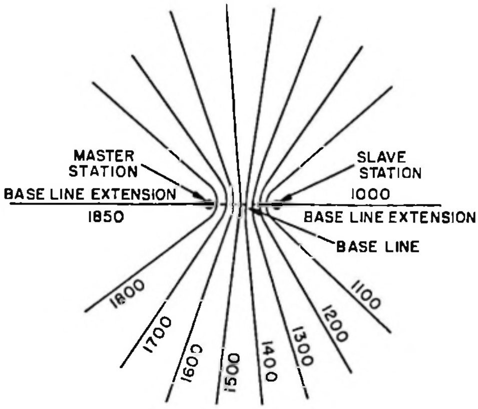


Figure 11-123. Hyperbolic Lines of Position, Showing Line Separation

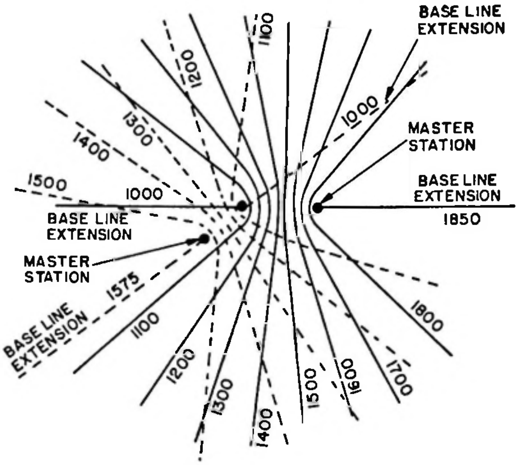


Figure 11-124. Hyperbolic Lines of Position from Two Pairs of Stations with Slave Station Common to Both Master Stations

11-828. Lines of position from two pairs of stations must be used to determine a loran fix. In order to conserve money and space, one station installation is made common to two pairs. For example, the slave station consists of two transmitters, each being synchronized by a different master station transmitter, as illustrated in figure 11-124. However, two master stations or a slave and a master may be common to one installation. In figure 11-124, the solid lines of position are the product of the master station at the right and a slave, while the dashed lines are the product of the other master and slave.

11-829. LORAN STATION PAIR IDENTIFICATION. Different channels and pulse recurrence rates are used in order that the receiver may distinguish between the pulses received from separate pairs of stations. There are four channels in the 1.70- to 2.00-mc band: channel 1 at 1.95 mc, channel 2 at 1.85 mc, channel 3 at 1.90 mc, and channel 4 at 1.75 mc. As many as 16 pairs of stations may operate in the same frequency channel and are distinguished by their recurrence rates. There are two basic rates: one at 25 pps (pulses per second), designated L (low), and the other at 33-1/3 pps, designated H (high). Each of the basic rates is divided into eight subrates. The recurrence rate designation and the channel number, together with the microsecond reading, are used to associate each line of position with a specific station pair on the loran charts. For example, a line of position corresponding to 1800 μsec produced by a station pair operating on channel 4 at a pulse recurrence rate of 33-5/9 pps is designated as "4H2-1800." A complete listing of the pulse recurrence rates together with their designations are given in table 11-6.

11-830. LORAN RECEIVER-INDICATOR. The loran receiving unit used in an aircraft or ship incorporates an indicator with a crt (cathode-ray tube) to present visually the received pulses from the transmitter station pairs. In one type of indicator, a pair of sweep lines, one placed above the other and having the same sweep time, are present on the crt. Each line is approximately one-half the recurrence rate of the transmitted pulses. Since the signal received

Table 11-6. Pulse-Recurrence Rates and Their Designations

LOW-RATE DESIGNATION	PPS	HIGH-RATE DESIGNATION	PPS
L0	25	H0	33-1/3
L1	25-1/16	H1	33-4/9
L2	25-2/16	H2	33-5/9
L3	25-3/16	H3	33-6/9
L4	25-4/16	H4	33-7/9
L5	25-5/16	H5	33-8/9
L6	25-6/16	H6	34
L7	25-7/16	H7	34-1/9

from the slave station has a fixed delay of slightly more than one-half the recurrence rate plus a small amount of coding delay, the two received signals appear on different sweep lines when the indicator is operated correctly. The correct position of the master station pulse is at the extreme left end of the upper crt trace; the slave station pulse then appears some place on the lower trace. The exact position is dependent upon the location of the two stations with respect to that of the indicator.

11-831. The loran facility has inherent errors which cannot be avoided by the operator and which are dependent upon the location of the indicator with respect to the transmitters. The closer the indicator is to the base line, as shown in figure 11-123, the more accurate the reading. The reason for this is that on the base line 1 μsec is equivalent to only a fractional part of a mile, while on the base-line extensions the time difference remains the same at all distances. Near the base-line extensions 1 μsec is equivalent to several miles, and since the accuracy of the indication is limited to approximately 2 μsec, a considerable error can exist. Expanding on this discussion, the lines of position passing between the stations are nearly parallel, but as their length is extended, the lines fan out and spread increasingly farther apart. This accounts for the difference in distance that 1 μsec represents at different locations. The readings taken near the base-line extensions therefore must be considered unreliable.

11-832. The receiver portion is a conventional superheterodyne and is designed for reception of pulse signals. The output is delivered to a crt indicator. The loran indicator scope has two horizontal traces. The top trace sweeps at a constant rate from left to right in little less than one half the pulse recurrence interval. The lower trace is swept during most of the time of the second half of the pulse recurrence interval.

Table 11-7. Pulse-Recurrence Rates

STATION RATE	S*		L		H	
	PRR	P/SEC	PRR	P/SEC	PRR	P/SEC
0	50,000	20	40,000	25	30,000	33-3/9
1	49,900	20-1/25	39,900	25-1/16	29,900	33-4/9
2	49,800	20-2/25	39,800	25-2/16	29,800	33-5/9
3	49,700	20-3/25	39,700	25-3/16	29,700	33-6/9
4	49,600	20-4/25	39,600	25-4/16	29,600	33-7/9
5	49,500	20-5/25	39,500	25-5/16	29,500	33-8/9
6	49,400	20-6/25	39,400	25-6/16	29,400	34
7	49,300	20-7/25	39,300	25-7/16	29,300	34-1/9

* The S basic rate is included in late equipment only and is not presently in operational use but is provided to allow future expansion of the Loran system.

The entire cycle is repeated at the pulse recurrence site of the transmitted signals. Thus the loran signals from the receiver appear as vertical pulses on the scope traces at points corresponding to the times of arrival of the loran signals.

11-833. The time difference is the horizontal distance from the master station signal to the slave station signal. To determine the time difference, the two signals are matched by superimposing them through the use of delay controls. In the course of this operation, the delay controls also move the "time-difference indicator," from which the time difference is read directly from a numerical, veeder-type dial. (Older receivers display timing pulses on the indicator.)

11-834. Timing is accomplished by an accurately controlled crystal oscillator and a series of divider (counter) circuits. Since loran performance depends upon the proper functioning of many interdependent electrical timing circuits, the basic and specific pulse-repetition-rate dials which provide station identification may give an erroneous identification if the dividers are not properly aligned. Loran receiving equipments are designed so that the crt indicator may be used to view test patterns which indicate whether or not the equipment is functioning accurately. A step-by-step performance procedure is contained in the technical manuals for the type loran receiver in use.

11-835. DIVIDER (COUNTER) CIRCUIT ALIGNMENT. The timing source (timer) for loran equipment is a crystal-controlled oscillator which operates continuously at a frequency of 100 kc. The entire timer (oscillator plus dividers) serves to produce output pulses which, by switching, may be adjusted according to the time intervals given in table 11-7.

11-836. The output from the 100-kc oscillator and divider chain (series of dividers) allows station identification by specific repetition rate and provides a standard for the time difference measurement. The intermediate outputs from the divider chain times the various delay circuits. The divider chain (in later type equipment) consists of five pulse divider circuits (B, C, D, E, and F dividers). The output of the B divider is controlled by the 100-kc oscillator and provides a 20 kc output. The C, D, E and F dividers have pulse outputs of 20, 000, 5000, 1000, 200, and 50 cycles per second. They provide timing pulses of 10, 50, 200, 1000, 5000, and 20, 000 microseconds, respectively. (Older equipments provide timing pulses of 10, 50, 100, 500, 1000, 5000, and 15, 000-20, 000 microseconds which are viewed and counted directly from the loran indicator for determining the time difference when making a loran position fix.)

11-837. By means of a TEST or FUNCTION switch, the outputs of the counter circuits can be viewed directly on the loran indicator. When the magnitude of the pulses, in addition to the number of pulses, is desired, an oscilloscope is connected directly to the divider circuit.

11-838. Divider circuits in all loran receivers operate on the same general principle: they are blocking oscillators arranged to give one pulse for a given number of input pulses (usually 5). The counter circuit shown in figure 11-125 is typical of a loran circuit. This circuit is a 50-microsecond counter and is similar in operation to other dividers.

11-839. The input to the divider is a series of 10-microsecond pulses from the 100-kc oscillator. The divider consists of a grid-plate coupled blocking oscillator arranged to provide one output pulse for each group of five input pulses. In common with other blocking oscillator circuits, the tube conducts when a sufficient negative charge has leaked off grid capacitor C2. When the tube conducts, the grid goes positive, causing the capacitor to recharge and finally causing the tube to be cut off. When the tube is conducting, the transformer is wound so as to impress a positive voltage on the grid through capacitor C2. Because of the large size of R4, the grid current flows into C2, causing the grid side to become negative with respect to the transformer side. As C2 charges, the grid voltage increases less and less rapidly. When the plate reaches saturation, the grid is no longer driven positive. The negative charge accumulated on the grid instantly reduces the plate current, thereby causing a collapse of the magnetic field of the transformer. This results in a negative voltage to be applied to the grid, thereby driving the tube beyond cutoff.

11-840. The cycle of operation is complete; the tube cannot conduct until most of the charge of C2 has leaked off through R4 to a voltage determined by R1, R4 and R3. The time constant of C2 and R4 has been selected so that the fifth input pulse causes the tube to conduct during the steep portion of the grid-voltage waveform (figure 11-125). This is actually the discharge curve of R2-R4, thus triggering the divider on the fifth rather than the fourth or sixth input pulse.

11-841. Since the time constant of R4-C2 is critical, R2 is variable so that the divider may be set for optimum stability. The normal procedure for aligning the divider circuits is to view the grid waveform, either on the loran indicator or with an oscilloscope, and adjust R2 (or its counterpart) for the

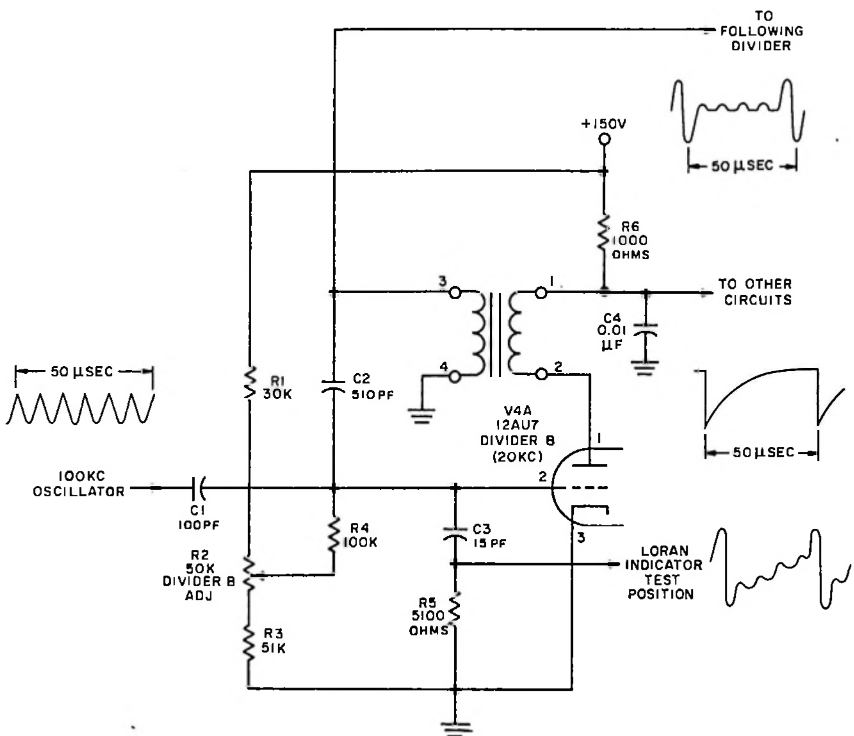


Figure 11-125. 50-Microsecond Counter Circuit (B Divider), Schematic Diagram

proper number of pulses using the following method.

11-842. If the pattern is incorrect, turn DIVIDER control R2 fully clockwise, then turn it counterclockwise until a point is found where four short pulses are between each pair of long ones and the pattern is stable. Continue counterclockwise rotation until the pattern becomes unstable; note this position on the dial of the control. Next, turn the control clockwise through the stable position until the patern again becomes unstable and note the number on the dial. Set the control on the number in the center of the two unstable readings. This procedure is repeated sequentially for each divider, starting with the 50-microsecond divider. Proper waveforms for each divider are included in the technical manuals for each equipment.

11-843. USE OF PULSE GENERATOR. Special pulse generators are supplied to provide simulated transmitted pulses for use in checking loran receivers. The test set is designed for use in performing the following tests on loran receivers:

a. Receiver alignment and sensitivity.

b. Timer performance and crystal oscillator frequency on all channels.

c. Stability of sweep generating circuits and accuracy of time delay measurements.

11-844. The equipment has a calibrated output for use in making sensitivity checks and to provide a series of calibrated pulses, at the standard pulse repetition rates of 20(S), 25(L), or 33-1/3(H), to the receiver. By selecting S, L, or H pulse repetition rates, a series of pulses at fixed delays (which are indicated on the equipment for the repetition rate selected) are made to appear on the loran traces. By taking a delay reading of the simulated pulses, the accuracy of the loran time difference reading is determined by a direct comparison with the delays introduced by the generator.

11-845. RECEIVER BANDWIDTH TESTING. A test procedure typical of loran receivers using an unmodulated signal generator is given below: the bandwidth at the 6 and 60 db down points may vary for the type loran receiver being tested and should be checked in the related technical manual.

a. Set channel switch to 1, INTERFERENCE REDUCER to OUT, and AMPLITUDE BALANCE to center.

b. Connect a signal generator to the receiver antenna input connector. The receiver input impedance is 50 ohms; the signal generator output should be matched to this value. Connect a dc vacuum-tube voltmeter across the diode (detector) load resistor.

c. Set the signal generator to 1950 kc and adjust its output to apply a 5-microvolt unmodulated signal across the receiver antenna terminals.

NOTE

If the signal generator output impedance is not 50 ohms and an external matching network must be used between the signal generator and the receiver, 5 microvolts at the signal generator output terminals will not be 5 microvolts at the receiver input.

d. Adjust the RECEIVER GAIN control to obtain 2.5 volts across diode load resistor.

e. Double the signal voltage input to the antenna terminals. Tune the signal generator to each side of resonance and note the frequencies at which the receiver output is again 2.5 volts. The difference between the lower frequencies gives the receiver bandwidth 5 db down from response at resonance. This bandwidth should be not less than 35 kc and should be centered about the correct channel frequency.

f. Increase the signal voltage input to the antenna terminals to 10,000 microvolts. Tune the signal generator to either side of resonance and again note the two frequencies at which the receiver output is 2.5 volts. This gives the bandwidth 60 db down. It should not exceed 180 kc and should be centered about the correct channel frequency.

g. Repeat the process on each of the three loran channels, if you desire. As an alternate method, a quick general check of proper bandwidth and alignment may be made with the loran test set which produces simulated loran signals; the specific procedures are included in the technical manual.

11-846. TACAN EQUIPMENT.

11-847. GENERAL.

11-848. OPERATION. Tacan, which is derived from TACtical Air Navigation, is a radio air navigation facility of the polar-coordinate type. It provides a properly equipped aircraft with bearing and distance from a ground beacon selected by the pilot. This facility incorporates 126 two-way operating channels spaced 1 megacycle apart within the band of 1025 to 1150 megacycles for air-to-ground transmission. A like number of channels for ground-to-air transmission are provided in the bands 962 to 1024 megacycles and 1151 to 1213 megacycles.

11-849. The distance from the ground beacon is visually displayed in the aircraft on a meter calibrated in nautical miles. Another meter indicates direction in degrees with respect to magnetic north. With this information, the pilot can fix his position at all times. A single tacan beacon can provide distance information simultaneously to as many as 100 aircraft and bearing information to an unlimited number of aircraft within a line-of-sight range of 200 nautical miles.

11-850. The operating principles of tacan are based on the time required for a radio pulse signal to travel to a given point and return. In distance operation, the airborne transmitter repeatedly sends out interrogation pulses that are picked up by the ground beacon receiver, which in turn triggers the associated transmitter to send out reply pulses on a different frequency. The reply pulses are picked up by the airborne receiver. Timing circuits automatically measure the interval between the interrogation and reply pulses and convert the time interval into electrical signals that operate the distance indicator. Bearing information is supplied by amplitude modulating the train of pulses generated by the transmitter; modulation is generated by the antenna equipment.

11-851. The antenna consists of a vertically polarized stationary array around which two sets of parasitic elements, contained in cylinders, rotate in a clockwise direction. The inner cylinder contains one parasitic element and modulates the pulse amplitude of the receiver at 15 cycles per second. The other cylinder contains nine elements and superimposes a 135 cps modulation of the fundamental 15 cps modulated signal. Accordingly, as a result of the rotation of this pattern, the signal received at any given direction from the beacon goes through corresponding cyclic variations in field strength as a function of time.

11-852. When demodulated, the cyclic variations produce components of 15 cps and 135 cps in fixed phase relation to each other in the airborne equipment. Bearing information is derived from the electrical phase of these audio signals with respect to reference pulses.

11-853. The tacan radio beacon, with its associated antenna groups and accessories and aircraft radio set, together provides means through which the position of an aircraft can be accurately determined. As many as 100 aircraft may simultaneously obtain navigational information from a single tacan installation. The tacan beacon is capable of receiving on any one of 126 frequencies in the range of 1025 to 1150 megacycles. The set can transmit on any one of 126 frequencies in the ranges of 962 to 1024 megacycles and 1151 to 1213 megacycles. Two types of antennas are available for use. Each antenna can operate on 63 channels, either in the low band of frequencies or in the high band of frequencies. Low band installations transmit at frequencies between 962 and 1024 megacycles inclusive, and receive at frequencies between 1025 and 1087 megacycles, inclusive. High band installations receive in the range of 1088 to 1150 megacycles and transmit in the range of 1151 to 1213 megacycles. Two frequencies are used in each channel, one for receiving and the other for transmitting. The frequency used for receiving in low band installations is 63 megacycles above the frequency used for transmitting in the same channel. In high band installations the receiving frequency is 63 megacycles below the transmitting frequency. The radio beacon output consists of the beacon's identification call, distance information signals, bearing information signals, and random pulses used to make up the constant duty cycle.

11-854. The radio beacon periodically transmits its identifying call in International Morse Code, thus enabling the aircraft to determine with which radio beacon it is in contact. The characters of the code consist of a train of pulse pairs generated at a fixed rate of 1350 cycles per second. A mechanical keyer accomplishes the coding. The pulses are phase-locked with the antenna rotation. The aircraft receiver detects these regularly occurring pulse pairs and reproduces the code as a keyed 1350-cps audio tone. Identification call signals have priority over the distance information signals. Bearing information reference bursts (described below) have priority over the identification call. The relative durations of these signals are such, however, that effectively there is no interruption of distance information or identification call.

11-855. The aircraft radio set transmits distance interrogation signals and receives replies from the radio beacon. The aircraft radio set sends out on the assigned channel pairs of pulses. The pulses of each pair are spaced 12 microseconds apart. The radio beacon receives these pulse pairs, termed distance interrogation pulse pairs, and in reply transmits back to the aircraft pulse pairs on the assigned channel. The time delay of the distance interrogation pulse pair to the corresponding distance reply pulse pair in the radio beacon is adjusted to exactly 50 microseconds, and the aircraft radio set deducts 50 microseconds from the total time elapsed between interrogation and reply. The distance between the aircraft and the radio beacon is thus determined by measuring the total time elapsed between initial transmission of the distance interrogation pulse pair and reception of the corresponding radio beacon reply.

11-856. Since all aircraft using the radio beacon operate on the same pair of receiving and transmitting frequencies, all radio beacon replies to all aircraft distance interrogations are received by all of these aircraft. It is therefore necessary for each aircraft to select those radio beacon replies which result from its own distance interrogations. This is done as follows.

11-857. Transmission of aircraft distance interrogation pulse pairs is continuous, and, in turn, aircraft reception of distance replies is also continuous. Transmission of the distance interrogation pulse pairs is semi-random; that is, the number of pulse pairs per second transmitted by a particular aircraft remains fairly constant, but the intervals between pulse pairs vary. The variation in time spacing is peculiar to each aircraft, and permits the aircraft to pick out the replies to its particular interrogations. The time-spacing pattern of reply pulse pairs is continuously compared with the time-spacing pattern of interrogation pulse pairs. Only those pulse pairs which lie in matching patterns operate the distance meter of the aircraft radio set. This meter displays continuously the distance of the aircraft from the radio beacon.

11-858. Bearing information originates within the radio beacon. The radio beacon antenna modulates the total pulse-pair output with subaudio frequency components, and the radio beacon transmits special bursts of pulse pairs which are inserted in the radio beacon output at appropriate intervals. The bearing of the radio beacon relative to the aircraft, as measured between a line connecting the aircraft to the radio beacon and a magnetic meridian, is directly proportional to the occurrence of a special burst of pulse pairs and the phase of the subaudio frequency components modulated on all the pulse pairs. In determining its bearing the aircraft utilizes 15-cps and 135-cps modulation on the total received signal, and 15-cps and 135-cps reference bursts which are components of the received signal.

11-859. Figure 11-126 represents a simplified version of the relative time occurrence of the components of the signal at the radio beacon. The 15-cps and 135-cps reference bursts are actually a train of rf pulse pairs. The 15-cps burst consists of 12 pulse pairs spaced 30 microseconds apart. The 135-cps burst consists of six pulse pairs spaced 24 microseconds apart. It is the number and spacing of the pulse pairs which identify each of the bursts. The aircraft radio set first reconstructs the 15-cycle and 135-cycle components from the envelope of the received pulse pairs. The 15-cps and 135-cps bursts serve as marker signals. The aircraft radio set compares the time occurrence of the 15-cps burst with the phase of the 15-cycle component, and the time occurrence of the 135-cps burst with the time occurrence of the 135-cps component. One cycle of the 15-cps wave represents 360 degrees of bearing. One cycle of the 135-cps wave represents 40 degrees of bearing. The beginning of the cycle on the 15-cps wave represents magnetic north. Thus, if the 15-cps reference burst occurs when the 15-cps wave is one third of the way to its first positive peak, and the first 135-cps reference burst occurs when the 135-cps component is at its first negative peak, the bearing is established as being 330 degrees to the beacon.

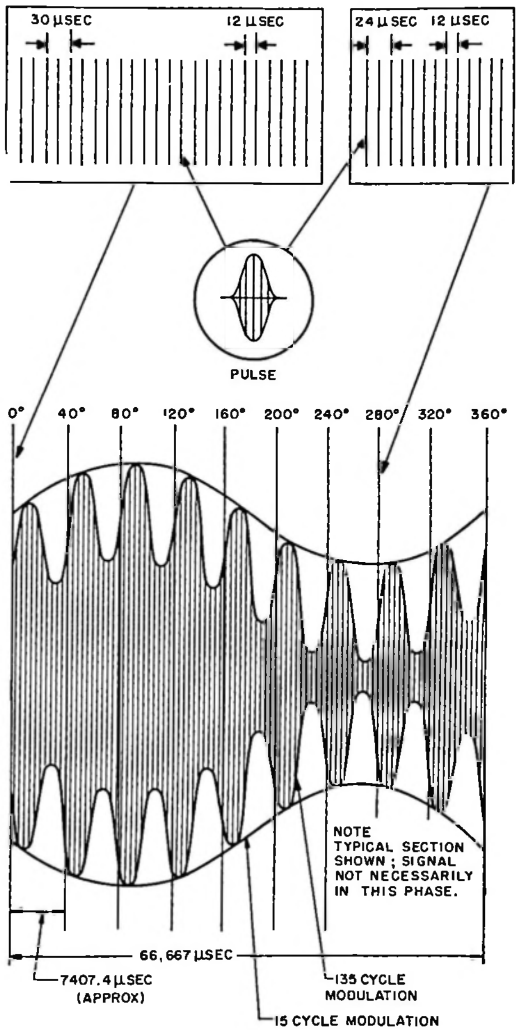


Figure 11-126. Relative Occurrence of Tacan Radio Beacon Signals

11-860. The total beacon output consists of rf pulse pairs transmitted at a rate of 3600 pulse pairs per second, 900 pulse pairs of which are 15-cps and 135-cps reference bursts. The additional 2700 pulse pairs per second which are required to keep the rate constant are either distance reply pulse pairs or random noise pulse pairs. The radio beacon generates these lest the number of replies become insufficient to maintain the required rate, since the number of replies is dependent on the number of interrogations received by the radio beacon. Should the number of interrogations exceed 2700 pulse pairs per second, the radio beacon will reply only to that number, thus maintaining the constant output pulse-pair rate. Identification call pulse pairs are substituted for distance replies or random noise pulse pairs, and consequently do not affect the output rate. A constant output rate is necessary in order that the aircraft may extract the 15-cps and 135-cps modulation.

11-861. TEST EQUIPMENT Because of the importance and complexity of the tacan beacon, built-in test equipments are included in the console which greatly facilitate performance testing and trouble isolation of the unit. The test equipment includes a pulse analyzer, pulse counter, pulse sweep generator and an oscilloscope. Specialized and simplified procedures are included in technical manuals for performing tests such as receiver sensitivity, transmitter power output, and spectrum analysis. Since the reference bursts, coding, and other pulses and their spacings are highly critical, frequent checking of these waveforms is mandatory for optimum performance of this equipment.

11-862. TROUBLE DETECTION. Some of the trouble encountered with the beacon is first detected by the aircraft and reported to the radio beacon location. Defects in the antenna output radiation, as indicated by failure of the aircraft to receive correct bearing information, distance information replies, or identification codes, may have their origin in units of the radio beacon other than the antennas. Ordinarily, trouble does not originate in the antenna unless it has suffered mechanical damage or has been dismantled for some reason. Because of the complex nature of the antenna group, the following tests can be performed only at a depot facility.

a. Measurement of the relative phase of the 15- and 135-cycle components.

b. North calibration of the antenna.

c. Harmonic analysis.

d. Measurement of percentage modulation and sideband tracking in the vertical plane.

e. Vertical pattern measurement.

11-863. ALIGNMENT AND MAINTENANCE.

11-864. CODE INDICATOR REFERENCE BURST. For each revolution of the antenna the 135-cps reference burst generator receives eight trigger pulses from the antenna, and the 15-cps reference burst generator receives one 15-cps trigger pulse. The 135-cps trigger pulses applied to the 135-cps reference burst generator are approximately 12 volts peak-to-peak and 150 microseconds in duration. Upon reception of the 135-cps trigger pulses, the reference burst generator produces six pulses spaced 24 microseconds apart.

11-865. The reference pulses are generated in a similar manner in the 15-cps reference burst generator. The output of the 15-cps reference burst consists of 12 pulses, 30 microseconds apart. At the input to the shaping amplifier stage the 15-cps and 135-cps reference pulse groups are mixed. From this point on, both pulse groups pass through the same stages. This action is possible because the two reference pulse groups never occur simultaneously.

11-866. From each reference burst sent to the shaping amplifiers, a blocking gate is sent to the priority gate. This gate blocks out the identification call and reply signals during transmission of reference bursts to establish the priority of reference bursts over all other signals.

11-867. CODE INDICATOR TONE KEYER ADJUSTMENT. To perform the tone keyer adjustment, you must accomplish the following sequence of operations:

a. Upon gaining access to the keying wheel, energize the coder-indicator unit.

b. Turn the keyer switch to the ON position.

c. Using a stopwatch, determine the time required for the necessary revolutions of the coding wheel. The coding wheel should complete the necessary revolutions within a certain number of seconds as determined by the specific equipment in operation.

d. Connect an indicator to the test output jack and interpret the code produced on the indicator. If no code is sensed, adjust the coding assembly; to do this, move the adjusting control in or out as required. Code should be transmitted for a required number of revolutions of the code wheel. The complete code should be sensed with no omission or repetition of code characters.

e. If the above requirements are not met, proceed as follows to adjust the code cutoff control. (If the requirements are met, omit this step.) Adjust the cam assembly until code just starts at the start code segment. The cam shaft should be moved so that the code cutoff control is not actuated at the start code segment.

f. Repeat the procedures called for in step e with the motor reconnected.

11-868. The code on the cam wheel is changed by loosening the screws in the segments, pulling out or pushing back the necessary cam segments, and then tightening the screws. Always set up the code starting with the cam segment start C code. It is desirable to let three segments, including the start code, remain pushed back before the code is actually started. Therefore, unless the code uses more than 50 segments, the first three segments should not be used, and the code should be set up starting with the fourth segment. A dot consists of a single pulled-out segment. A dash consists of three pulled-out segments. To separate dots and dashes, one segment is pushed back between a dot and a dash, two dots, or two dashes. Three segments are pushed back between characters of the code.

11-869. For example, the code USA is inserted in the manner shown in figure 11-126A. The segment marked START CODE and the next two segments should be pushed back. The fourth segment should be pulled out. This comprises the first dot of the character U. The fifth segment should be pushed back, the sixth pulled out (second dot of character U), the seventh pushed back, and the next three segments pulled out. This completes the character U, "dot-dot-dash". After completing the character U, three segments are pushed back, to separate the U from the following S. For the character S, three alternate segments are pulled out, forming "dot-dot-dot". After this character is completed, three segments are again pushed back, to separate the S from the following A. The character A is formed by pulling out the next segment, pushing back the following one, and finally pulling out the three following segments, thereby completing the "dot-dash" of the A. The complete code, USA, has now been set up. All other segments should be pushed back.

11-870. Replace the coding wheel by engaging the locating pins on the wheel hub, using caution during this procedure to avoid damage to the coding control. Then replace other necessary parts in the reverse order of disassembly. During the interval of time that the code changes are being made, as outlined in the previous paragraph, no identification is transmitted from the radio beacon. However, bearing and distance information may continue uninterrupted during this interval.

11-871. The beacon time delay may be defined as the time lag between the reception of the radio-frequency pulse and the transmission of the first radio frequency reply pulse by the beacon antennas.

11-872. DELAY LINE CHECK AND ADJUSTMENT TO OBTAIN OVER-ALL RADIO BEACON ZERO DISTANCE DELAY. The

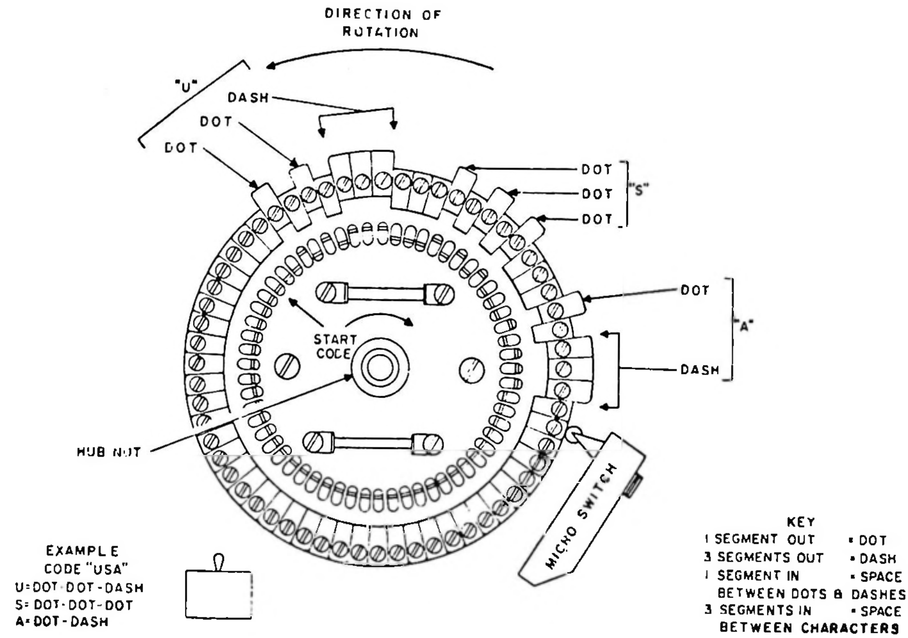


Figure 11-126A. Setting the Code

following procedure is based on checking the over-all radio beacon delay of 50 microseconds and adjusting the delay line as necessary to obtain that delay. This method is employed to permit you to use the built-in test equipment:

a. Set the function switches on an oscilloscope and the power meter pulse counter to VIDEO DELAY.

b. Set the pulse-sweep generator TRIGGER SELECTOR switch to the SWEEP position, and the PULSE CODING and CRYSTAL SELECTOR switches to the desired positions.

c. Adjust the pulse analyzer-signal generator to provide a modulated signal. Then, set the MODULATION SELECTOR control to PULSE and the RF OUTPUT to the desired decibel level.

d. Observe the display on the oscilloscope screen, and if necessary, adjust the controls of the oscilloscope for a clear presentation. Note that there are two sets of pulses displayed—the radio set output pulse and a pair of reference pulses.

e. Adjust the pulse-sweep generator BALANCE control so that the reference pulses have slightly larger amplitude than the radio set output pulses. A pattern similar to that shown in figure 11-126B should be obtained on the oscilloscope.

f. If a pattern such as that shown in part A of figure 11-126B is obtained, that is, with the radio set output pulse-pair to the left of the reference pulse-pair, the over-all zero-distance delay is less than 50.2 microseconds. In that case, the next step should be performed. If it is to the right as

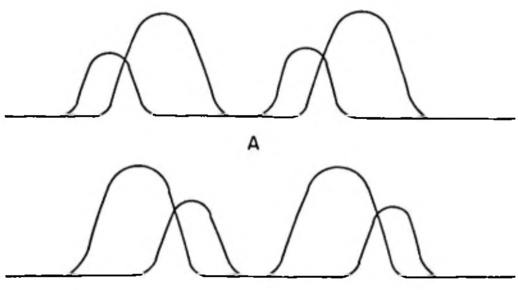


Figure 11-126B. Delay Measurement Waveforms

shown in part B, the zero delay is too great, and it is definitely established that adjustment of the delay line is necessary. The procedure for doing that is provided by step h.

g. Change the CRYSTAL SELECTOR control on the pulse-sweep generator to the desired position. Observe that the pattern is as shown in figure 11-126B. If the radio set output pulse-pair is to the left of the reference pulse-pair, the zero-distance delay is less than 49.8 microseconds. In that case, a readjustment of the delay line is necessary.

h. Upon gaining access to the delay line, make necessary adjustments to obtain a net delay of between 49.8 and 50.2 microseconds, as determined by repeating steps f and g.

11-873. MAGNETIC VARIATION ADJUSTMENT FOR NORTH CALIBRATION OF ANTENNA. Check the reading of the magnetic variation dial to see that the correct compensation for magnetic north has been set. At shore installations, the servo motor control in the antenna base, must be loosened before compensation for magnetic north can be set. At fixed shore installations, the antenna base and antenna assembly have been lined up with true north. During initial adjustments it is necessary to compensate for the angular difference between magnetic north and true north, as determined from charts for the particular geographic area. Once this compensation has been set, further adjustments will not be necessary during normal operations.

11-874. At shipboard installations, the antenna base and antenna assembly have been lined up with the fore-and-aft direction of the ship. Depending upon the geographic location, magnetic north and true north must be compensated for. This is necessary because the ship's compass is calibrated to true north, whereas the 15-cps, or north reference for the aircraft, is oriented to magnetic north. Once the adjustment has been made, the bearing of the radio beacon to magnetic north will be automatically maintained by the gyrocompass of the ship within geographic limits. The bearing information transmitted to the aircraft will remain accurate as long as the ship remains in areas within which the angular difference between true north and magnetic north remains approximately the same as that for which the magnetic variation unit has been set. Readjustments for magnetic variation must be made whenever the angular difference changes appreciably, as determined from navigational charts and maps.

11-875. To make the adjustments, proceed as follows:

a. Set the magnetic variation subassembly until the main control is free.

b. For variations west of true north set control toward increasing numbers, that is, from 0 degrees up to 10 degrees, 20 degrees, etc. For variations to the east of true north, set control from 0 degrees (360 degrees) down, that is, toward 350 degrees, 340 degrees, etc.

11-876. ANTENNA GROUPS FUNCTIONS. The antenna group of the radio beacon performs the following functions:

a. Receives distance interrogation pulse pairs from the aircraft.

b. Radiates the radio beacon output signals, which the aircraft receive and process to obtain distance and bearing information.

c. Provides bearing reference pulses to trigger the coder-indicator at specific intervals.

11-877. To detect a reversal in phase of an ac signal, it is necessary to compare it with a fixed reference. A circuit which does this is shown in figure 11-126C. The circuit operates as follows: Assume that no error voltage is present, and that an ac reference voltage is applied through T_1. When point C is positive with respect to point F, points A and B are also positive with respect to point F, and diodes 1 and 2 conduct equally. This condition results in equal voltage drops across R_1 and R_2, and the voltage from point D to point E is zero. When point C is negative, no current will flow because of the polarity of the diodes.

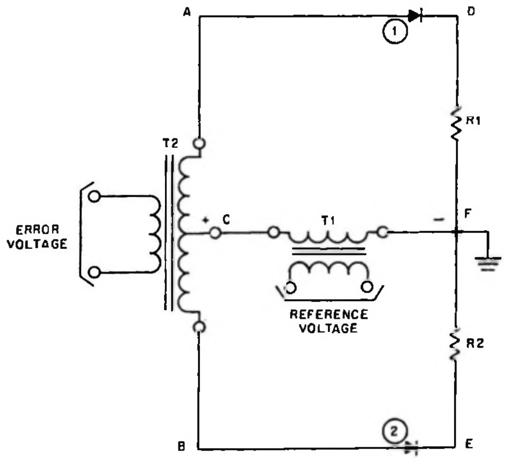


Figure 11-126C. Simple Phase Detector

11-878. Assume that an error voltage of the same frequency as the reference voltage is now applied across transformer T_2 in phase with the reference voltage. The error voltage is now added to the reference voltage. When point C is positive, the voltage in the secondary of T_2 from point C to point A adds to the voltage at point C, causing point A to become more positive and the current flow through diode 1 to increase. At the same time the voltage across the secondary of T_2 from point C to point B is negative and, therefore, subtracts from the voltage at point C. This action causes point B to become less positive and the current flow through diode 2 to decrease. When the current through diode 1 increases, the voltage across DF exceeds the voltage across EF. Since the voltage drop across R_1 is greater than the drop across R_2, point D is positive with respect to point E.

11-879. If the error voltage is out of phase with the reference voltage, causing point B to become positive with respect to point A, the voltage drops across R_1 and R_2 will be reversed, as will the polarity across DE. The polarity of voltage DE changes with the polarity of the error signal, and the voltage is proportional in magnitude to the error voltage.

11-880. One other important feature of this phase detector is that it acts as a limiter if the amplitude of the error voltage equals or exceeds the reference voltage. This condition occurs since the resultant voltage of path FCB, as shown in figure 11-126C, will keep decreasing until the reference equals the error. Thereafter this voltage will increase again with the opposite phase. However, since the diodes are insensitive to phase and are affected only by the magnitude of the voltage across them, the dc rectified current no longer decreases for any further increase in error voltage. The difference between the resultants of paths FCA and FCB remains constant and is the net error voltage. By changing the magnitude of the reference voltage, it is possible to control the level at which the output dc from the phase detector saturates. A curve of output dc as a function of ac error voltage for various values of reference voltage is shown in figure 11-126E. One important disadvantage of the phase detector shown in figure 11-126C is that the fundamental power frequency appears across the load since the output current is half-wave rectified. This can be a very serious difficulty in magnetic amplifier servo applications.

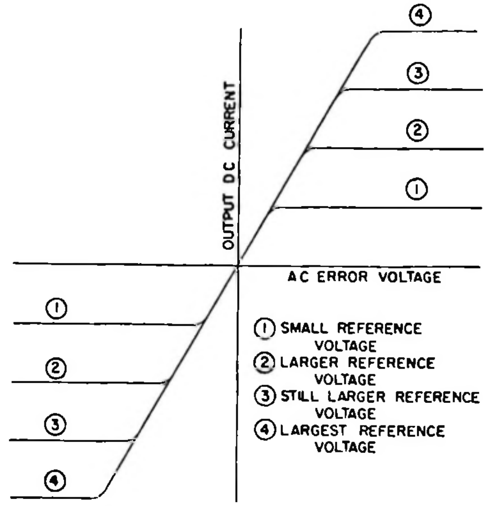


Figure 11-126E. Comparison of Input AC Error with DC Output, Graphic Analyzer

11-881. A phase detector which eliminates the fundamental power frequency is shown in figure 11-126F. Two bridge rectifiers are connected across the reference voltage with current-limiting and balancing resistors in series. One of these bridge rectifiers is illustrated separately in figure 11-126D, while both rectifiers are shown in figure 11-126F. Bridge rectifier CR_1 is oriented across the reference voltage so that all four diodes conduct during the positive half-cycle, and rectifier CR_2 has full conduction during the negative half-cycle. Thus each rectifier has a conduction path between terminals 1 and 2 during successive half-cycles. If the error voltage has a phase with respect to the reference voltage, as shown in figure 11-126G, then conduction will take place for the positive half-cycle as shown in part A of figure 11-126G and for the negative half-cycle as shown in part B of the same figure. In both cases the current through the load is in the same direction; that is, the error voltage has been full-wave rectified. In addition, if the phase between the error and reference voltages should reverse, a similar set of conduction path diagrams will be obtained, except that the polarity of the dc voltage across the load will also reverse.

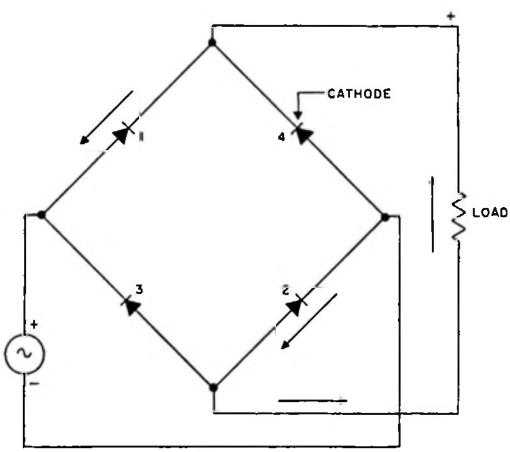


Figure 11-126D. Bridge Rectifier Circuit, Showing Direction of Load Current During Positive Half-Cycle

11-882. The significant features of a phase detector is that the output is a direct-current voltage proportional in magnitude to the

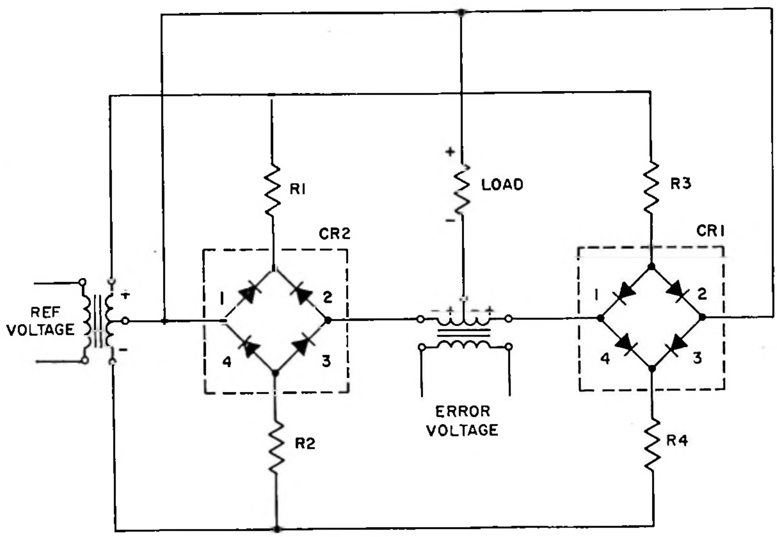


Figure 11-126F. Phase Detector with Bridge Circuits

ac input signal and of a polarity depending on the phase of the ac error voltage fed to it. To accomplish a synchro alignment, perform the following procedure:

a. Connect the synchro control transformer to a source of approximately 78 volts, as shown in figure 11-126H.

b. Obtain a minimum reading on the voltmeter, thereby indicating the approximate electrical zero position. Disconnect the 78 volts from S_1 and S_3.

c. Connect a jumper across terminals S_1 and S_3, applying a source of 78 volts across terminals S_1 and S_2 as illustrated in the figure.

d. Connect a voltmeter across terminals R1 and R2, and adjust the stator for minimum reading on the voltmeter to obtain the accurate zero position.

e. Lock the synchro in position.

11-883. RECEIVER SQUITTER CIRCUIT ADJUSTMENTS. Variation of the squitter rate adjustment may result in an extremely high squitter rate and thereby cause high klystron beam current to trip the overload relay of the transmitter. The following procedure is for squitter count and adjustment using the built-in test equipment.

a. Set the meter selector controls to the desired positions. The meter should indicate about half-scale deflection in each position.

b. Turn the function switch on the power meter pulse counter to receiver sensitivity.

c. Set the counter selector switch on the power meter pulse counter to the squitter position.

d. Set the range switch on the power meter pulse counter to the desired position.

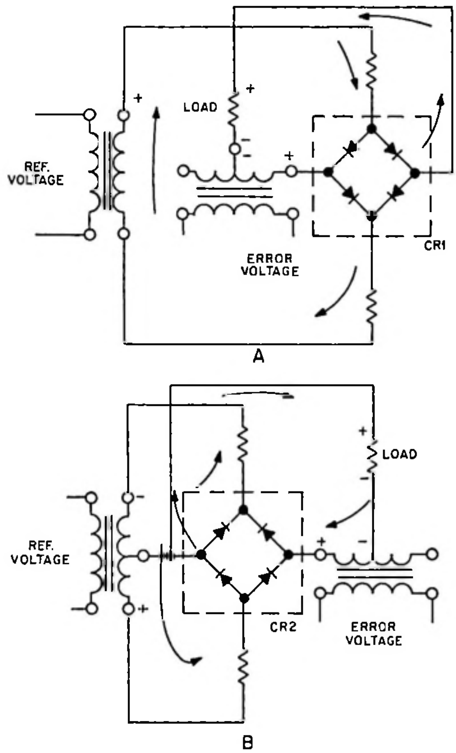


Figure 11-126G. Current Flow Through Phase Detectors

The squitter rate, as read on the pulse count meter, should be a certain number of pulses per second as specified for the equipment (e.g., 2700 $\pm$90 pulses per second). If the number of pulses per second is not correct, adjust the equipment controls to provide the number of pulses per second required.

11-884. Use an oscilloscope to check the waveform at the blanking gate test point in the radio receiver. The waveform should be similar to that shown in figure 11-126I. Measure the waveform for pulse width, and then adjust the set to the desired microseconds needed for proper operation. Readjust the squitter count as previously discussed.

11-885. TRANSMITTER BEAM PULSE ADJUSTMENT FOR DROOP ELIMINATION. Connect an oscilloscope, with the FUNCTION switch in the general test position, to the antenna incident jack and observe the north reference burst. Compare the amplitude of the first pulse in the reference burst pulse train with the amplitude of the last pulse in the pulse train. Adjust the set until all pulses in the reference burst pulse train are approximately equal in amplitude.

11-886. TRANSMITTER ADJUSTMENT OF BEAM PULSE WIDTH Connect an oscilloscope to the beam pulse test jack with the function switch set in the desired position. Set the marker selector switch on the oscilloscope to the 1-microsecond position, to display 1-microsecond markers on the scope screen. Measure the width of the beam pulse appearing on the scope screen by counting the number of 1-microsecond markers superimposed on the beam pulse from edge to edge. The beam pulse should be a certain number of microseconds in width at the half-amplitude points, as specified for the equipment. If these requirements are not met, the frequency multiplier-oscillator must be adjusted to obtain the proper pulse width.

11-887. Connect the oscilloscope to the shaped-pulse test jack of the frequency multiplier-oscillator, and observe the shaped pulse. Center the shaped pulse about a vertical hairline on the scale. Note carefully the positions of the half-amplitude points of the shaped pulse. Once these points have been determined, do not touch any of the horizontal controls of the oscilloscope. Connect the oscilloscope to the beam pulse test jack, and observe the position of the beam pulse relative to the two points previously noted for the half-amplitude points of the

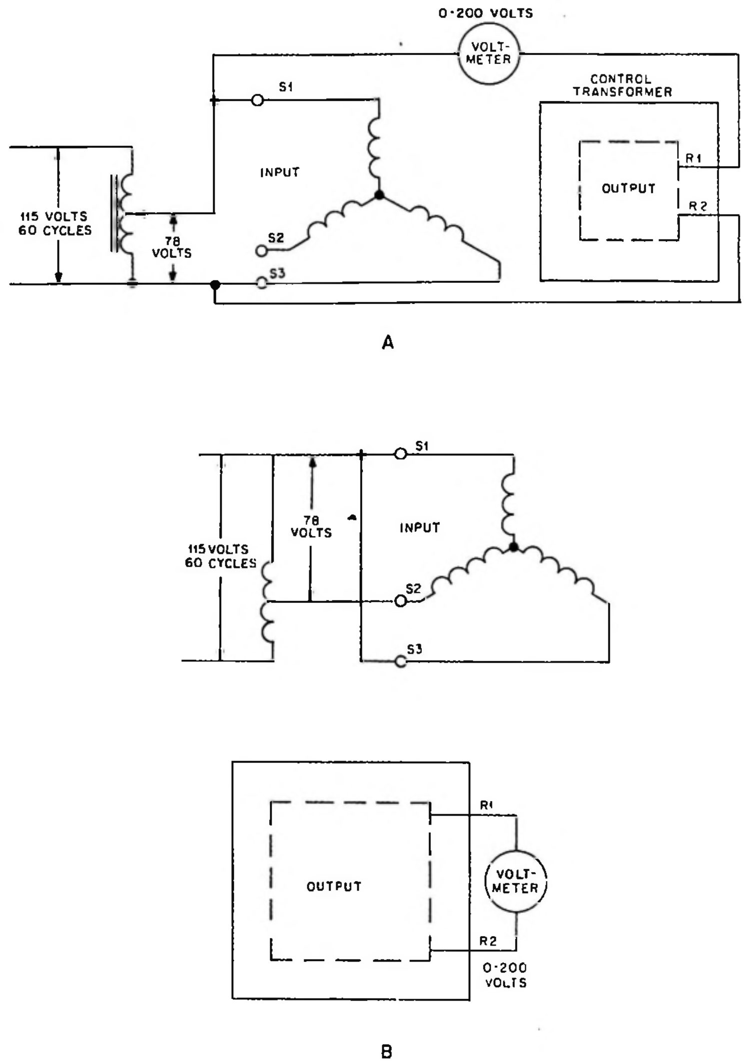


Figure 11-126H. Alignment Setup for Antenna Synchros

shaped pulse. The two half-amplitude points of the shaped pulse should be centered within the beam pulse. If the above requirement is not met, move the load connected to the delay line to one of the other connections on the delay line as necessary to meet the aforementioned requirement.

11-888. ANTENNA OUTPUT TROUBLES. Ordinarily, troubles will not originate in the antenna unless it has suffered mechanical damage or has been dismantled for some reason. Defects in the antenna output radiation, as indicated by failure of the aircraft to receive correct bearing information,

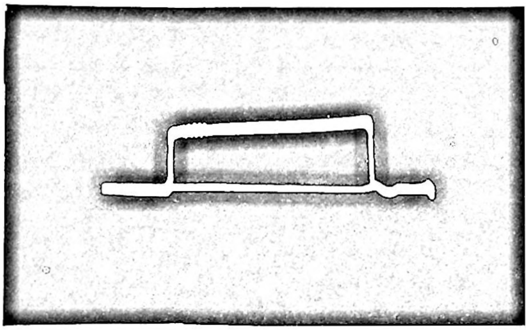


Figure 11-126I. Blanking Gate Waveform

distance information replies, or identification code, may have their origin in other units of the radio beacon.

11-889. Selection of the proper sites for these antenna tests is of prime concern because the precision of the information transmitted by the beacon is essential for proper operation. It should be noted that certain parts of these tests and alignment procedures can be performed only at depot installations. Thus, shipboard beacons will have to be brought to depot installations before some of the prescribed tests can be performed. Caution must be taken concerning reflecting objects such as metal fences, reinforced steel buildings, large trucks, and other objects in the test areas during tests (see figure 11-126J). People walking in the test areas during testing should maintain a distance of at least 30 feet from the antenna test site.

11-890. HARMONIC ANALYSIS. This test should be performed only in depot installations. The following equipment is required in testing antenna operational performance (see figure 11-126K):

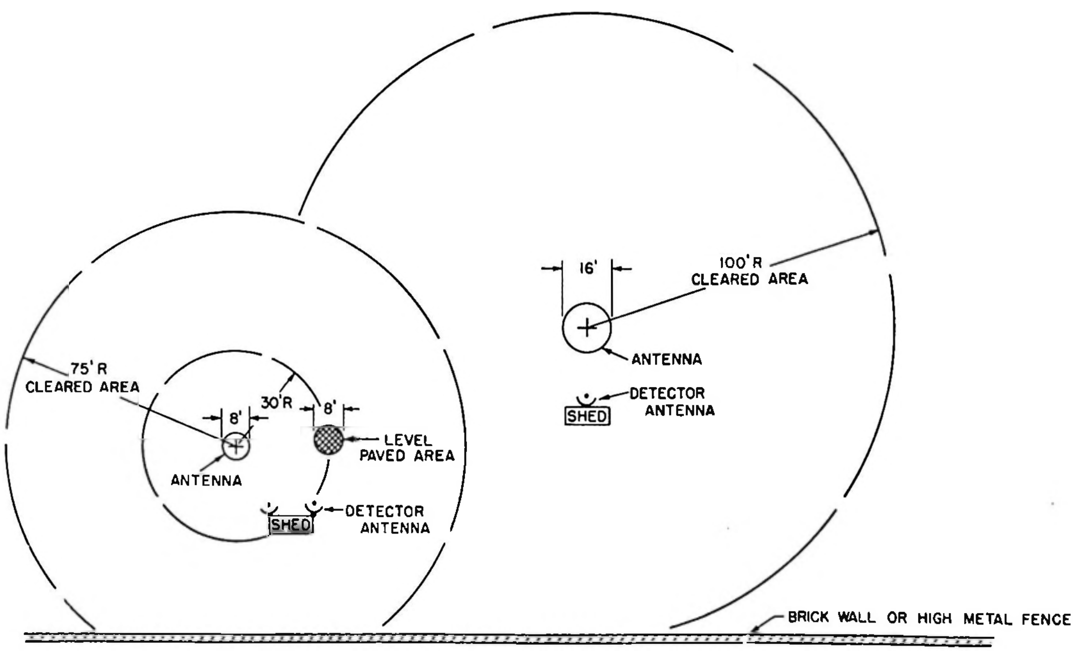


Figure 11-126J. Typical Antenna Test Sites

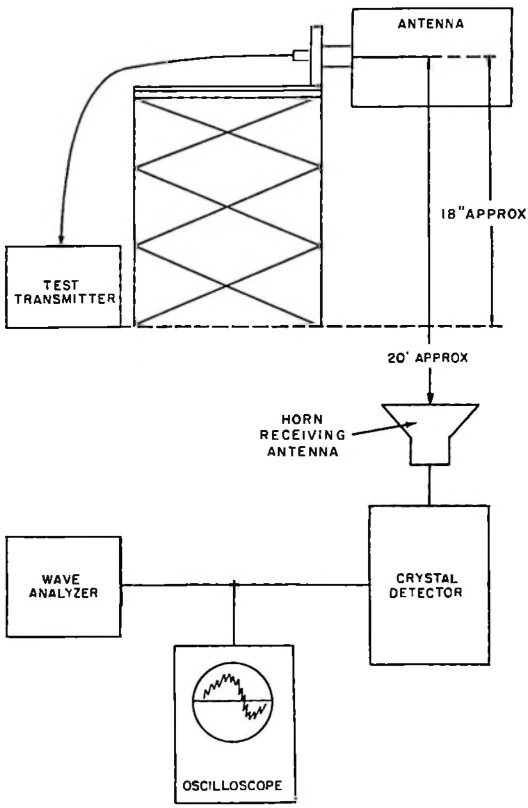


Figure 11-126K. Harmonic Analysis Test Setup

a. Crystal Detector. This device is used to rectify the radiated signal from the antenna for application to a dc meter or an oscilloscope. The detector should be linear within 3.0 percent.

b. Wave Analyzer.

c. Test Transmitter.

d. Oscilloscope.

11-891. Since this test is performed with the antenna lying on its side, the oil must be drained from the pitch and roll gear boxes. The antenna axis for this test will be horizontal. This position facilitates taking measurements of the performance of the antenna in the vertical plane. The antenna is generally installed approximately 18 feet above the ground. Horizontal rotation of the antenna represents a change in elevation angle. A receiving antenna horn is used so as to minimize reception of ground reflections. Care should be taken to place the receiving antenna at the same level as the center of radiation, and to make sure that the antenna rotates on a level plane. Failure to do this will cause the detecting point to change azimuth about the antenna.

a. Energize the beacon, and then tune the test transmitter to 987-mc continuous wave output for a low-band antenna or to 1188-mc continuous wave output for a high-band antenna.

b. Set the antenna to zero degrees in azimuth. The axis of the antenna must be accurately set at right angles to the line joining the apex to the antenna.

c. Energize the spin motor, and obtain a complete wave analysis of the signal using the wave analyzer in the standard manner. Measure the following frequencies: 15, 30, 45, 60, 135, 270, 405 cycles per second. Note that the 30-cycle modulation should be less than 15 percent of the 15-cycle modulation. The rms (.707 multiplied by peak voltage) sum voltage of all specified modulation frequencies should not exceed 20 percent of the fundamental, using as the fundamental the 15- or 135-cycle component, whichever is of higher magnitude. The 270-cycle modulation should be less than 15 percent of the 135-cycle modulation, and the vector sum of the modulation frequencies less cross modulation products should not exceed 20 percent of the 135-cycle modulation.

11-892. PERCENT MODULATION AND SIDEBAND TRACKING IN VERTICAL

PHASE. This test should be undertaken only at depot or shore installations. The following equipment is required in the measurement of the percent modulation and sideband tracking.

a. Test Transmitter.

b. Crystal Detector.

c. Oscilloscope.

d. Thermocouple Detector. This is a square-law device that is used to detect the radiated signal from the antenna and apply the output to a dc millivoltmeter. The thermocouple detector should be such that, when used with a correct dc millivoltmeter, it is linear within 3.0 percent.

e. Millivoltmeter (low internal resistance).

f. Remote Meter Box. This device is used in conjunction with the crystal detector for the convenience of remotely viewing the self-contained crystal detection meter.

g. Antenna control arrangement for speed control of the spin motor.

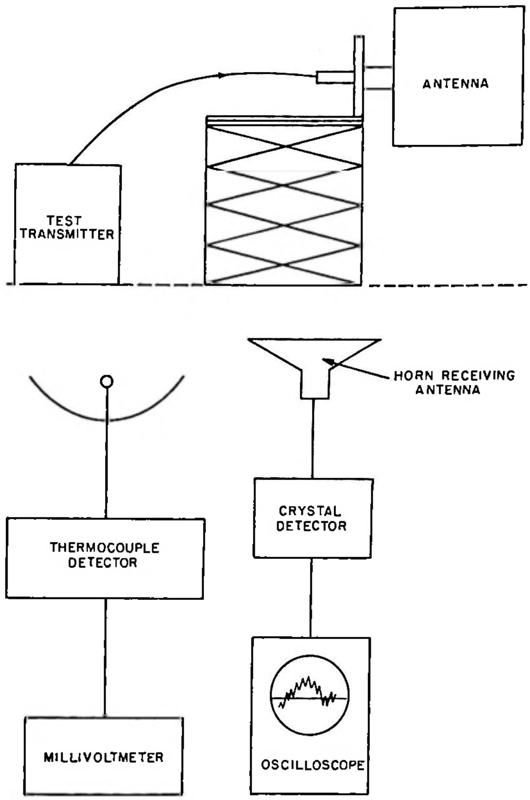


Figure 11-126L. Test Setup for Measurement of Percent modulation and Sideband Tracking

11-893. Connect the equipment as shown in figure 11-126L and proceed as follows:

a. Energize the equipment. Set the transmitter at 987 mc with square-wave modulation.

b. Install the thermocouple detector at a convenient distance so as to obtain full-scale deflection of the millivoltmeter. Rotate the antenna manually to obtain a maximum signal.

c. Install the crystal detector at the same height as the antenna axis and approximately 30 feet away. Connect the remote meter box. In this manner the crystal detector dc level, the thermocouple output, and the oscilloscope are side by side for easy reference.

d. Vary the output of the transmitter by detuning or by inserting various lengths of connecting cables. Read and record the thermocouple output and the deflection of the oscilloscope at as many points as is convenient within the range of the instrument. Using the thermocouple output as a standard, the calibration of the oscilloscope can be accomplished. Use as finely divided a reticle as practicable on the oscilloscope.

e. Adjust the test transmitter to 962 mc for low-band antennas and to 1151 mc for high-band antennas, using square-wave modulation. Adjust the antenna to zero degrees, representing the horizon.

f. Observe the oscilloscope pattern. With the antenna not rotating, an unmodulated square wave will be obtained. Adjust the oscilloscope gain so that the number of oscilloscope deflections equals the dc crystal current reading. Rotate the antenna spin motor and stop the antenna at approximately the peak of the modulation cycle; then adjust the oscilloscope so that the deflection does not exceed the reticle divisions.

g. Using a portable crystal detector, rotate the whole antenna by hand until the peak of the modulation cycle occurs in the direction of the horn antenna.

h. Apply power to the spin motor. Read the dc value of the detector current representing the carrier level.

i. Observe the oscilloscope pattern, which should resemble the pattern shown in figure 11-126M. Record the amplitudes shown as V_1, V_2, V_3, and V_4 for 10, 20, and 30 degrees respectively, and then repeat at 1024 mc for low-band antennas and at 1213 mc for high-band antennas. Calculate the percentage of modulation using the following formulas:

$$\text{Percent 15-cycle modulation} = \frac{(V_1 + V_2) - (V_3 - V_4)}{V_1 + V_2 + V_3 + V_4} \cdot 100$$

$$\text{Percent 135-cycle modulation} = \frac{(V_1 + V_3) - (V_2 + V_4)}{V_1 + V_2 + V_3 + V_4} \cdot 100$$

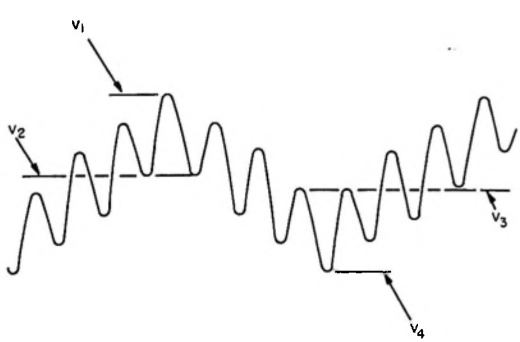


Figure 11-126M. Pattern for Determining Percent of Modulation

j. Calculate the relative sideband levels as follows:

$$\text{15-cycle sideband level} = \frac{V_1 + V_2 - V_3 - V_4}{2}$$

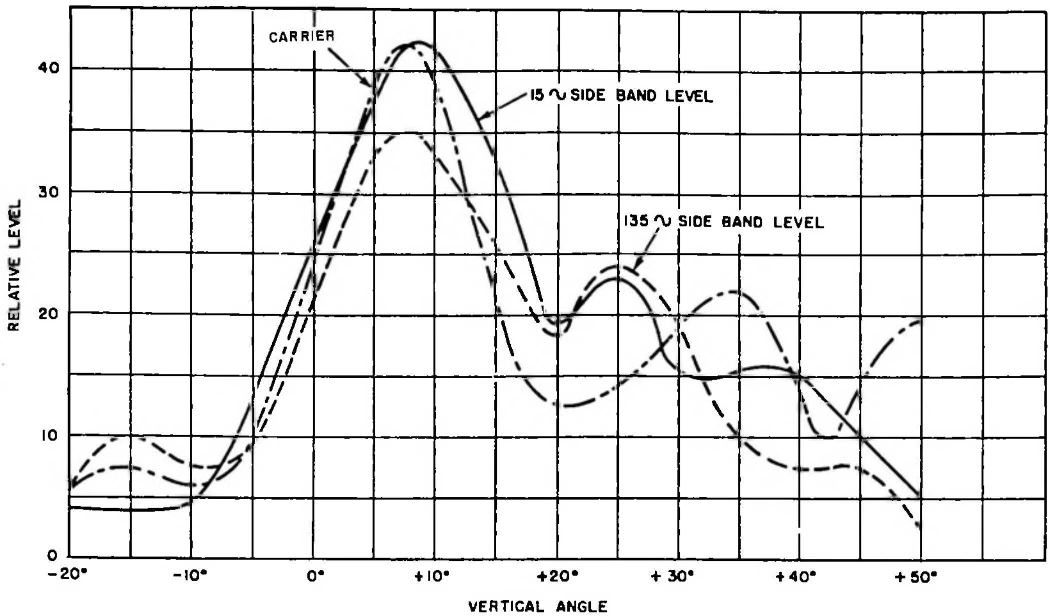
$$\text{135-cycle sideband level} = \frac{V_1 - V_2 + V_3 - V_4}{2}$$

k. Plot the carrier level, the percentages of modulation, and the sideband level as illustrated in figure 11-126N. Note that at the horizon the percentage of modulation for each modulation frequency should be 20 percent ± 8 percent, but the total sum of the percentages should not exceed 50 percent.

11-894. RELATIVE PHASE OF 15- AND 135-CPS COMPONENTS. This test should be performed only in depot or shore installations and only if either one of the parasite cylinders has been disassembled from the mounting plate. The following equipment is required in the measurement of the 15- and 135-cycle components (see figure 11-126O).

a. Crystal Detector.

b. Oscilloscope.



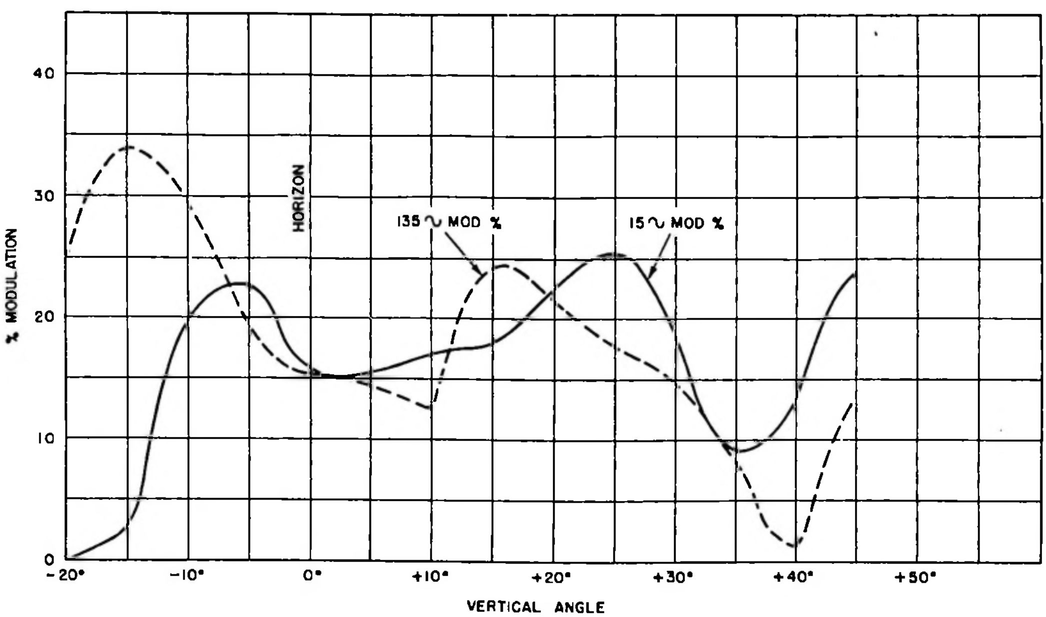


Figure 11-126N. Representative Curves, Percent of Modulation Test

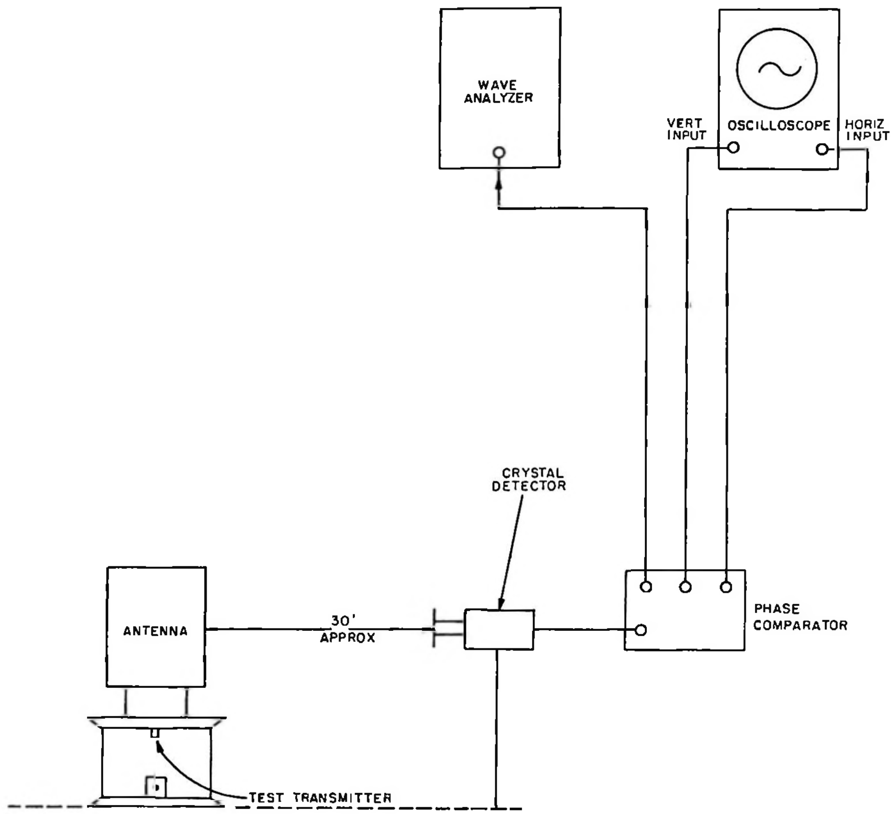


Figure 11-126O. Test Setup for Measurement of Relative Phase of the 15- and 135-Cycle Components

c. Wave Analyzer.

d. Phase Comparator. Since the relative phase between the 15-cps signal and the 135-cps signal must be adjusted to within 2-1/2 degrees over the entire radio-frequency spectrum, the comparator must be capable of measuring to an accuracy of 0.2 degree of the 135-cps component.

e. Test Transmitter.

f. Synchronous Motor Jig. This device is used to convert the motor of the shore antenna and shipboard antennas to a synchronous motor. This is done by substitution. Because of the difference in the rotational speeds of the motors, a belting arrangement is used to convert the antenna rotation to 900 revolutions per minute when using the synchronous motor.

11-895. Connect the equipment as shown in figure 11-126O and make sure that the antenna is in its normal erect position.

a. Adjust the test transmitter to a frequency of 962 mc cw output for low-band antennas and to 151 mc cw output for high-band antennas.

b. Set the turntable to any convenient reference point, which will hereafter be referred to as zero degrees azimuth. This azimuth need not necessarily correspond to any particular axis of the antenna.

c. Install the detector approximately 30 feet from the antenna in such a manner that its height is level with the center of the radiating elements.

d. Observe the waveshape on the oscilloscope to insure that proper operation is obtained. For this purpose the detector output is fed to the vertical deflection terminals of the oscilloscope, and the resulting pattern should indicate the presence of modulation resembling that shown in figure 11-126P. If interior modulation is obtained, determine the cause before proceeding with the test.

e. With the equipment connected as in figure 11-126O, set the switch of the phase comparator to the 135-cycle position and tune the wave analyzer to 135 cycles.

f. Adjust the phase and amplitude controls of the 135-cycle component of the phase comparator until a null is obtained on the wave analyzer. The controls must be adjusted alternately and the sensitivity of the wave analyzer increased progressively until the null is obtained. Record the 135-cycle phase reading. Do not disturb the amplitude of the phase settings until the measurement has been completed.

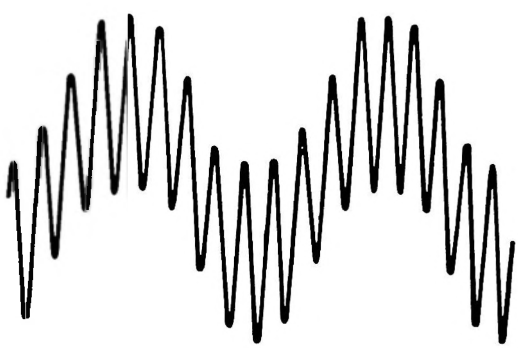


Figure 11-126P. Detected Antenna Pattern

g. Set the switch of the phase comparator to the 15-cycle position, and tune the wave analyzer to 15 cycles. Follow the procedure of step f to obtain a 15-cycle null by manipulating the amplitude and phase adjustment of the 15-cycle component of the phase comparator. Record the 15-cycle reading.

h. Set the switch of the phase comparator to the phase position, and note the Lissajous pattern on the oscilloscope. The oscilloscope must be set to use dc amplifiers in both the horizontal and vertical channels. The pattern will appear as shown in figure 11-126Q. If the Lissajous pattern is closed, the proper relative phase exists between the 15-cycle and 135-cycle components. If the pattern is not closed, adjust the 15-cycle phase of the phase comparator to obtain the closed condition. Note the new 15-cycle phase reading, and subtract it from that noted in step g. This represents the phase error. Record the sign of the phase error, the sign being negative if the reading obtained in step h is greater than that obtained in step g.

j. Repeat steps a through h at bearing angles of 90, 180, and 270 degrees with respect to the zero position chosen in step b.

k. Repeat steps a through j at 1024 mc for low-band antennas and at 1213 mc for high-band antennas.

l. Repeat steps a through j at a vertical elevation of 10 degrees and at a frequency of

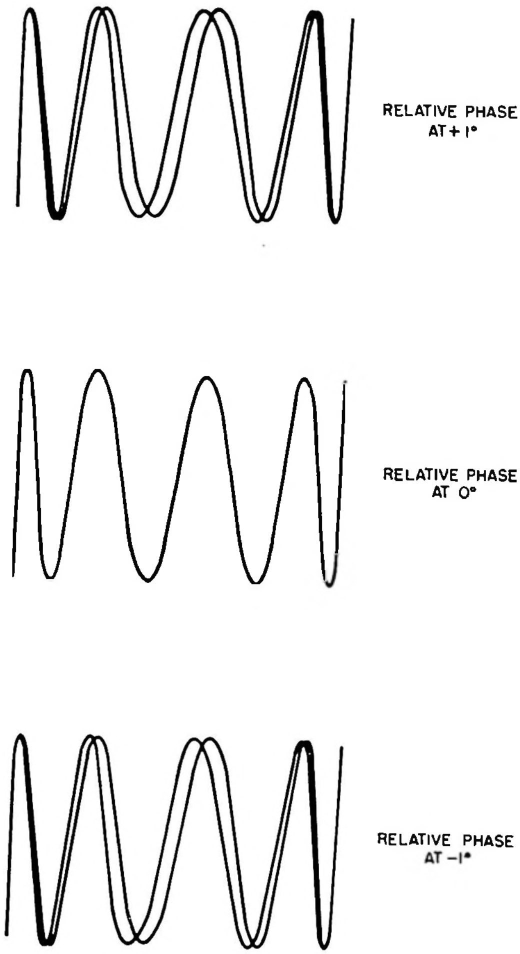


Figure 11-126Q. Phase Measurement Patterns of the 15- and 135-Cycle Components

987 mc for low-band antennas and at 1188 mc for high-band antennas.

m. The above procedure will give a total of 12 phase measurements consisting of combinations of four azimuths, three frequencies, and two elevations. At least 11 of the 12 phase measurements should not exceed a phase error of ± 2.5 degrees. The remaining measurement should not exceed an error of ± 3.5 degrees.

n. If the acceptable limits are not met, the position of the inner parasite must be adjusted. The amount and direction of the adjustment are derived from the deviation readings and should be such as to place the antenna within the acceptable limits.

11-896. CENTRAL ARRAY TESTS. Because of the inaccessibility of the central array, the following tests are suggested if a faulty array is suspected. This test may be made at any installation.

11-897. IMPEDANCE MEASUREMENT. The following equipment is required for the central array test:

a. Slotted Line.

b. Standing-Wave Indicator.

c. Signal Generator.

11-898. Connect the equipment as illustrated in figure 11-126R, and then proceed with the test.

a. Set the signal generator to a frequency of 960 mc for low-band antennas and to 1160 mc for high-band antennas, and adjust for pulse modulation using maximum pulse width and a repetition rate corresponding to the tuned frequency of the standing-wave indicator (normally 1000 cps).

b. Tune the slotted-line probe for maximum deflection of the indicator.

c. Slide the probe along the line to a position of maximum output, and set the gain of the indicator to read 1.0.

d. Slide the probe along the line to a position of minimum output, and read the standing-wave ratio as indicated.

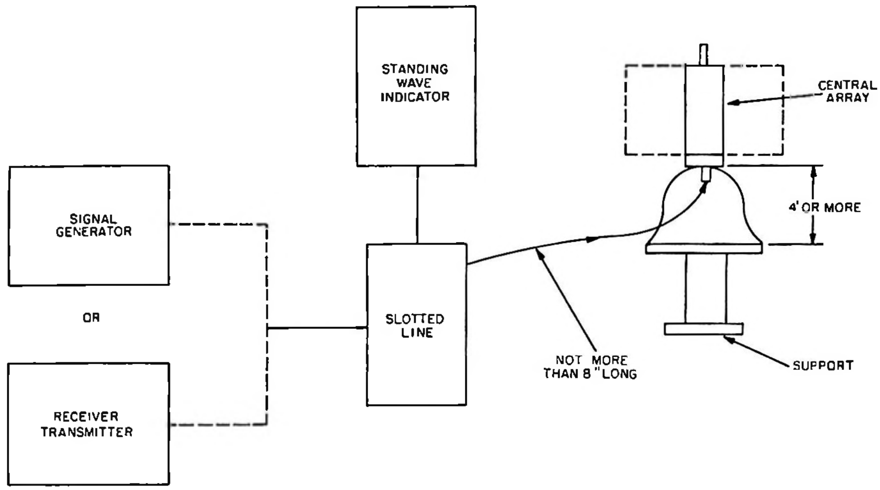


Figure 11-126R. Central Array Test Setup (Impedance Measurement)

e. Repeat steps a through d for frequencies of every 10 mc for low-band antennas and for every 10 mc up to 1220 mc for high-band antennas.

11-899. VERTICAL PATTERN MEASUREMENT. This test should be undertaken only in depot or shore installations. The following equipment is required in measuring the vertical pattern:

a. Crystal Detector.

b. Remote Meter Box.

c. Test Transmitter.

11-900. TEST CONNECTIONS. Connect the equipment required as illustrated in figure 11-126S. If a receiver-transmitter unit is not available, use a test transmitter. The antenna should be installed with its axis horizontal and approximately 10 feet above the ground. The detector should be placed at least 30 feet away, and at the same height as the axis of the antenna. Upon completion of the above, make the following adjustments:

a. Energize the beacon. If no receiver-transmitter unit is available, tune the test transmitter to 987 mc for low-band arrays and to 1188 mc for high-band arrays, using a cw output.

b. Observe and record the detector reading at —2, 0, and +2 degrees and at all peaks and nulls from 20 degrees to 60 degrees. Record the angles and values.

c. Correct all readings for relative rf level.

11-901. ACCEPTABLE LIMITS.

a. The slope determined by the values measured at +2 degrees shall be a minimum of 0.5 db per degree. Therefore, the value at —2 degrees should be at least two decibels below the value at +2 degrees.

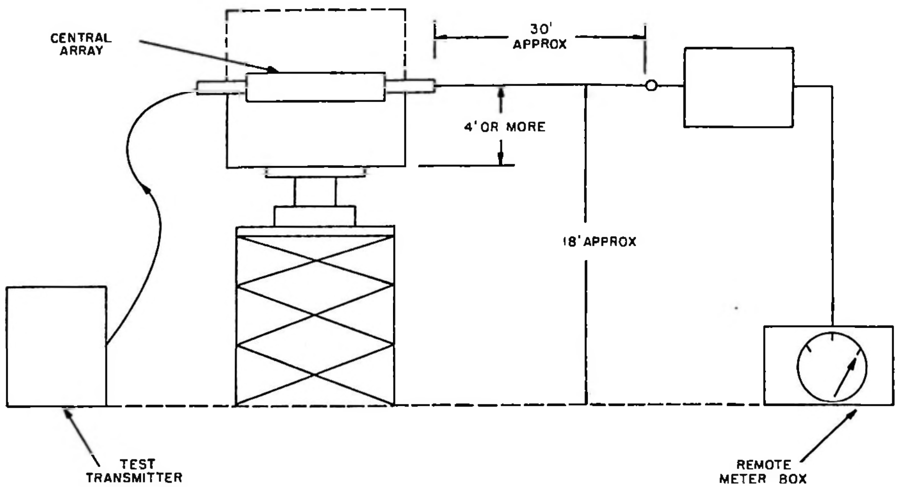


Figure 11-126S. Central Array Test Setup (Vertical Pattern Measurement)

b. The amplitude at zero degrees should be not less than 50 percent of the maximum amplitude.

c. No nulls deeper than 25 db below the maximum should occur between zero and 50 degrees.

d. No nulls deeper than 12 db below the maximum should occur between zero and 15 degrees.

11-902. VOR EQUIPMENT.

11-903. TRANSMITTER.

11-904. RESOLVER. A resolver is made up of rotor and stator windings, and its operation is somewhat analogous to that of a transformer, in that it is used to produce voltages rather than rotation. The most common type of resolver, shown in the schematic diagram of figure 11-126T, has two distributed stator windings wound inside a cylindrical form with the two coil axes at right angles to each other. The stator windings act as transformer primaries. Two secondary windings, also with coil axes perpendicular, are carried on a cylindrical rotor. Some models have extra windings for

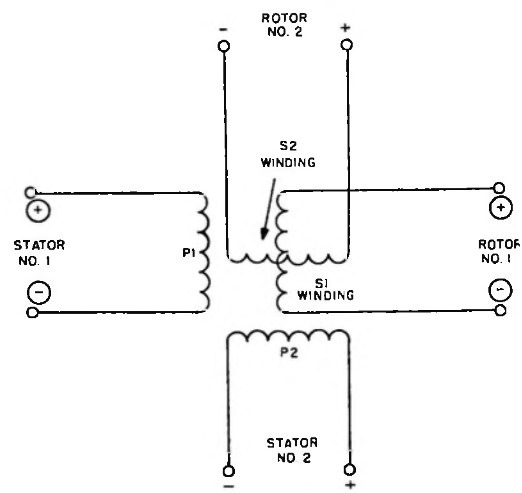


Figure 11-126T. Resolver in the Zero Position feedback purposes; others may have one primary or one secondary omitted.

11-905. The figure shows the resolver with the rotor in the zero position, the designations P and S referring to the primary and secondary windings. The polarities indicated are instantaneous values. The encircled polarities refer to the P1 — S1 transformer to distinguish them from the polarities of the P2 — S2 transformer. The magnetic coupling for the zero position is such that primary P1 induces maximum voltage in secondary S1, but no voltage in secondary S2. Similarly, P2 induces maximum voltage in S2, and no voltage in S1. In addition, note that the resolver secondary windings are designed so that the voltages induced in each are in phase with their respective primary voltages, as indicated in the figure. In common design, the transformation ratio between a primary winding and a secondary winding is one-to-one.

11-906. Part A of figure 11-126U gives an example of how combined voltages are derived when the rotor is turned 45 degrees from the zero position. Part B shows a graphical representation of the resolver function. When voltages are applied to resolver primaries P1 and P2, a resultant magnetic field is produced which is the sum of the individual primary fields. The induced voltages in secondaries S1 and S2 are proportional to the components of the resultant fields which lie parallel to each of the windings, as shown in part B of the figure. The fields produced by the equal potentials across P1 and P2 are represented by vectors having a magnitude of 1 volt. The resultant field shows a vector equivalent of 1.4 volts. Since the S1 axis is parallel to the resultant field, its induced voltage is maximum, or 1.4 volts. On the other hand, the S2 axis, being displaced by 90 degrees, causes the induced voltage in S2 to be zero.

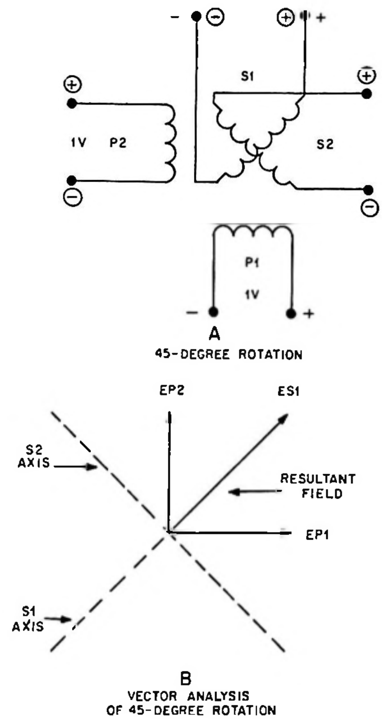


Figure 11-126U. Resolver Action

11-907. AUTOSYN TRACKING ADJUSTMENTS. The Autosyn tracking adjustments can be performed as follows:

a. Using the resolver track adjust control, set both the "OMNIRANGE", phase control, and the azimuth selector (autosyn control) to the zero position.

b. Disassemble the locking arrangement about the autosyn itself to enable it to be rotated.

c. Now rotate the autosyn as required to obtain a reading of exactly "0" on the phase balance indicator.

d. Reassemble the locking arrangement, and place it back about the autosyn unit to hold it stationary.

e. With controls still set as in step a, turn the resolver track adjust control fully counterclockwise.

f. Note this reading on the phase balance indicator. If it does not read zero, a phase error exists. Measure the exact magnitude of this phase error by rotating the azimuth selector (azimuth control) as required to get a zero reading on the phase balance indicator.

g. Plot this error on a graph, with error versus bearing as co-ordinates.

h. Following this procedure, plot the phase error at 30-degree intervals by setting both the "OMNIRANGE", phase control, and azimuth selector on 0, 30, 60, 90, etc.

i. Now turn the resolver track adjust control fully clockwise, and repeat steps f through h, entering these error readings on the same graph.

j. The curve for the counterclockwise setting of the resolver track adjust control crosses the curve for the clockwise settings at four points, forming two large and two small loops. Selecting the crossover point to the left of either large loop, read forward or back to the nearest degree. Set both the phase control and the autosyn control to this bearing. Rotate the resolver track adjust control from full clockwise to full counterclockwise. There should be no change in reading on the azimuth selector and phase balance indicator. If there is a phase error, shift both the phase control and the autosyn control slightly clockwise or counterclockwise (both at the same setting) to obtain zero indication on the phase balance indicator. Now shift both phasing controls forward 60 degrees from this critical setting (into the large loop on the graph), and adjust the resolver track adjust control to obtain phase balance, i.e., "0" reading on the phase balance indicator.

11-908. VARIABLE TRACKING ADJUSTMENTS. The variable tracking adjustments can be performed as follows:

a. With the "OMNIRANGE" and the azimuth selector (autosyn control) at "0", rotate the variable track adjust control to obtain a phase balance, i.e., a reading of zero on the phase balance indicator.

b. Keeping the "OMNIRANGE" control at the zero phase position, rotate the azimuth selector to "180". The phase balance indicator should still read zero.

c. If the phase balance indicator does not now give a zero indication, adjust the azimuth selector to establish zero, noting the error in degrees, and reset the variable track adjust control to a point where the error in degrees reflected on the azimuth selector dial is halved.

d. Now repeat steps a and b as often as required to reduce the error in degrees, as noted on the azimuth selector dial, to zero.

11-909. MODULATION ELIMINATOR. The modulation eliminator is designed to remove the amplitude modulation from that portion of the transmitter output which is fed to the vhf goniometer. It is essential that the rf input to the goniometer be practically free from modulation, to prevent cross-modulation between the carrier and sideband signals. The modulation eliminator consists essentially of two uhf diodes connected in an rf bridge circuit. A simplified schematic of the modulation eliminator is shown in figure 11-126V. The diodes are connected at one junction of the bridge as shown in the

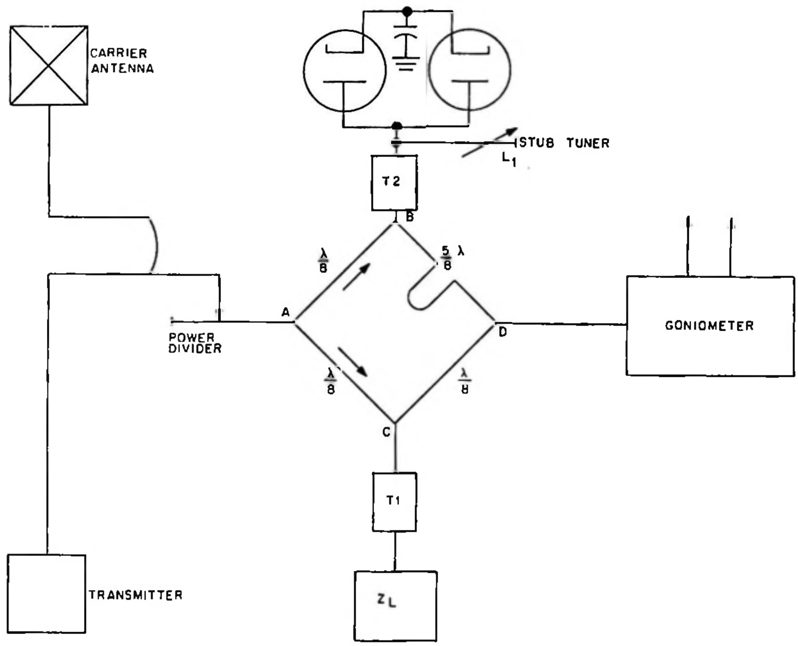


Figure 11-126V. Modulation Eliminator, Simplified Schematic

figure. A resistive load, Z_L, is connected at the junction of the bridge opposite the diodes.

11-910. The rf from the transmitter is fed to the bridge at point A and divides along the arms of the bridge as indicated. Since one of the arms of the bridge is made one-half wave length longer than the other arms, the rf from point B arrives at point D, 180 degrees out of phase with the rf from point C The bridge output is equal to the difference in the amplitudes of the rf energy at points B and C. This difference in amplitudes is determined by the difference in impedance of the diodes and Z_L. As the difference in impedance decreases, the bridge tends to approach a balance. The impedance of load Z_L is fixed, whereas the impedance of the diodes varies with the envelope of the modulation on the carrier arriving at point B. That is, the diode impedance decreases as the amplitude of the modulation increases. Thus, the bridge tends to become more balanced on the peaks of the modulation. In other words, the output from the bridge tends to remain at a constant level, the modulation peaks being "compressed" is shown in figure 11-126W.

11-911. PHASING ADJUSTMENTS. You can perform phasing adjustments by using the following procedure:

a. With the "OMNIRANGE" (phasing control) set at its zero phase position, turn the phase sense adjustment control fully clockwise, and set the azimuth selector (autosyn control) to 359 degrees.

b. Now slowly rotate the phase sense adjustment control until a point is reached

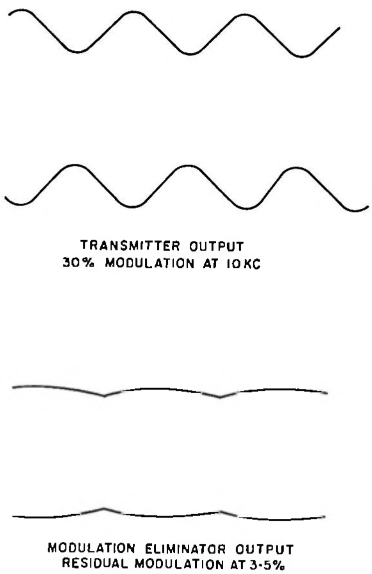


Figure 11-126W. Modulation Eliminator Waveform Patterns

when the alarm on-lamp lights and the alarm off-lamp is extinguished. If the lamps flash alternately, turn the phase sense adjustment control a little more in the clockwise direction until the alarm on-lamp remains lighted.

c. Set the azimuth selector back to the phase position. The alarm on-lamp should go off and the alarm off-lamp should again glow.

11-912. MODULATOR.

11-913. SPACE MODULATION ADJUSTMENT. Space modulation is a form of amplitude modulation which requires two carriers on the same frequency to be in phase. One carrier remains fixed in amplitude while the other must vary in a manner which will add to, or subtract from, the fixed carrier. The modulation therefore takes place in space, and the two signals received by the receiver combine and vary in amplitude. Space modulation is similar to some types of fading and flutter that occur in receivers when receiving both a ground wave and a reflected sky wave. In the omnirange signal, the two carriers remain at constant amplitudes. Because of the rotation of the eccentric variable carrier lobe, the receiving antenna, at a fixed point in space, will receive varying amounts of signal depending on the position of the variable carrier lobe maximum or minimum. The ratio between the power in the reference carrier radiation field and that in the variable carrier field determines the percentage of modulation at the receiver. If the radiation field of a lobe were equal to the reference carrier in amplitude, 100 percent modulation would result. The amount of variable carrier power delivered to the antenna in omnirange equipments is adjusted to produce 30-percent modulation. The variable carrier is also rotated at the cps rate by the variable carrier outputs from the goniometer. The percentage of space modulation is increased by increasing the power fed to the sideband antennas without increasing the power to the carrier fed to the center antenna. Adjust the percentage of modulation as follows:

a. Place the controls on the transmitter to the transmit operational position, and properly position the controls of the monitor receiver.

b. With the goniometer mechanism off, rotate the tone switch to "315" and note exact reading (minimum). Then, rotate the tone wheel to "315" and note the maximum reading.

c. Compute the space modulation according to the following formula:

$$\text{Percent of Modulation} = \frac{E_{max} - E_{min}}{E_{max} + E_{min}} \times 100\%$$

11-914. FM SUBCARRIER ADJUSTMENT. Adjust the level of the subcarrier to obtain

30-percent modulation of the carrier, as follows:

a. Close the interlock switch on the transmitter to turn power on with door open.

b. Turn the goniometer "on".

c. Tune the high-voltage power supply for full output, and then, position the controls on the transmitter to the operational transmit position.

d. Adjust the modulator driver to obtain a reading of "30" on the modulator output meter.

11-915. CODE IDENTIFICATION ADJUSTMENT. Adjust the level of the keyer 1020-cycle-per-second code identification signal to obtain 10-percent modulation of the carrier, as follows:

a. Close the interlock switch on the transmitter to turn the power on with the door open.

b. Tune the high-voltage power supply for full output, and then position the controls on the transmitter to the operational transmit position.

c. Turn the toggle switches on the keyer to the off position.

d. Using clip leads, connect a voltmeter across the transformer of the audio-frequency oscillator.

e. Turn the power "on".

f. Adjust the gain to obtain a reading of 1.9 volts. Release the interlock switch.

g. Adjust the modulator driver to obtain a reading of "10" on the modulation output meter of the modulator.

11-916. VOICE CHANNEL ADJUSTMENT. To perform the voice-channel adjustment, proceed as follows:

a. Close the interlock switch on the transmitter to turn the power on with the door open.

b. Connect the plug from the modulator driver to feed the unkeyed 1020-cycle-per-second code identification signal into the speech input.

c. Tune the high-voltage power supply for full output, and then position the controls on the transmitter to the operational transmit position.

d. Turn the keyer switch to the "on" position.

e. Adjust the modulator driver to obtain a reading of "30" on the modulation output meter of the modulator.

11-917. KEYING HEAD ADJUSTMENT. To perform the keying head adjustment, proceed as follows:

a. Throw the toggle switches on the keyer to the "off" position.

b. Loosen the setscrews which lock the keying head pins and pull out pins in a sequence corresponding to the dots and dashes of the station's code identification. Pull out three pins for each dash, one pin for each dot, one pin between dots and dashes, and three pins between letters.

c. Tighten all setscrews.

d. Rotate the keying head by hand to make sure that the pins operate the microswitch properly.

e. Turn the toggle switches on the keyer to the "on" position, and then position the controls on the transmitter to the operational transmit position.

f. Visually check to see that the modulation output meter on the modulator reads "O" during the intervals between letters and that the needle overshoots "10" for dashes.

g. Check the indicator connected to the transmitter control. The code should be distinct, and the length of each dot should be 1/10 second plus or minus 10 percent.

h. If it is necessary to adjust the length of the dot, loosen the microswitch mounting screws and move the switch close to the pins to lengthen the dot and away from the pins to shorten the dot. Tighten the screws with the switch in the correct position.

11-918. ANTENNA.

11-919. RADIATED PATTERN-CIRCULATION CHECK. The radiation from the antenna is an omnidirectional reference carrier radiation pattern and is in the form of a clockwise rotating eccentric lobe (also called a limacon) produced by the variable carrier. The reference carrier, as the name implies, is the reference signal with which the rotational position (phase) of the variable carrier is compared at any specific instant. The eccentric variable carrier lobe is produced by the addition of the reference carrier omnidirectional pattern and the two variable carrier figure-of-eight patterns of a specific instant of time. The specific instant of time is measured in degrees of arc from magnetic north in a clockwise direction. The eccentric variable carrier lobe rotates at the specified cycles-per-second rate, and the point of azimuth at which the resulting vector maximum falls, at a specific instant, is known as a radial. All radials are referenced to magnetic north regardless of antenna geographic location.

11-920. Vertically polarized radiation is undesirable and is held to a minimum by the slot dimensions of the antenna. Another factor governing vertical radiation is the size of the counterpoise under the antenna and the distance of the counterpoise from the antenna. The equipment shelter has been designed and positioned to reflect as small a vertical radiation component as practicable. Any vertical radiation, mixed with horizontal radiation, will be slightly out of phase with the horizontal radiation at the receiving antenna and will cause bearing errors. These errors will be more pronounced close to the station and will vary with altitude and receiving antenna attitude. The 'cone of confusion', which indicates over-the-station position to the pilot, will be extremely erratic when excessive vertically polarized radiation exists. Therefore, over-all design has reduced the vertical radiation of the antenna to a negligible amount, and the antenna produces only horizontally polarized radiation fields.

11-921. The reference carrier radiation pattern is horizontally polarized and circular with the antenna at the center, as shown in figure 11-126X. Because all antenna slots are excited equally and in the same polarity, the phase of the reference carrier radiation pattern is the same at any point of azimuth around the antenna. The field strength at a given distance from the antenna is also the same at any point of azimuth around the antenna.

11-922. The two variable carriers, displaced electrically by 90 degrees, excite pairs of antenna slots which are displaced physically by 90 degrees, as shown in figure 11-126Y. One variable carrier excites the northwest/southeast slots, and another variable carrier excites the northeast/southwest slots. This displacement con-

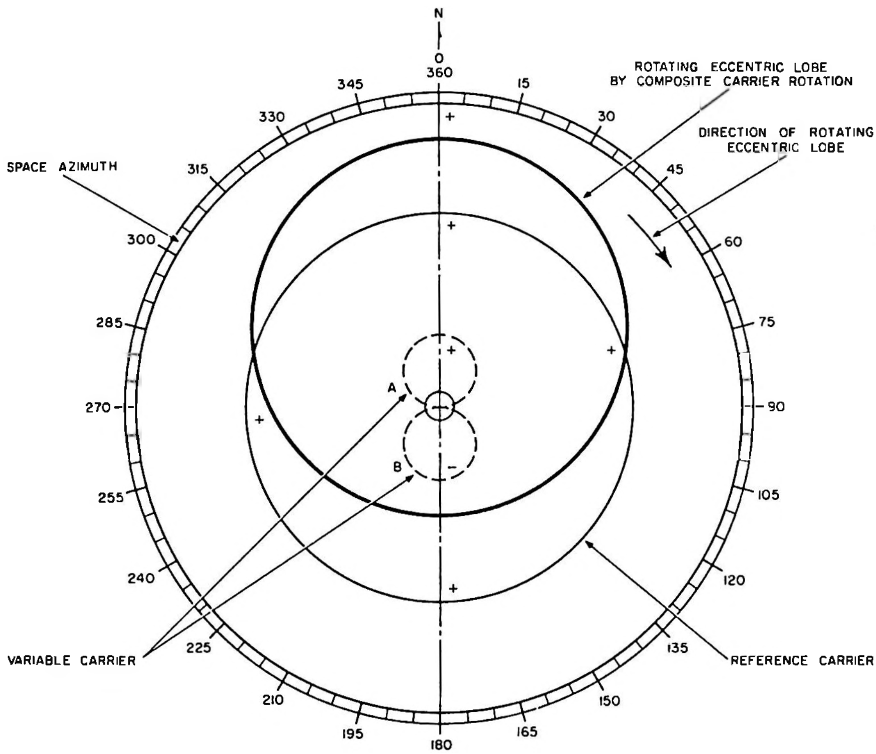


Figure 11-126X. Antenna Total Radiation Pattern

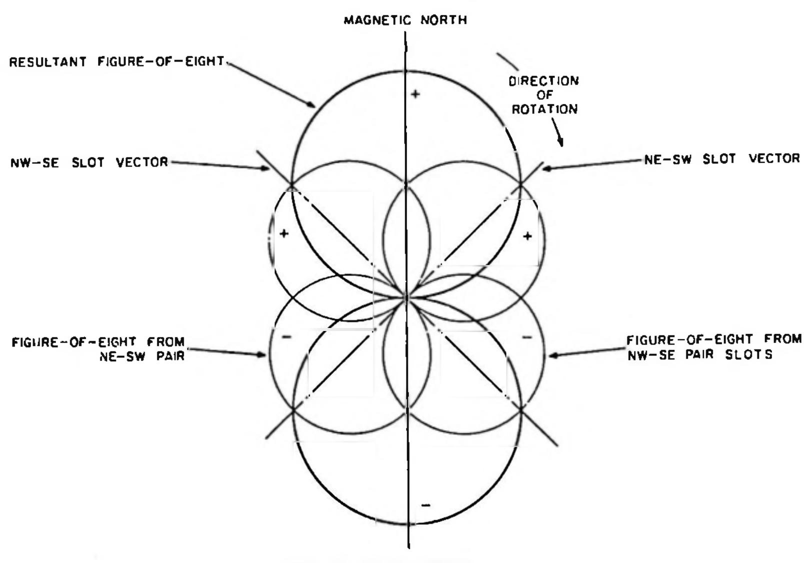


Figure 11-126Y. Antenna Figure-of-Eight Radiation Pattern figuration does not change physically at any time. The figure-of-eight radiation patterns produced by either slot pair take the form of two diagonally opposite circular fields of opposite radiation polarity. The magnitude, or field strength, of the polarized fields varies from zero to a fixed maximum, and the polarity reverses. These changes are controlled by the output delivered to the slot pair by the goniometer. When equal power and like polarities are fed to the antenna slots from the goniometer, the resultant figure-of-eight radiation pattern of figure 11-126Y is produced in the north and south direction. There is no reference input.

11-923. Rotation is supplied to the resultant figure-of-eight radiation pattern by varying the amplitude and polarity relationship between the two antenna slot pairs. In part A of figure 11-126Z, the northwest/southeast slots are receiving maximum excitation while the northeast/southwest slots are at zero excitation. This places the resultant lobe at 315 degrees. However, in part B of the figure, the excitation to the northwest/southeast slots has been reduced, and the excitation to the northeast/southwest slots is increasing. The resultant radiation pattern has moved to 337.5 degrees. Part C of the figure duplicates figure 11-126Y and results in movement to zero degrees. Progressing from 315 degrees through 45 degrees, the northwest and northeast minor lobes are all positive, and amplitude variations produce rotation. In part F of the figure, the polarity of the northwest/southeast fields has reversed, but because of their low amplitude in comparison to the northeast/southwest fields, the resultant radiation pattern moves smoothly to 67.5 degrees. Figure 11-126Z indicates the method by which the resultant radiation pattern is rotated through 180 degrees and, subsequently, through 360 degrees.

11-924. ANTENNA POSITION ALIGNMENT. The omnirange antenna is a four-slot cylindrical antenna used to radiate a composite carrier consisting of a reference carrier and two variable carrier signals. The four equally spaced longitudinal slots in the wall of the cylindrical cavity are the radiating elements. These four parallel slots, equally spaced around the circumference of the metal cylinder, are usually designated northwest, northeast, southeast, and southwest. Small adjustable capacitors are placed across each slot to compensate for manufacturing tolerances in slot dimensions.

11-925. The antenna is centered and oriented so that the center between the northeast and northwest slots faces magnetic north.

11-926. GONIOMETER.

11-927. GENERAL. A typical goniometer assembly (figure 11-126AA) is composed of a synchronous motor, a reference signal generator, a capacitive type goniometer, and line balance transformers. One purpose of this equipment is to generate a 9960-cps tone, frequency-modulated at 30 cps (± 480-cps deviation). This tone is then used to amplitude-modulate the rf carrier of the transmitter, the output of which provides the reference signal in an omnirange equipment. The tone is generated by means of a toothed wheel that rotates in the field of a coil energized by a permanent magnet. The second function of this equipment is to generate a pair of rf signals which are modulated 90 degrees out of phase with each other. The rf carrier of the transmitter is, in turn, modulated by these two signals to provide separate outputs to two pairs of omnirange antennas.

11-928. The following description of the operation of the major components comprising the goniometer will provide an under-

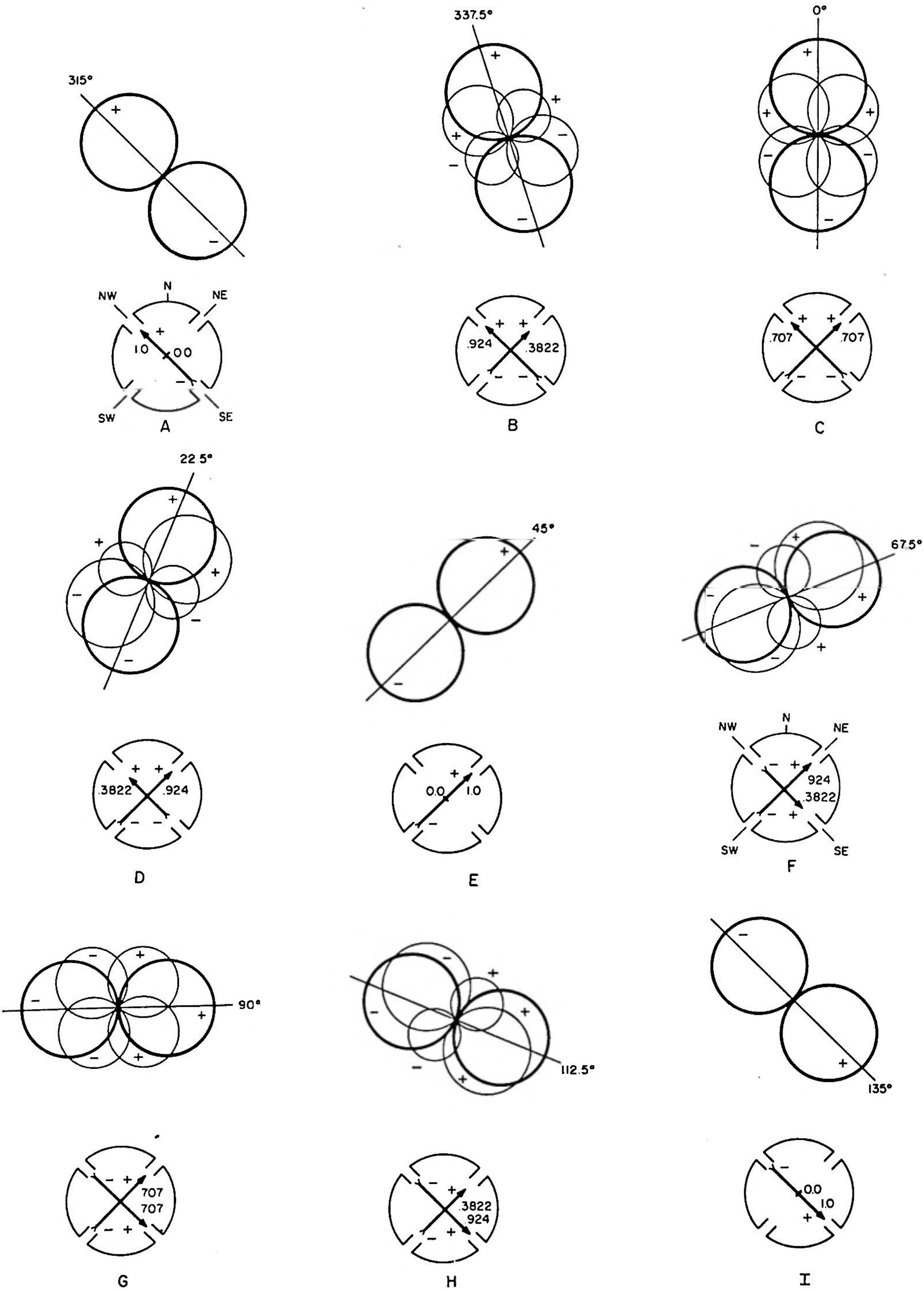


Figure 11-126Z. Antenna Figure-of-Eight Pattern Rotation

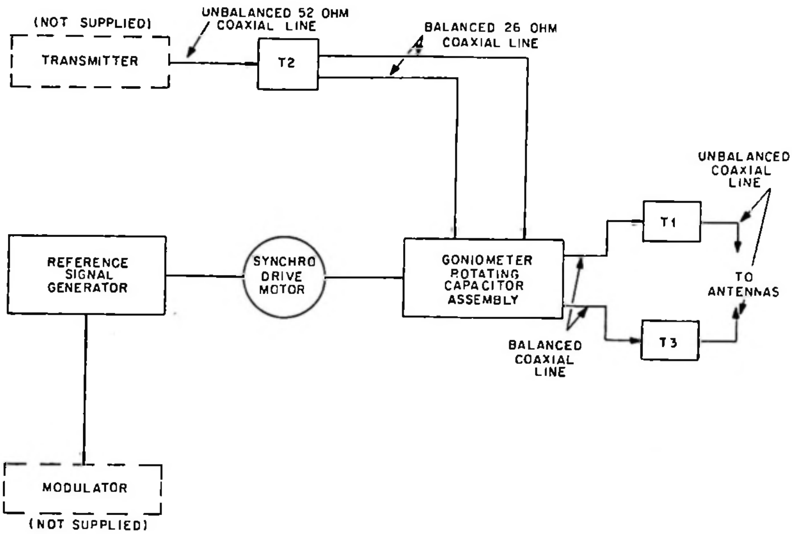


Figure 11-126AA. Typical Goniometer Assembly, Block Diagram

standing of the operational function of each component.

a. The synchronous drive motor is a single-phase, line-starting motor equipped with a thermal overload switch. This motor drives the reference signal generator and goniometer capacitor in synchronism. The speed of rotation is 1800 rpm when supplied from a 60-cps source. It is important that this speed be maintained as it directly affects both the center frequency and the modulation frequency of the tone used to modulate the rf carrier of the transmitter.

b. The reference signal generator consists of a coil of wire wound over a permanent magnet whose lines of flux pass through a toothed gear called the tone wheel. As the teeth of this gear pass the pole piece, a varying magnetic field is produced around the magnet, inducing a voltage in the coil. The basic frequency of this voltage is dependent upon the number of teeth on the wheel and the speed of rotation, and is designed to be 9960 cps. In addition, the spacing between the teeth is varied around the circumference of the wheel so that the frequency varies ± 480 cps from 9960 cps as the wheel revolves. You can see than, that a tone wheel having 332 teeth and rotating at 30 cps (1800 rpm/60 sec = 30 cps) results in a center frequency of 9960 cps (332 x 30 cps = 9960 cps). The resulting signal, 9960 cps deviating ± 460 cps at a 30-cps rate, is referred to as the 10-kc frequency-modulated subcarrier. This signal is fed to the modulator driver and used to amplitude-modulate the rf carrier, as shown in figure 11-126AB.

c. The capacitor portion of the goniometer assembly consists of eight rotor plates meshed with eight stator plates. Four rotor plates and four stator plates are located at each end of the rotor shaft. The rotor shaft is split along its rotational axis, and each shaft section is insulated from the other.

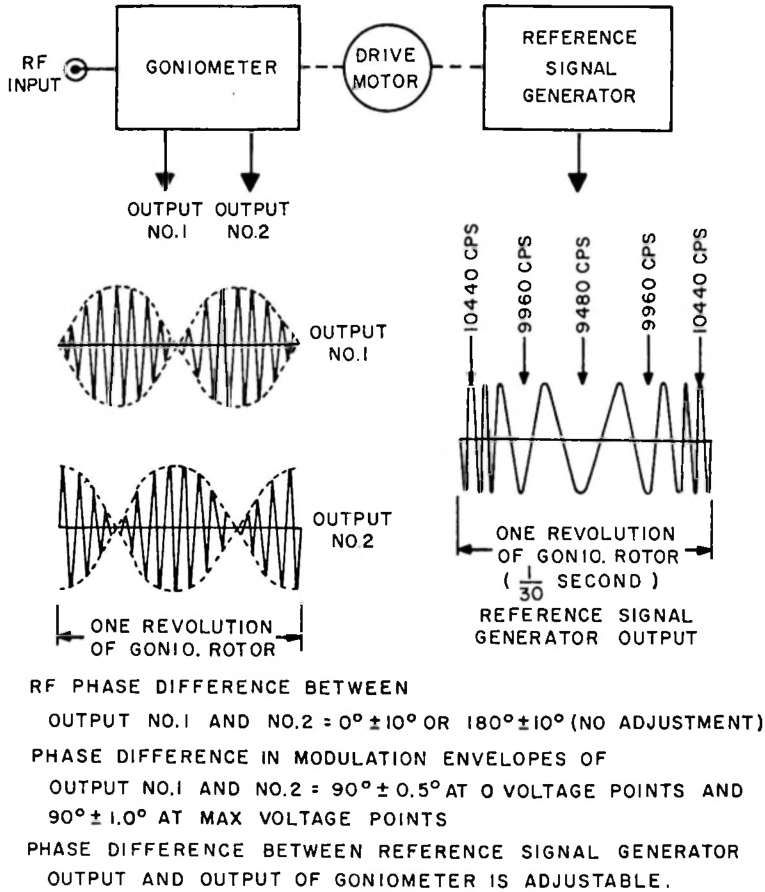


Figure 11-126AB. Reference Signal Generator and Goniometer Outputs

The rotor plates are designed so that they provide sinusoidal modulation of the rf carrier as the shaft rotates. Since the stator plates at one end of the shaft are physically displaced 90 degrees from those at the other end, one set of plates provide sine-modulated rf energy to one of the goniometer output circuits, while the other set of plates provides cosine-modulated rf energy to the remaining output circuit. Consider goniometer output No. 1 as an example and adopt the following terminology:

R_1 = Rotor No. 1

R_2 = Rotor No. 2 (insulated from Rotor No. 1)

S_1 = Stator No. 1

S_2 = Stator No. 2 (insulated from stator No. 1)

C_i = A fixed value capacitor which couples rf power to the rotor

C_o = The capacitance between one set of rotor plates and one set of stator plates when half of the rotor area is exposed to one set of stator plates

θ = Angle of rotation

The various capacitances at any instant can then be calculated by the following equations:

Capacitance between $R_1S_1 = C_o (1 + \sin\theta)$

Capacitance between $R_1S_1 = C_o (1 - \sin\theta)$

Capacitance between $R_2S_2 = C_o (1 + \sin\theta)$

Capacitance between $R_2S_1 = C_o (1 - \sin\theta)$

The relationships of the capacitors to the four principal angles of rotation, along with goniometer capacitor equivalent circuits, are illustrated in figure 11-126AC.

d. Line balance transformers are used to provide electrical quarter-wave matching sections at 115 mc, which is the mean operating frequency of this equipment. One problem in converting unbalanced lines to balanced lines, or vice versa, is the possibility of coupling between current flowing outside the shield of a conductor to that flowing inside the shield. This is overcome by maintaining a continuous shield from one line to the other so that there is no way for outside current to get into the line. The second problem is to maintain equal impedances from the conductors of the balanced line to their respective shields. This is done to maintain equal currents in these lines and is accomplished by the use of quarter-wave matching stubs. The third problem is to maintain a constant line impedance to obtain maximum power transfer. To provide constant line impedance, the matching wave stubs are made to match each other. It should be noted that over the range of operation (112 mc to 118 mc) the stub impedance, and hence the line currents, will vary. however, the currents in the two lines remain equal to each other and properly balanced.

11-929. ALIGNMENT AND ADJUSTMENTS. Whenever the goniometer capacitor, tone wheel, rotating shafts, or the coupling between the two shafts is replaced or disturbed in any manner, you must accomplish the following alignment procedure:

a. Remove the capacitor housing cover.

b. Loosen the setscrews on the coupling between the capacitor shaft and the reference signal generator shaft.

c. Rotate the tone wheel (figure 11-126AD, point A) for a reading of exactly 135 degrees on the 360-degree dial (figure 11-126AD, point B).

d. Tighten the setscrews on the reference signal generator side of the shaft coupling.

e. While holding the tone wheel at exactly 135 degrees, rotate the goniometer capacitor so that slot of induction ring is toward the base of the housing (this is to prevent a 180-degree error in rotational relationship of the capacitor to the tone wheel).

f. Slide a 0.02-inch feeler gage blade between the two stator plates nearest the 360-degree dial, and engage the slot in the split rotor shaft. (This will cause the rotor to position itself so that the main axis of the blades is parallel to the capacitor mounting base.)

g. With the feeler blade positioning the capacitor rotor and the dial set at 135 de-

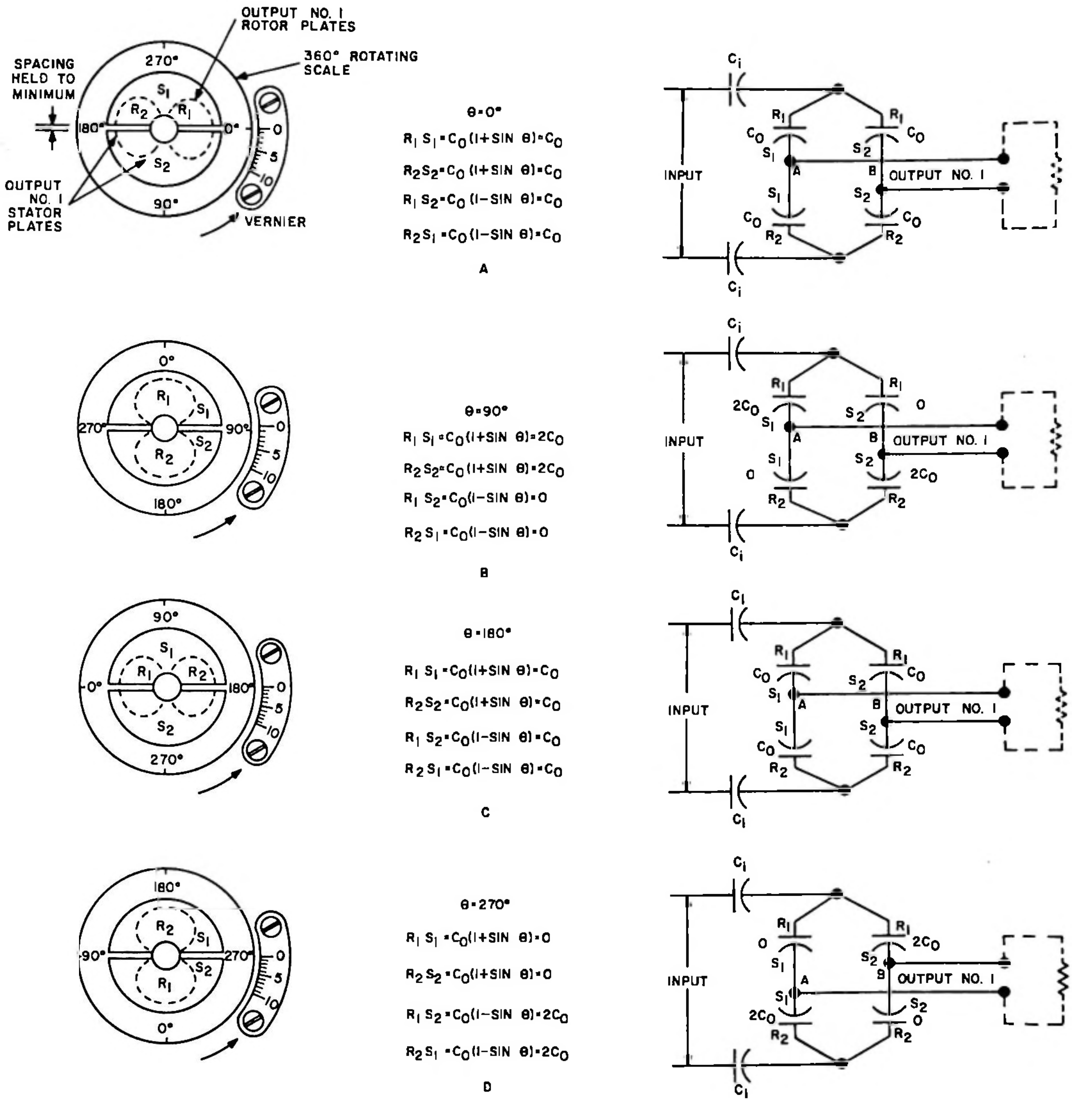


Figure 11-126AC. VHF Goniometer, Capacitor Output and Equivalent Circuits

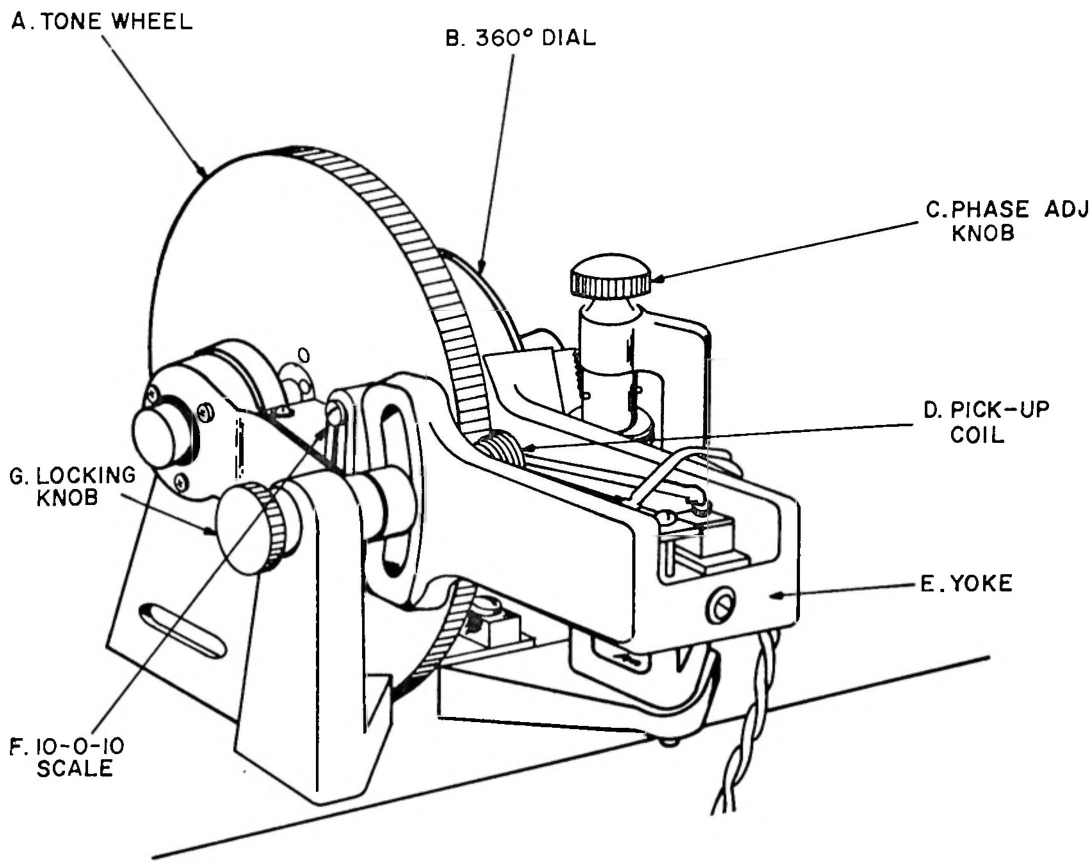


Figure 11-126AD. Close-up of a Reference Signal Generator

grees, press the coupling together and tighten the setscrews.

h. Replace the capacitor housing cover and secure the mounting screws. You have now established the proper phase relationship between the goniometer capacitor and the reference signal generator.

11-930. With the rotational relationship of the goniometer capacitor and tone wheel properly adjusted as previously discussed, the 360-degree dial can now be oriented using the following procedure:

a. Loosen the setscrews that hold the dial in place on the shaft.

b. Turn the tone wheel so that the arrow on its outer edge is directly opposite the permanent magnet of the pick-up coil.

c. Holding the tone wheel in this position, rotate the 360-degree dial so that it reads 135 degrees on the vernier.

d. Tighten the dial setscrews. You have now properly oriented the dial with respect to the tone wheel and goniometer capacitor.

11-931. To adjust the pick-up coil with respect to the tone wheel, you must perform the following procedure:

a. Terminate the output of the reference signal generator with a non-inductive 600-ohm resistor.

b. Connect a Ballentine Model 300 vacuum-tube voltmeter or equivalent across the load resistor, and apply primary power.

c. Measure the voltage across the load resistor, the reading should be 1 volt.

d. If the measured voltage is higher or lower than 1 volt, loosen the mounting screws at the bottom of the yoke assembly (figure 11-126AD, point E), and turn the adjusting screw at the end of the assembly until you obtain a reading of 1 volt. (A higher reading indicates too little spacing, and a lower reading indicates that the spacing is too great.)

e. Make sure that the pick-up coil does not touch the tone wheel.

f. Tighten the mounting screws and again check the voltage.

11-932. The pointer adjustment is made as follows:

a. Rotate the tone wheel until the dial indicates 135 degrees.

b. Align the pointer on the yoke assembly to read zero on the 10-0-10 degree scale (figure 11-126AD, point F).

c. Loosen the locking knob (figure 11-126AD, point G), and turn the phase adjustment knob (figure 11-126AD, point C) in first one direction and then the other to determine whether the pick-up coil assembly can be moved ± 10 degrees from zero, as indicated by the pointer.

d. If this is not possible, you must readjust the pointer by manipulation of its mounting screws.

11-933. If fine phase adjustment is necessary, you must loosen the locking knob (figure 11-126AD, point G) and turn the phase adjustment knob (point C) for correct reference signal generator phase. Fine phase adjustment is then made by moving the yoke assembly of the reference signal generator up or down a maximum of 10 degrees from the horizontal (zero) position. Movement of the yoke changes the position of the pick-up coil with respect to the teeth on the tone wheel, and hence changes the output phase of the reference signal generator. Next, tighten the locking knob. The phase change will then be indicated on the 10-0-10 degree scale.

11-934. MONITOR RECEIVER.

11-935. FIELD DETECTOR BALANCING. To perform field detector balancing, accomplish the following procedure while referencing figure 11-126AE:

a. Turn the proper transmitter switch to the standby position. Turn the monitor receiver power switch to the "on" position.

b. After an hour's warm-up, set the input selector switch to FS motor input No. 1 and input attenuator No. 1 to 0 decibels.

c. Adjust the detector No. 1 balance control to obtain a reading of 0 on the field intensity meter.

d. Set the input selector switch to FS

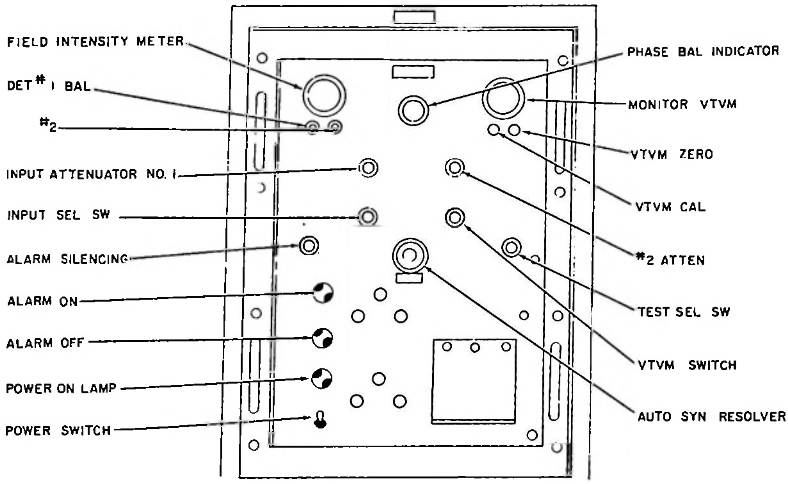


Figure 11-126AE. Typical Monitor Receiver

motor input No. 2 and input attenuator No. 2 to 0 decibels.

e. Adjust the setting of the detector No. 2 balance control to obtain a reading of 0 on the field intensity meter.

11-936. FIELD DETECTOR LEVEL. To set the output level of the field detector, perform the following procedure:

a. Turn the transmitter switches to transmit with the high-voltage power supply set for full power output.

b. Turn the input selector switch to FS motor input No. 1 and input attenuator No. 1 control to 15 decibels.

c. Turn the input attenuator No. 1 control to obtain the largest possible reading under 25-scale divisions on the field intensity meter.

d. Turn the input selector switch to FS motor input No. 2 and input attenuator No. 2 to 15 decibels.

e. Repeat step c, using input attenuator No. 2.

f. If the readings obtained in steps c and e differ by more than 20 percent, repeat steps a through f, positioning the controls as required to obtain more nearly equal readings. The meter should now read within ±10 percent of 25 scale divisions.

11-937. BEACON EQUIPMENT.

11-938. RADIO BEACONS.

11-939. General. The radio beacon transmits a signal for reception by aircraft, enabling them to determine their position. The beacon can transmit on any one of 126 different 1-megacycle channel frequencies; from 962 mc to 1024 mc and from 1151 mc to 1213 mc. It can receive on any one of 126 different 1-mc channel frequencies, from 1025 mc to 1087 mc and 1088 mc to 1150 mc. The 126 pairs of frequencies can be segregated into two groups; one using a high-band antenna and the other a low-band antenna. Thus the radio beacon can operate over one of 63 different channels. One hundred different aircraft can be serviced with information simultaneously. To provide distance information, the maximum range of the radio beacon is approximately 200 miles, and line-of-sight distance is required to provide bearing (Azimuth) information. The following elements, containing information required to determine the geographic position of the aircraft, is provided by a radio beacon:

a. Azimuth relative to the beacon referenced to magnetic north

b. Radio beacon identification

c. Distance from the beacon

11-940. The azimuth and distance information helps to determine an aircraft's position. The azimuth information is transmitted by modulating the antenna radiation pattern and transmitting bursts of reference pulse-pair groups. Bearing is determined by a comparison of the phase of the antenna modulation envelope with the time occurrence of the reference bursts. Distance information is determined by measuring the total time elapsed between the signal sent and the signal received. The radio beacon periodically transmits its identifying call in International Morse Code, thus enabling an aircraft to determine which radio beacon is transmitting on each frequency band. The characteristics of the transmitted beacon code consist of a train of 1350 double pulses per second, keyed by a coding wheel built into the radio beacon. The aircraft receiver decodes the pulses and reproduces them as identification call signals. All signals transmitted by the radio beacon are characterized by the fact that they consist of pulse pairs with 12-μsec spacing between the two pulses of the pair. The number of pulse pairs per second is the spacing between the leading edge of the first pulse of one pair and the leading edge of the first pulse of the next pair, and the spacing between pulse pairs is a characteristic of that particular signal element. However, it is the spacing of 12 μsec between the pulses of a pair that provides the aircraft with the means for distinguishing between the signal pulses from the radio beacon and any other pulses that may be present on the received radio frequency.

11-941. ANTENNA 1350-CPS OSCILLATOR. The antenna synchronization 1350-cps oscillator (shown in figure 11-125AF) is functionally a part of the antenna group and is no part of any receiver-transmitter circuits. The oscillator frequency is controlled by a tuning fork whose 1350-cps output is applied to the oscillator grid. Part of the oscillator output is used as a 1350-cps reference to check the rotation speed of the antenna; the balance is coupled to the second stage of the amplifier, a cathode follower with the input coil of the tuning fork as its cathode load. This feedback to the tuning fork results in the generation of sufficient voltage to sustain oscillation within the oscillator circuit.

11-942. The azimuth information is measured in degrees clockwise from magnetic north. This is the reciprocal of the information obtained from the modulated antenna pattern, which actually shows the direction of the aircraft from the beacon station. The azimuth information, composed of four separate components, is transmitted to the interrogating aircraft. These four components are identified as follows:

a. 15-cps amplitude modulation. The 15-cps amplitude-modulated signal will be imposed on the rf carrier energy radiated by the beacon antenna.

b. North reference burst (15-cps). A group of coded pulses is introduced into the 15-cps amplitude-modulated field of energy

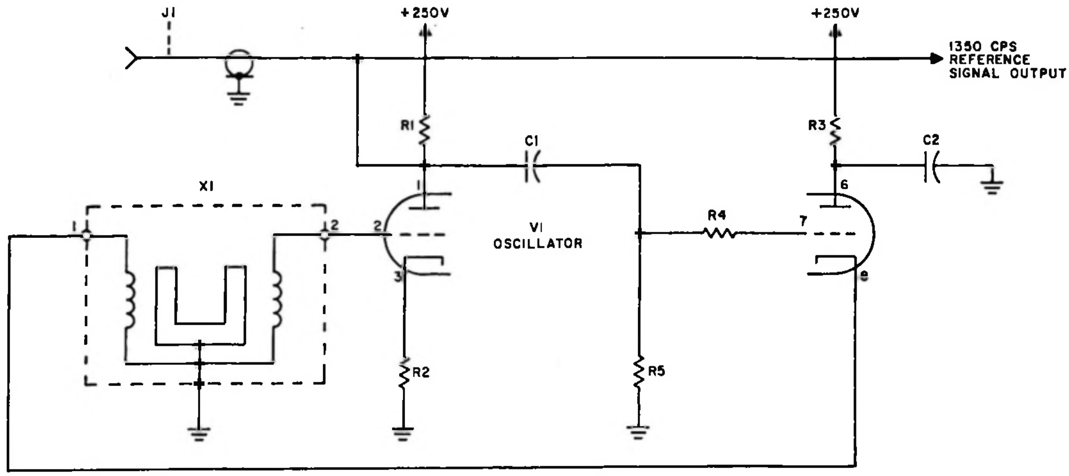


Figure 11-126AF. Antenna Synchronization 1350-cps Oscillator, Simplified Schematic Diagram

radiated by the beacon antenna. This coded burst orients the beacon's bearing information with respect to magnetic north, once during each revolution of the reflector (every time the peak of the radiation lobe points due east with respect to magnetic north). At the instant the peak of the radiation lobe points due east, a trigger pulse, generated by the antenna, is used by the radio beacons to initiate the coded burst formation. This is accomplished by having an iron slug pressed into the outer edge of an aluminum disc (which rotates with the antenna shaft) pass through a magnetic pick-up coil during each revolution. The passing of the slug through the coil changes its reluctance and thereby causes a trigger pulse.

c. 135-cps amplitude modulation. The 135-cps amplitude modulation imposed on the rf carrier is produced from having directors, spaced 40 degrees apart about the radiating element and rotating at a 15-cps rate, impose an output on the rf carrier.

d. Auxiliary reference (135-cps) burst. Iron slugs passing through a magnetic pick-up coil initiate the formation of the 135-cps reference bursts, as in the case of the 15-cps reference bursts, but at 40-degree rather than 360-degree intervals. The 135-cps amplitude-modulated component and the 135-cps reference bursts are used in a manner similar to that previously described for the 15-cps signals.

11-943. ANTENNA ADJUSTMENT. To check the speed of antenna rotation, compare the tachometer output voltage with a frequency standard. The procedure is as follows:

a. Connect a coaxial test lead from the reference signal jack on the order-indicator to the vertical input of the test oscilloscope.

b. Connect a coaxial test lead from the tachometer test jack on the coder-indicator to the horizontal input of the test oscilloscope.

c. Position the pattern in the center of the oscilloscope and adjust the pattern amplitude.

d. If the antenna is rotating at the proper frequency, a steady figure-eight pattern will appear on the oscilloscope screen.

e. If the Lissajous pattern (figure eight) is unsteady, slowly turn the speed control adjust until the pattern becomes steady. If the speed control adjustment does not steady the pattern, apply appropriate troubleshooting procedures.

11-944. KEYER ADJUSTMENT. The adjustment for the radio beacon keyer is similar to the adjustment previously discussed for the code indicating keyer.

11-945. RADAR BEACONS.

11-946. GENERAL. Fixed at known points as indicated on a map, radar beacons can provide an aircraft with positive identification of its position. A radar beacon consists of a transponder, which is a ground-based radar transmitter-receiver, and two separate antennas, one for receiving and one for transmitting. Just the opposite of a radar station, a radar beacon receives a signal first, then returns a signal to the sending source. Extraneous signals are not desirable; therefore, interrogating radars (aircraft) and the beacon transponder are provided with coding circuits, enabling them to identify each other positively. The signals must be of a certain frequency, pulse length, and spacing. Beacons need no operators because they are designed to be triggered by a coded signal received from an interrogating aircraft. When the signal is received, it will send out a coded reply, identifying the radar beacon station. This information, when received by the interrogating aircraft, will indicate the bearing and distance of the radar beacon station. Airborne radar sets thus have longer range because of the retransmission of a signal, rather than the reflection of its own, which is considerably weaker. A typical ground beacon (see figure 11-126AG) is a high-power unit which accepts interrogation pulses and transmits a range-coded reply.

11-947. BEACON TRANSMITTERS. Beacon transmitters generally use a magnetron or klystron as an rf source with a pulse power of approximately 10 kw and an rf scatter-band (including drift) of 9375 mc $\pm$ 45 mc. The pulser is usually a thyratron switch tube with a pulse repetition rate for radar of

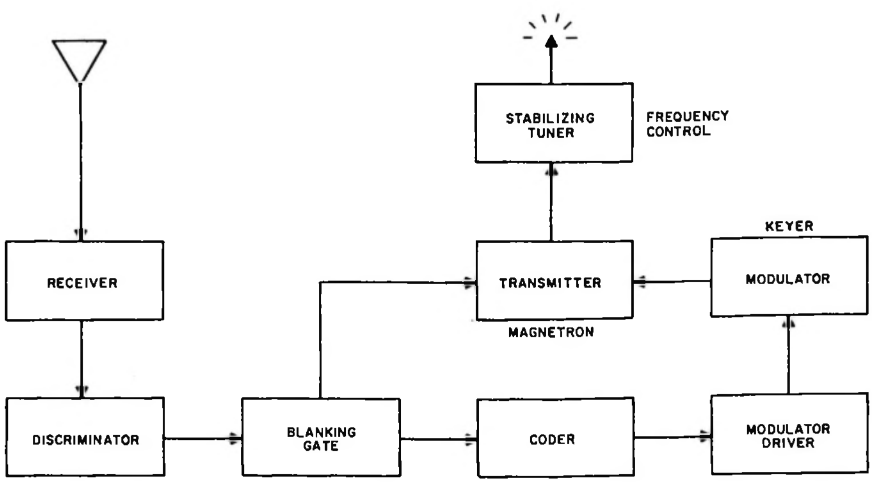


Figure 11-126AG. Block Diagram of a Typical Radar Beacon

810 pps with a duration of 0.8 μsec; and for the beacon, a pulse repetition rate of 405 pps with a duration of 2.2 μsec.

11-948. BEACON RECEIVERS. Beacon receivers are usually superheterodyne with six i-f stages and two video stages. The i-f pass band is approximately 5.5 mc and the sensitivity is 131 dbw (decibel watts) on search and 125 dbw on beacon. The special features include a separate beacon local oscillator with afc and video stretching.

11-949. BEACON ANTENNAS. The beacon antenna is usually a horizontally polarized dipole antenna with a gain of 700, and an altitude coverage of approximately 7500 ft. This azimuth is covered with a circular scan of 360 degrees, and this type of antenna can scan at the rate of 30 revolutions per minute. The elevation antenna has a tilt adjustment from —21 degrees to +3 degrees.

11-950. ANTENNA INSTALLATION. Several important factors must be taken into consideration before an antenna is mounted at a beacon site. Weather, location, altitude, and material are the most important. Antennas should be mounted on a well supported mast with guy wires holding the mast against wind velocities. A heating unit is needed in high latitudes to prevent icing. The rf lines must be as short as possible for maximum reception. This is determined by two factors: the power attenuation and the danger of long-line effects. Balancing the antenna mechanically is equally important because of the danger aircraft may encounter if given information from a slightly tilted beacon. The antenna must therefore be mounted exactly perpendicular to the earth.

11-951. BEACON CODER. The primary reason for the beacon coder is to provide a means of discriminating against signals not possessing the required characteristics programmed for the beacon site. Coding is used so that the site will not be burdened with extraneous signals. The following are the characteristics which may be coded at a beacon site:

a. Signal repetition frequency

b. Pulse spacing

c. Pulse shape (and duration)

d. Number of pulses per interrogating signal

e. Frequency (or frequencies)

When used in conjunction with the airlink equipment of a close support control set, the beacon coder provides navigation and bombing information to a remotely guided plane or missile by means of coded pulse-time-modulated commands within the maximum range of the radar set. The beacon coder automatically converts commands originating in the airlink equipment into coded pulse-time-modulated signals capable of modulating the output of the radar set. The coded radar beam provides remote control of the radar set, which is carried aboard the guided plane or missile. Thus the beacon coder with its associated equipment provides a direct ground-to-plane communications link without the use of any auxiliary communications equipment.

11-952. DELAY LINE ADJUSTMENTS. The magnetostriction delay line adjustments can be performed as follows:

a. Connect the probe of an oscilloscope to the modulator drive section of the equipment.

b. Turn the radar set control switch to the STDBY position. This means that voltages are available but have not been applied to the output equipment.

c. Connect a pulse generator, adjusted to provide output pulses of the desired width at the repetition frequency of the equipment being tested, to the input of the equipment under test. A group of negative pulses should be seen followed by a group of undesirable reflections.

d. To minimize the amplitude of the undesirable reflections, start at one end of the modulator drive section, and while observing the oscilloscope, turn each adjusting screw on the bottom of the front panel of the modulator drive section until the lowest possible amplitude of reflections is obtained for the final setting of each screw. Following the sequential adjustment, be sure that all screws are seated securely.

e. While you observe the oscilloscope screen, loosen the magnet mounting screw and orient the magnet so that the maximum negative amplitude is obtained; then secure the magnet firmly in this position. If the second pulse is much lower in amplitude than the other pulses, reverse the polarity of the magnet. Upon completion of this adjustment, place glyptal on the adjustment screws.

11-953. PULSE LENGTH DISCRIMINATOR. The pulse length discriminator, shown in figure 11-126AH, has three stages: a pulse integrator, a short pulse shaper, and a long pulse shaper.

11-954. This pulse integrator converts the negative video pulses into positive sawtooth waveforms. It operates with a small negative self-bias, obtained through cathode resistor R2; consequently the quiescent plate voltage is low, and capacitor C2 is

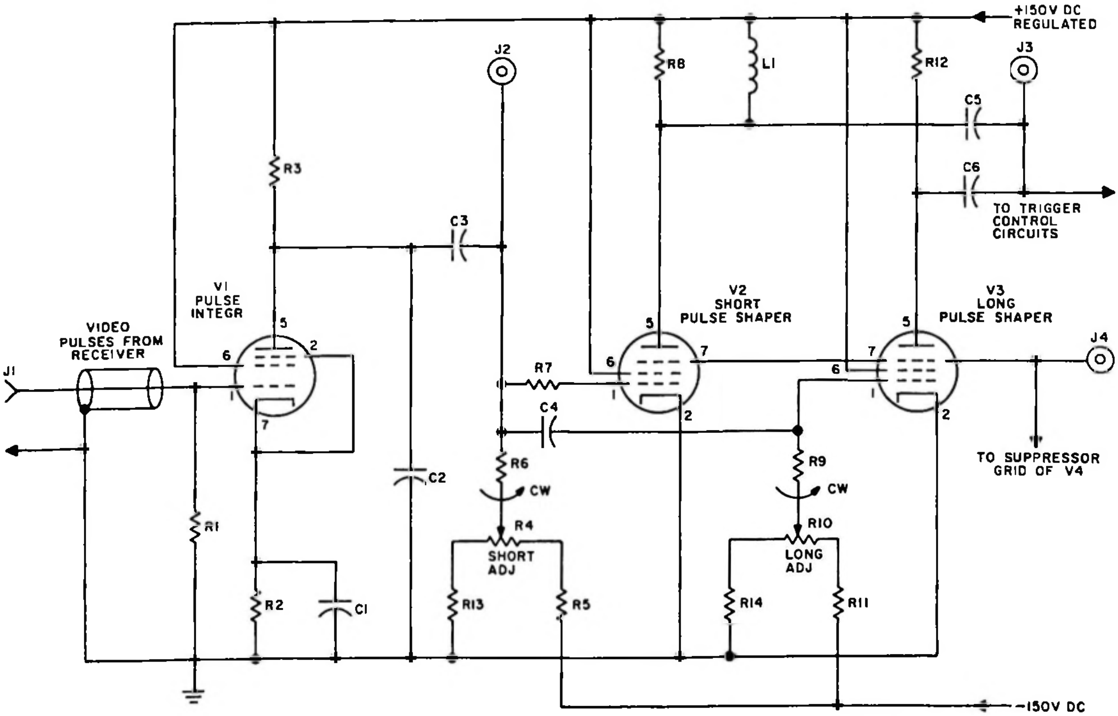


Figure 11-126AH. Pulse Length Discriminator, Simplified Schematic Diagram charged to a low voltage. The negative video pulses cut off the V1 plate current, and C2 starts to charge through plate load resistor R3 toward +150 volts. Since the charging circuit rc time constant is approximately 6.6 microseconds and the video pulse duration is normally 2.35 microseconds, the voltage across C2 is a sawtooth of approximately constant slope (6 volts per microsecond); consequently, the peak amplitude of the sawtooth voltage is proportional to the duration of the video pulse applied to the V1 control grid. When the video pulse ends, the low dc plate resistance of V1 rapidly discharges C2 and prepares the stage for reception of the next pulse.

11-955. The short pulse shaper, V2, produces a negative pulse followed by a positive pulse when the video pulse length exceeds 2.0 μsec. The sawtooth output voltage of V1 is applied to the grids of V2 and V3. The V2 plate circuit contains inductor L1 shunted by R8. Tube V2 is biased beyond cutoff; the bias is adjusted by R4 so that a 2.0-μsec sawtooth pulse from V1 has just sufficient amplitude to raise the V2 control grid above cutoff. The quiescent voltage across L1 is zero; when V2 starts to conduct, the plate current through the tube causes the plate voltage to fall. When the sawtooth ceases, the V2 grid is driven beyond cutoff again, the plate voltage rises rapidly and, because of inductor L1, overshoots its quiescent value, thereby forming a positive trigger. Resistor R8 rapidly damps the oscillations; consequently only one cycle occurs.

11-956. The long pulse shaper, V3, produces negative pulses. These pulses begin 2.7 μsec after the start of the video pulse if the video pulse length exceeds 2.7 μsec. The bias of V3 is adjusted by means of R10 so that a 2.7-μsec sawtooth will just raise the control grid voltage to cutoff. Hence, if the sawtooth duration exceeds 2.7 μsec, V3 starts to conduct. Therefore, the portion of the sawtooth input that exceeds the cutoff level appears inverted at the V3 plate. This voltage is mixed with the V2 output, and the resultant voltage is applied to the V4 control grid. Thus, when the length of the video pulse applied to the V1 control grid is less than 2.0 μsec, no output is applied to V4 because V2 is not brought out of cutoff. When the video pulse length is between 2.0 and 2.7 μsec, only V2 is brought out of cutoff, and a negative pulse followed by a positive pulse is applied to V4. When the video pulse length exceeds 2.7 μsec, the sum of the V2 and V3 outputs is not positive, because the negative output of V3 has a greater amplitude than the positive output of V2. Since the trigger control circuits can be triggered only by positive pulses, it is evident that this can be accomplished only by pulses having a length between 2.0 and 2.7 μsec. The pulse length discriminator circuits are so arranged that a trigger control circuit output can occur only when the mixed output of V2 and V3 contains positive pulses. This occurs when the pulse length lies between 2.0 and 2.7 μsec.

11-957. PULSE LENGTH DISCRIMINATOR ADJUSTMENTS. The pulse length discriminator adjustments can be performed as follows:

a. Turn switches to the test positions.

b. Turn both the LONG ADJ and the SHORT ADJ controls fully clockwise.

c. Connect the output of an rf pulse generator to the input of the pulse length discriminator, and decrease the output pulse length while increasing the input signal strength to a level of 5 db above the equipment sensitivity.

d. Plug a headset into the output headset jack of the pulse length discriminator.

e. Slowly turn the SHORT ADJ control counterclockwise until a pure, uninterrupted tone is heard in the headset.

f. Increase the generator output pulse length.

g. Slowly turn the LONG ADJ control counterclockwise until the tone in the headset ceases.

h. Decrease the generator output pulse length and repeat step e .

i. Slowly vary the generator output pulse length. Steady triggering should be heard in the headset except for the low and high limits, which will be unsteady. This is a normal condition.

j. Return all switches to the normal operating positions. The adjustment is complete.

11-958. DIRECTION-FINDER EQUIPMENT.

11-959. GENERAL.

11-960. RADIO DIRECTION FINDERS. Radio direction-finding equipments are used both for homing and for obtaining position fixes in navigation. The equipment basically consists of a directional antenna, an antenna-to-receiver coupling unit, a receiver, and some form of azimuth indicator. One of the most important items of a radio direction finder is the antenna, which must have directional properties. The versatility of the equipment is derived by having an antenna which can be rotated to detect the direction of incoming signals. This involves the use of special types of signal coupling units. The receiver is conventional in design, and its output is applied to the indicator, which provides a visual indication of the azimuth in degrees, either directly or indirectly.

11-961. DIRECTIONAL ANTENNAS. The simplest type of directional antenna is a loop antenna, which may take the form of a circle, a square, or a triangle. All forms of the loop antenna have a field pattern which resembles the figure eight, as shown in figure 11-126AI. The figure-eight pattern has two maximum and two null (minimum) signal points, with the nulls being perpendicular to the plane of the loop, as shown in the figure. The null points, which are the easiest to detect, are used to determine the direction of the received signal; however, since the loop antenna has two such points, it would be impossible without special provision to determine whether the signal is arriving from one direction or from a direction 180 degrees removed. This ambiguity is extremely undesirable, and is eliminated by using a unidirectional antenna, known as a sense antenna, in conjunction with the loop. The signals received by both antennas are applied to the receiver, and when the physical arrangement and relative size of the two

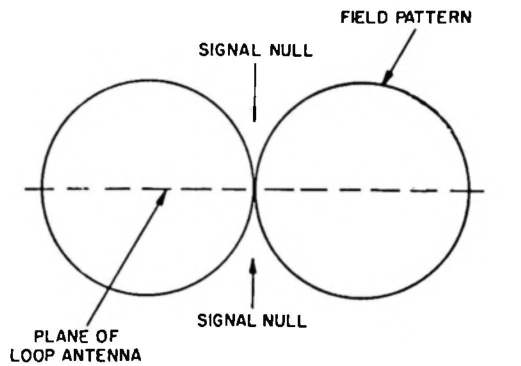


Figure 11-126AI. Field Pattern of Loop Antenna are correct, the receiver antenna field pattern will have the shape shown in figure 11-126AJ. This field pattern, known as a cardioid, is a result of the combination of the two antenna field patterns and has a single null 90 degrees displaced from the figure-eight nulls. (The two halves of the figure-eight pattern of the loop antenna are 180 degrees out of phase with each other, and the pattern of the sense antenna is in phase with one of them, the right-hand one in figure 11-126AJ; the resulting addition of the signals produce the cardioid pattern.) This condition exists only if the lobes of the sense and loop antennas are of the proper relative magnitudes. If the sense antenna is more effective than the loop, a modified cardioid with no complete null is produced, as shown in part A of figure 11-126AK. If the loop antenna is more effective, a modified cardioid with a small lobe at its null point is produced, as shown in part B of figure 11-126AK. Several types of errors are encountered with the loop-sense antenna direction-finder equipment.

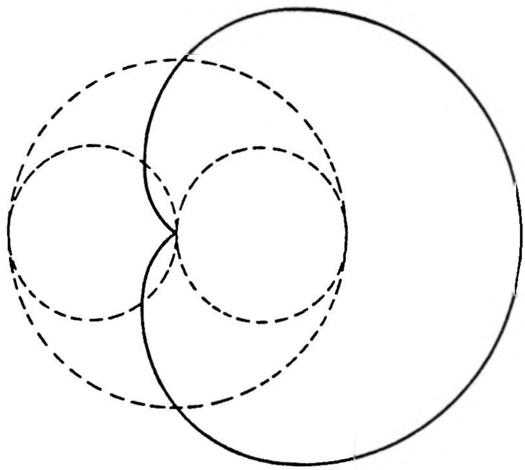


Figure 11-126AJ. Cardioid Pattern Resulting from Combination of Loop and Sense-Antenna Field Patterns

11-962. Except for the nonuniform field pattern of the sense antenna, which may be caused by nearby objects, all errors can be

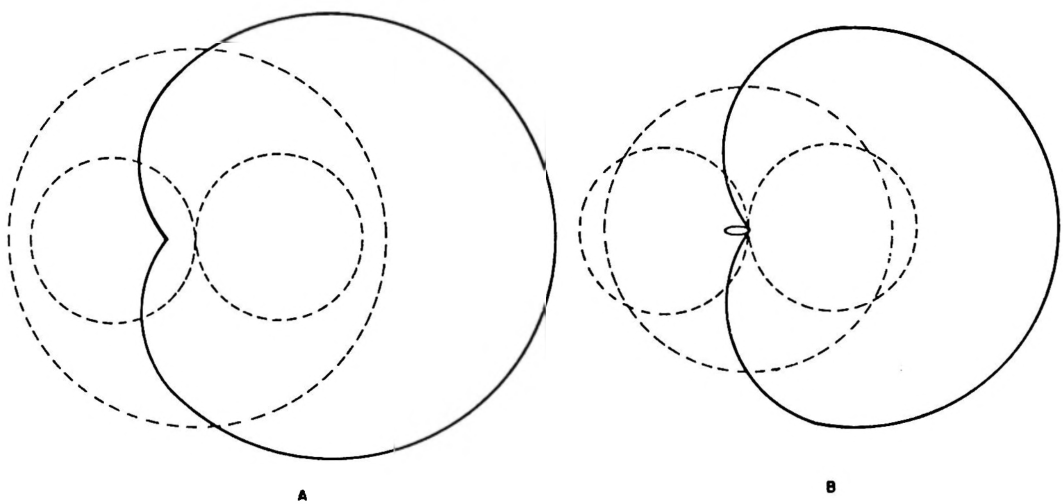


Figure 11-126AK. Field Patterns Resulting from Too Much (A) and Too Little (B) Sense-Antenna Voltage attributed to the loop antenna. Irregularities in the winding of the loop antenna and in the relationship between the loop and nearby objects may introduce errors into the readings obtained with the direction finder. These irregularities cause a small sense voltage to be developed in the loop, which produces an error known as antenna effect. This sense voltage adds to the signal received by the loop antenna; if the sense voltage is in phase with one lobe of the figure-eight field pattern, that lobe will be larger than the other, and the nulls will be less than 180 degrees apart. If this sense voltage is 90 degrees or 270 degrees out of phase with the loop voltage, the nulls are filled and a sharp null is not obtained. A similar effect will be produced by metal objects in the vicinity of the radio direction-finder antenna. In this case, the reflection of the incoming signal by the metal causes a signal to be produced at some random phase relationship with the direct signal. Since reflection produces random polarization of radio waves, the reradiated signals from nearby objects can also cause errors in another way. The field pattern of the loop antenna shown in figure 11-126AI is produced when vertically polarized signals are received. If only horizontally polarized signals are received, the null points will be shifted 90 degrees, or they will lie parallel to the plane of the loop. When reradiated signals are received, some horizontal components may be present; if so, a shifting of the null points will result. The amount by which the nulls are shifted depends on the degree to which the signals are horizontally polarized. The polarization change caused by ionospheric reflections is called night effect, and that caused by the polarization of an airplane's antenna is called airplane effect.

11-963. Another source of error in a radio direction finder using a loop antenna can be attributed to asymmetry in the arrangement of the coils of the loop. Any asymmetry of the windings will distort the figure-eight field pattern produced by the loop, and will thus change the position of the nulls or obscure them. Asymmetry to ground can be eliminated by an electrostatic shield, consisting of a metal case designed to enclose the loop almost completely. The halves of the shield are separated by an insulator at the top of the loop, to prevent the shield itself from functioning as a single-turn loop antenna. In addition to eliminating asymmetry to ground, this shield eliminates asymmetry to nearby metal objects except those which are very near the insulated section of the shield. Since the shield is metal, it shorts out precipitation static caused by the striking of the antenna by charged particles, and provides mechanical protection to the wires composing the loop.

11-964. Next to the loop, the most common direction-finder antenna is the Adcock antenna, the basic form of which is shown in figure 11-126AL. The field pattern which is developed about the vertical elements resembles that of the loop antenna, and the sources of error are very similar; however, night effect is greatly reduced. The horizontal members of the antenna are located at its center, and are arranged so that the horizontal signal components due to downward-moving sky waves cancel. Since the vertical members cannot respond to horizontally polarized waves, there is little error in bearing readings because of skywave horizontal polarization. Despite the

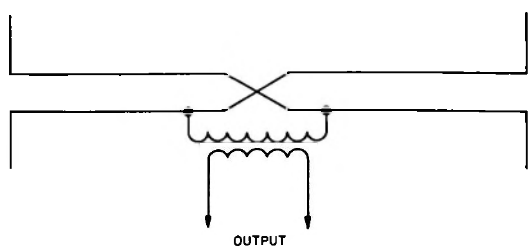


Figure 11-126AL. Basic Adcock Antenna

best shielding and balancing available, however, there is always some residual pickup from nearby metal objects and ground.

11-965. The grounded Adcock antenna, shown in figure 11-126AM, eliminates nearly all horizontal polarization errors. In this form, each vertical section of the antenna is separately coupled to the receiver. The mutual inductance between the vertical sections provides a high impedance to effectively eliminate signal voltages received by the horizontal sections because of horizontal polarization. With the addition of electrostatic shields between the coupling coils, an antenna theoretically free of polarization error can be constructed. Practical antennas of this type have been designed with a polarization error of less than 1 degree. There is little difference between the Adcock and loop antennas regarding factors such as ambiguity, sense, and errors other than polarization error. The Adcock antenna, however, must occupy a larger space to provide the same sensitivity, since it is composed of a single wire, whereas a loop antenna is composed of a coil of wire. Crossed-loop antennas, or crossed-Adcock antennas, which consist of two identical antennas mounted so that their field patterns are perpendicular to each other, are sometimes used in direction finders. These crossed antennas are normally stationary, and various methods of electrically phasing the antennas to achieve simulated rotation are used to obtain the azimuth of received signals. Since crossed antennas permit you to differentiate among four nulls, a more complicated sense mechanism is necessary to determine direction. The true and reverse azimuth are separated from the azimuth at right angles by a balancing arrangement, after which a typical sense antenna eliminates the ambiguity.

11-966. Another type of direction-finder antenna, which is composed of two identical loop antennas, is known as the spaced-loop antenna. In this type, the loop antennas are mounted in a fixed position with respect to each other, but the extire unit is rotatable. Two forms of the antenna are used, as shown in figures 11-126AN and 11-126AO. In the type shown in figure 11-126AN, the planes of both loops are parallel and aligned. This arrangement produces a field pattern which is similar to that of the single loop shown in figure 11-126AI. The advantage of this antenna is that night effect is eliminated, because horizontally polarized, downcoming waves from the ionosphere induce voltages in the two loops which are equal in amplitude and opposite in phase, and therefore cancel. The major disadvantage of the antenna is that, because of the space separa-

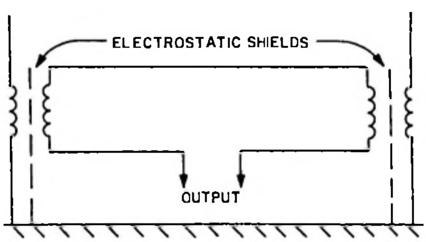


Figure 11-126AM. Grounded Adcock Antenna

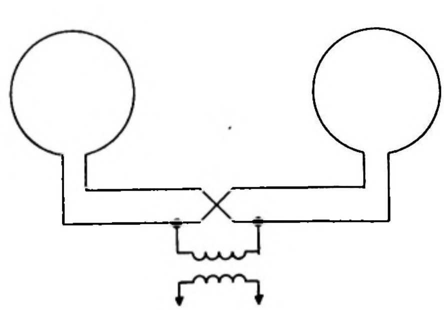


Figure 11-126AN. Spaced-Loop Antenna (Two Identical Loop Antennas)

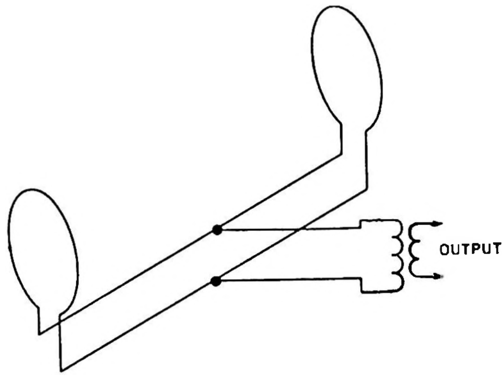


Figure 11-126AO. Coaxial Spaced-Loop Antenna

tion, much care must be taken to obtain symmetry in the arrangement.

11-967. With the coaxial spaced-loop antenna shown in figure 11-126AO, the effects of horizontal polarization are not cancelled, but are rather effectively used to provide the signal nulls at the correct place. In this type of antenna the planes of both loops are also parallel, but they are not aligned, and the true nulls are parallel to the plane of the loops. Figure 11-126AP shows the effect of horizontal polarization upon the field pattern. Notice that even though the field pattern changes with polarization changes, the true signal null points are always in the correct direction, whereas the incorrect nulls shift and even disappear when the horizontal component is great enough; therefore, with this antenna, an ordinary sense antenna is all that is necessary to determine the correct bearing to a signal source. The spaced-loop antenna (either type) is generally used when good direction-finding accuracy is required and little antenna space is available. If the over-all width of a loop antenna, or the distance between the elements of an Adcock antenna, is an appreciable part of a wavelength of the received signal, the null points produced will deviate from the direction of the transmitting station, and will change with signal-frequency changes. As the frequency increases, the antenna will become a greater portion of a wavelength.

11-968. ANTENNA-COUPLING METHODS. In direction finders that use rotatable antennas, special methods must be employed to connect the antenna to the receiver. One method is to use slip rings; however, slip rings often introduce a

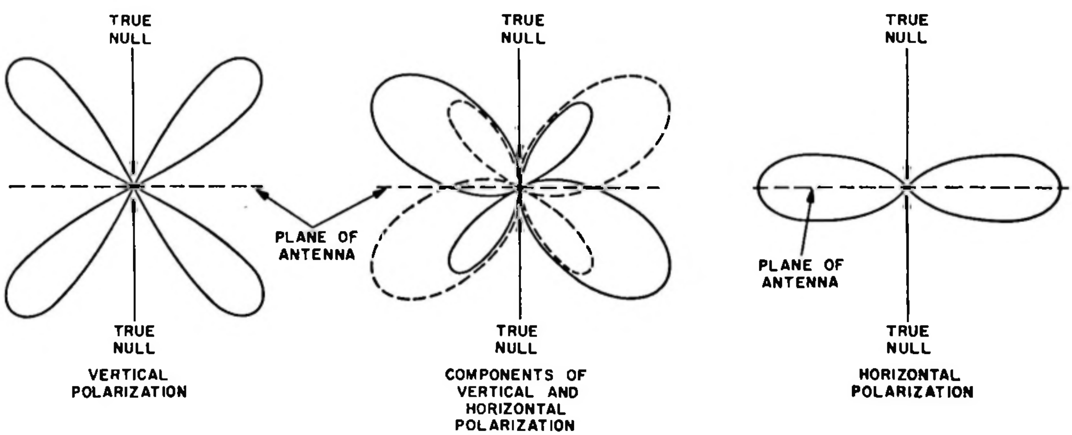


Figure 11-126AP. Coaxial Spaced-Loop Antenna Field Patterns great deal of noise, in addition to providing another possible source of unbalance in the antenna equipment. Another method is the use of pigtails, or flexible-wire connections; however, their use allows stray pickup and limits the rotation of the antenna, since the wire will wrap around the shaft and break if the antenna is turned too far. Nevertheless, pigtails do not introduce noise and are therefore often used in preference to slip rings. Electronic coupling devices, such as the rotating transformer shown in figure 11-126AQ, are sometimes used. Since the two windings are coaxial, the degree of coupling is constant, regardless of the direction in which the antenna faces. The secondary of the transformer can be tuned, along with the other parts of the receiver rf strip, to cover a wide band of frequencies with equal input. At higher frequencies a rotatable capacitor may be used, since a small value of capacitance is sufficient for adequate coupling.

11-969. When a fixed directional antenna such as a crossed-loop or crossed-Adcock antenna is used, the method of determining the direction of an incoming signal may be accomplished in the actual coupling device. The most commonly used device is an inductive goniometer. As illustrated in figure 11-126AR, an induction goniometer consists of two fixed primary windings, oriented at right angles to each other, and a rotatable secondary winding. Each primary is connected to one of the antennas. Since each antenna receives a signal whose strength is proportional to the direction of the signal source from the antenna, the secondary will perceive a null when it is in such a position that it receives equal and opposite signals from the two primaries. Because the amount of coupling between each primary and secondary is dependent upon the position of the primary relative to the secondary, signals of equal strength can be received from the antennas by properly orienting the secondary. The angle of orientation depends upon the direction of the incoming signal. For example, if no signel is induced in one antenna, the maximum signal will be induced in the other, so that if the secondary is aligned with the primary connected to the antenna receiving no signal, a null is produced in the output, since the secondary has no inductive coupling with the other primary. If a signal arrives at a 45 degree angle to the antennas, each antenna will have the same output, and the secondary must be positioned at a 45 degree angle with respect to each primary to receive equal and opposite outputs from the

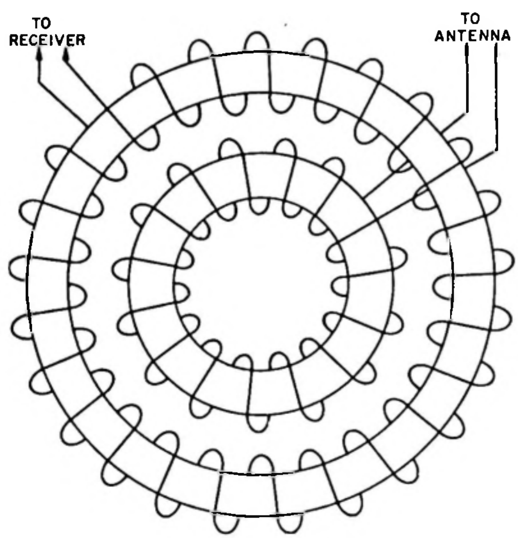


Figure 11-126AQ. Rotating-Transformer Coupling Device

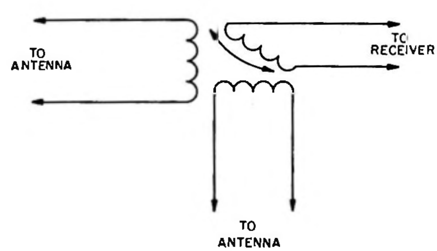


Figure 11-126AR. Inductive-Goniometer Coupling Device two antennas. Regardless of the received-signal direction, the secondary can be positioned to produce a null, and this position will be directly determined by the signal bearing. If a visual direction indicator is linked with the secondary, a reading in degrees can be obtained. The secondary may be rotated either manually or automatically. A capacitive goniometer, which uses capacitors instead of inductors, but which operates on the same principles as the inductive goniometer previously discussed, is also used. The inductive goniometer is employed for high-frequency signals, and the capacitive type for low-frequency signals, where sufficient capacitive signal coupling can be obtained.

11-970. Another method of obtaining simulated rotation of fixed antennas is to employ electronic switching units. The results obtained by this method are similar to those using an inductive or capacitive goniometer except that no mechanical motion is required; hence, the term electronic goniometer. Even though it contains tubes, resistors, and capacitors, this goniometer may still be considered as a coupling unit, since its output is applied to the first rf stage of the receiver. Figure 11-126AS shows the basic requirements for an electronic goniometer. The signal received by a pair of vertical antennas (a loop antenna or an Adcock antenna may also be used) is applied to the electronic switch circuit. At the same time,

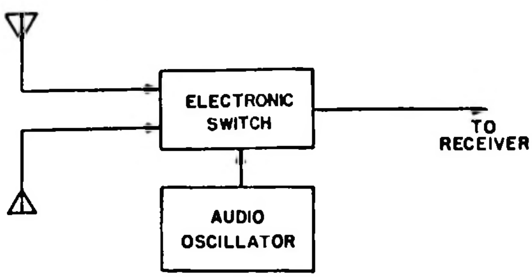


Figure 11-126AS. Basic Elements of an Electronic Goniometer

an audio signal is applied to the electronic switch from an audio oscillator. Effectively, the action of the switch is to connect the signal from one of the vertical antennas to the receiver during one half of the audio cycle, and the signal from the other antenna during the other half of the cycle, the signal strength being dependent upon the audio-signal level at any instant. Since two pairs of vertical antennas are necessary, a second circuit identical to the one just described is employed. The only difference in the operation of these two switch circuits is that, since the second pair of antennas provides a figure-eight field pattern which is at right angles to the first, the audio oscillator must provide signals that are in quadrature phase to the first. The outputs of both switch circuits are combined in the receiver input, and the result is an apparent rotation of the antenna equipment at the audio-frequency rate. The rotating effect can be explained as follows: First, arbitrarily assign the directions N and S to the two halves of one figure-eight pattern, and E and W to the two halves of the other figure-eight pattern. Then assume that the action as perceived at the input of the receiver begins when the signal from N is maximum. As the N signal falls toward zero, because of modulation by the audio signal, the E signal rises from its zero point toward maximum and reaches maximum when N reaches zero. As the E signal passes its maximum and starts toward zero, the S signal begins to increase, etc, until the cycle is completed. The result is that signals can be received from all directions, as from a nondirectional antenna, with approximately the same strength. Since the effective rotation is produced by electronic circuits, high rotational speeds can be obtained; by using an oscilloscope as an indicator, with the deflection circuits operated in synchronism with the rotating field pattern, a visual presentation of the received signals can be obtained. The electronic switching circuit is a balanced modulator.

The audio-oscillator frequency varies widely with the type of equipment used, but is generally between 40 cps and 100 cps. Goniometers of all types are subject to all of the direction-finder errors of the particular kind of antenna used, and also introduce eight-wavelength errors. These errors are the result of the nonuniform flux fields in the stationary windings of inductive goniometers and the nonuniform dimensions of the antenna with respect to ground.

11-971. DETECTION METHODS. Any standard receiver may be used for radio direction-finder application to find the two nulls of a direction-finder antenna, by aural (hearing) means. If a visual indicator is available, it is possible to determine directly the direction of the incoming signal. To receive cw signals, the receiver must contain a beat-frequency oscillator or an audio-injection oscillator to produce the audible note. The audio-injection oscillator introduces an audio signal of the desired frequency into one of the i-f or rf stages. This method has the advantage of producing an output of constant tone, regardless of drift of the input signal; however, it also allows some signal to be present in the output, even at the null, because the audio signal modulates incoming noise. This residual signal, which is a function of noise, can be reduced to a minimum by coupling directly into the af stages an audio signal 180 degrees out of phase with, and equal in amplitude to, the residual signal; however, since the noise is always fluctuating, perfect cancellation cannot be attained. One of the most important considerations in any direction-finding receiver is the careful shielding of circuits, especially the input circuit and the first few amplifier stages, because any stray signal introduced at these points will produce an error at the indicator by acting in the same manner as the signal from the sense antenna. Wherever adequate shielding cannot be attained, the circuits must be balanced to avoid introducing errors in the output. For certain applications when two or more antennas are used, there must be two or more equivalent amplifier circuits in a receiver to amplify the signals separately before they are combined. The sense-antenna signal may be amplified and changed in phase before being combined with the loop signal, or, when fixed antennas are used, the inputs from the two antennas may be amplified separately and combined in the indicator unit.

11-972. BEARING INDICATORS. After a signal is detected and changed to interpretable information, some means must be devised to present this data to the operator. The simplest form of indicator is a headset (when used in conjunction with a magnetic compass). To find the azimuth to a distant transmitter, the direction-finder receiver is tuned to the transmitter frequency, and the directional antenna is rotated to a position of signal null in the headset. This method is quite cumbersome, however, since the directional antenna must be mounted in a fixed position and the entire antenna station must be turned to obtain the null. The bearing to the station can then be obtained by reference to the magnetic compass heading. (At frequencies over 100 megacycles, the maximum signal may be sharper than the null, in which case the maximum signal point is used to determine the direction of a radio signal.) A more practical method of determining direction is to use some form of visual indicator which is a part of the direction-finding equipment.

11-973. In navigation, several meter arrangements are used for indicating purposes. One type is the left-right indicator, which employs a single-pointer, zero-center meter. The direction of pointer deflection depends upon whether the aircraft heading is to the right or left of the bearing to the transmitting station. If crossed antennas are used, the signal received by each antenna is amplified and detected separately. Then the two signals are combined in the indicator circuits so that the strongest signal will deflect the meter. If the signals are of equal strength, the meter pointer will, of course, remain centered. In some equipments a crossed-pointer meter is used to indicate signal direction. Two signal-amplifying channels are incorporated in the receiver used with this type of meter, and each channel receives a signal from a separate antenna. When the inputs to the indicator from the two channels are equal, the two pointers cross on the vertical center line of the meter, thus indicating that the aircraft is on-course. An increase in signal strength will deflect both pointers equally upward, and a decrease will deflect them equally downward, with the pointers still crossed on the vertical on-course line. If the aircraft drifts off-course, one of the amplifier channels will provide a stronger signal than the other, and the crossover point of the pointers will be on the side of the stronger signal. Two electron-ray tuning-indicator tubes (commonly referred to as magic-eye tubes), or one tube of the dual type, can be used instead of the cross-pointer meter previously discussed. The two shadows serve the same function as the pointers of the meters. When the received signals are equal, no shadows appear on the faces of the tubes; however, when one signal becomes weaker, a shadow appears on the corresponding tube face, thus indicating the direction in which the aircraft has drifted off-course. A cathode-ray tube can also be used as an indicator when two signal channels are used. The two signals are shown as parallel lines (either vertical or horizontal) on the face of the tube. If the signals are equal in strength, the lines appear equal in length, but if one signal is stronger, this signal produces more deflection of the electron beam and, therefore, a longer line. The navigator is thus provided with an indication of the direction in which his craft has drifted off-course. The cathode-ray tube, used in conjunction with the electronic goniometer previously discussed, presents a visual indication, as illustrated in figure 11-126AT. In order to achieve this indication, which shows the true and reverse direction of the signal directly in degrees, three signals must be applied to the deflection circuits of the tube: a sweep signal (200 kc is commonly used), the audio signal which causes the electronic rotation of the antenna equipment, and the amplified output signal of the receiver. The amplitude of the sweep signal, which moves the beam from the center of the tube toward the outer edge, is dependent upon the amplitude of the signal from the receiver. The rotation of the sweep about the face of the tube is dependent upon the audio signal, which rotates the indicator pattern to correspond with the electronic orientation of the antenna equipment. As the antenna is rotated through the direction from which a signal is received (or at the 180 degree point), an output from the receiver is obtained. This output is applied to the indicator, and the sweep is

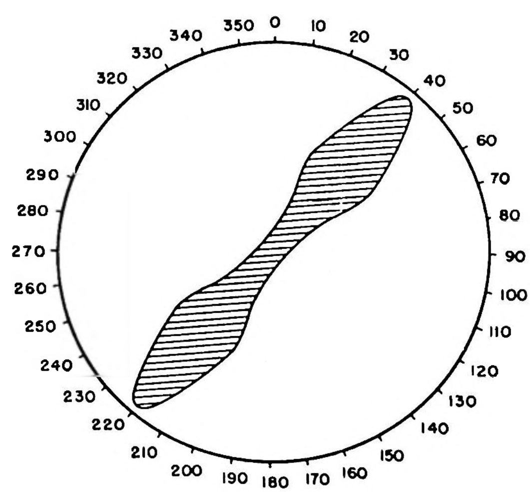


Figure 11-126AT. Cathode-Ray Tube Bearing Indicator deflected toward the edge of the tube. As illustrated in figure 11-126AT, the points of maximum sweep deflection indicate both the direction to, and the direction away from, the transmitter. This 180 degree ambiguity is eliminated by incorporating a blanking circuit to prevent the display of the false bearing indication.

11-974. Another type of device used to obtain bearings directly in degrees is the automatic direction finder, or radio compass, in which a loop (or Adcock) antenna is automatically turned to the direction of the signal being received. This type of direction finder always seeks the true bearing - never the 180 degree ambiguous bearings. The input circuits of the unit are very similar in operation to the electronic goniometer previously discussed, differing mainly in the use of a rotating loop and sense antenna in place of four fixed vertical antennas. An audio frequency is used to operate a balanced modulator (electronic switch), as in the electronic goniometer. When the loop is faced to one side of the true null point, a loop signal is received. This signal is applied to the balanced modulator, where it acquires the audio component as modulation. Because the loop signal is applied to the grids of the balanced modulator in parallel, and is obtained from the plates in push-pull, the loop signal of one half of the audio-modulated cycle is reversed in phase. The output of the balanced modulator is applied, in conjunction with the sense-antenna signal, to the input of the receiver. The sense signal is in phase with the loop signals of one half of the audio signal cycle, and out of phase with the signals of the other half-cycle. At the output of the receiver, the audio component is filtered out and applied to a pair of control tubes. Because the polarity of the audio signal is established by the sense antenna, only one control tube conducts, and that tube applies a voltage to a reversible motor. This motor, which is connected mechanically with the loop antenna, rotates in the proper direction to turn the antenna toward the null. When the null is reached, the balanced modulator ceases to function and the motor stops. When the loop is faced toward the opposite side of the null, the sense signals add to the loop signals of the other audio half-cycle, and the second control tube conducts. This causes the motor to rotate the loop in the reverse direction. A meter which has a 360 degree scale is connected to the loop antenna through a synchro arrangement, to provide a direct indication of bearing, in degrees, with respect to craft heading. The discussion above does not cover all possible combinations of the components used in direction-finder equipments, but the representative equipments included will give you an idea of the more important combinations.

11-975. Radio range is used principally for the homing of aircraft. The beam produced by the station of departure is followed until the beam of the next radio-range station is crossed. If this second station is the destination, its beam is followed to the airport, but if the destination is beyond the second station, the bearing of the second station is followed until the beam of a third station is crossed. For long-distance flights this process is repeated until the destination is reached. A navigator can obtain a position fix by taking the radio bearing of a second radio station (or obtaining a second bearing by other means) and plotting the intersection of this bearing with the radio-range bearing. Where the beams of two radio ranges cross, a position fix can be obtained if two receivers are used (one tuned to each range station). When both receivers produce a steady tone, the aircraft is over the cross point of the two radio ranges. Marker beacons are incorporated in radio ranges to enable the navigator to determine locations more accurately. These are located at strategic points along the range to provide absolute identification of these points. The main advantage of the radio range is the small amount of equipment required in the aircraft. A radio receiver with a speaker or headset to give aural indication is all that is necessary. On the ground, one or more transmitters and an array of antennas are required, but since the main consideration is the weight added to the aircraft, the weight of the ground equipment is not important. Since a different frequency is used, the marker beacons require the use of another receiver in the aircraft, but all beacons of this type (because of their field pattern) are operated on the same frequency; therefore, a simple, fixed-tuned receiver is employed. On the ground, a separate beacon transmitter is required at each point to be identified. The radio-range beams are produced by the interference between two radio signals transmitted at the same frequency. Actually, these two radio signals are produced by one transmitter, which is alternately switched from one pair of vertical antennas to another. Each pair of vertical antennas produces the same type of figure-eight field pattern as that produced by the loop antenna. (In some cases loop antennas have been used at ground radio-range stations.) The over-all field pattern produced by both pairs of antennas is two crossed figure eights. The transmitter switching is such that one pair receives a dash followed by a dot (Morse code N), while the other pair receives a dot followed by a dash (Morse code A). The timing of the A and N transmission is such that in the region of signal interference where the signals are of equal amplitude a steady tone results. As shown in figure 11-126AU, the region of steady tone is a beam, which may be from 1 degree to 3 degrees wide. On either side of the beam there exists a twilight area in which one or the other of the letters is received against a tone background. The twilight zones and the beam together are approximately 9 degrees wide. When flying the radio range, the navigator could

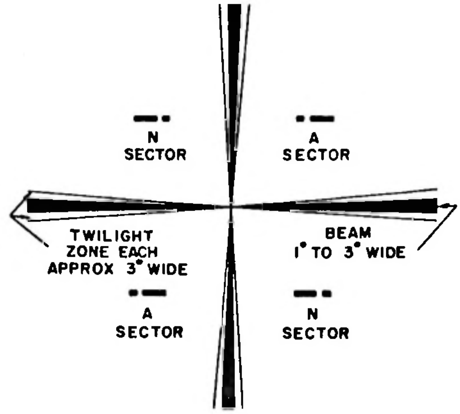


Figure 11-126AU. Radio-Range Beam Pattern

simply proceed along a course which would maintain a steady tone; however, for reasons of safety, the normal procedure is to fly at the right edge of the beam, where a slight indication of one letter is heard above the steady tone. False beams are often produced in mountainous terrain as a result of the refraction of radio waves passing near obstructions. If the signals from both the A and N antennas are refracted, the beam may actually curve around a mountain. If the obstruction is not near the cross of the A and N signals, some signals from the antenna radiating N signals may be refracted into the radio-range sector which normally contains only A signals. If this happens, the steady tone will be heard in the headset even though the aircraft is not on-course. These false beams (which often branch away from the proper beam), if followed, can result in serious consequences, and are therefore shown on the radio-range maps to ensure that the pilot will be aware of their existence. If a pilot becomes lost as a result of following a false beam, static, missing a marker beacon, etc, the beam can be located by following a standard flight procedure. However, this flight procedure, which is given in the Federal Air Aeronautics (FAA) Handbook of Air Navigation, is quite time consuming, especially when the range station is at a considerable distance. If the aircraft is equipped with a radio compass or if the pilot knows his approximate location, finding the beam can be much easier. Using a radio compass, the pilot can determine the azimuth of the radio-range station, and, by plotting this bearing on a radio-range chart of the area, the course to the nearest beam can be established. When the radio-range beam is being flown, direction can be checked by determining on which side of the aircraft each letter of the pattern is located, and whether the range station is approximately at the bearing expected when the beam is reached. The earlier radio-range stations interrupted transmissions at regular intervals to give weather and other pertinent flight information. This was not an entirely satisfactory arrangement, however, because the interruption could occur when a navigator was at a crucial point in flying the range. To overcome this difficulty, another transmitter and antenna are used. The range A and N transmitter is tuned 1020 cycles above the normal carrier frequency, and a second transmitter connected to a single vertical antenna is tuned to the assigned frequency of the station. This second transmitter is voice-modulated with weather and flight information, but all frequencies near 1020 cycles are filtered out. Thus, the A and N signals are received as a 1020-cycle tone, while the voice modulation covers the audio range except those frequencies near 1020 cycles. This combination is resolved by the receiver so that the navigator can listen to either the radio-range signals or the voice signals, as required. In the United States, the two transmitters are normally one dual transmitter. Both use the same chassis, and the two oscillator crystals use the same crystal oven, so that any drift in frequency due to temperature change will not affect the difference frequency between the two crystals. To ensure continuous operation, a standby dual transmitter is kept available in case the operating transmitter fails.

11-976. VISUAL RADIO RANGE. A visual radio range was developed by the Bureau of Standards, but after field test it was discarded because of mechanical deficiencies. In this method a common oscillator drove two amplifiers—one modulated at 65 cps and the other at 86.7 cps. The sideband frequencies were transmitted from an antenna equipment of the low-frequency radio-range type, and the relative strength of the sidebands received at the aircraft was dependent on the direction of the transmitting station from the aircraft. After amplification and detection, the 65-cps and 86.7-cps signals were used to operate a bearing indicator which contained an electromagnetic vibrator. There were two vibrator elements (reeds) on this instrument, each of which was mechanically resonant with one of the signals. The amplitude of vibration of each reed was proportional to the amplitude of the signal with which it was resonant. When the two reeds vibrated equally, the aircraft was on-course; when one reed vibrated more than the other, the aircraft was off-course and had to be turned toward the weaker signal. In a modification to this arrangement, a third amplifier modulated at 108.3-cps was added to the transmitter. The three amplifiers fed three pairs of vertical antennas through a goniometer having six secondaries, with coils connected in such a manner that three figure-eight field patterns 60 degrees apart were formed. These patterns are shown in figure 11-126AV. This arrangement gave 12 possible courses rather than the four obtainable with only two amplifiers, since any two reeds could be compared. Three of the possible 12 combinations are indicated in figure 11-126AV. The modulation frequencies used for the arrangement previously described were

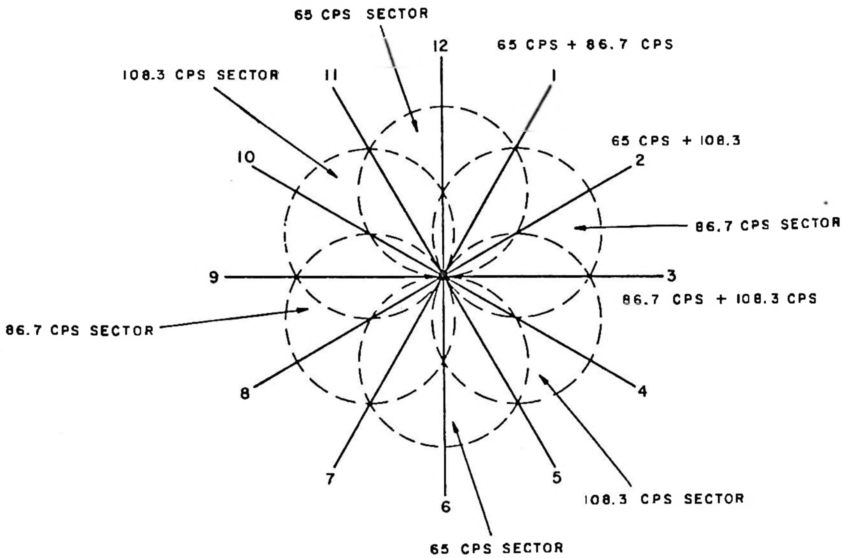


Figure 11-126AV. Visual Radio Range, with 12 Possible Courses

chosen so as to prevent any possible harmonics of one frequency from affecting another reed. The 12-course range was never put to practical use because, among other things, if the relative gain or one of the amplifiers changed, a change in course would be indicated.

11-977. RADIO-RANGE ANTENNAS. To obtain the field patterns required for radio-range work, special antenna equipments are necessary. Loop or Adcock arrays, as previously explained in connection with radio direction finders, provide the desired range patterns, but are not normally used because of the limited radiated power obtainable from them. Instead, simple vertical antennas in arrays and fed in pairs, as previously described, are used. These antennas must be tuned to the operating frequency of the transmitter. In the United States radio-range facilities operate in the 200- to 400-kilocycle band; therefore, antenna towers from 100 to 300 feet are used. In spite of their size, these antennas require loading in order to bring them into resonance; therefore, they do not operate a maximum radiation efficiency. The transmitter is rapidly switched from one antenna pair to the other by means of a relay. Special filters and tuning are used to prevent arcing at the relay contacts, and thus to keep spurious sidebands at a minimum. The beam patterns of radio ranges do not usually have a right-angle (90-degree) relationship as shown in figure 11-126AU. Normally, each beam is directed toward an important landmark, such as an airport, a city, etc, and beams differing in azimuth by angles other than 90 degrees often result. Nearly any angle can be obtained by varying the size of the individual antennas or the energy fed to them, so that lopsided figure-eight field patterns result. An example of this is shown in figure 11-126AW. Another method of obtaining the same objective is to add a vertical antenna phased in such a manner that its radiation adds to two adjacent field

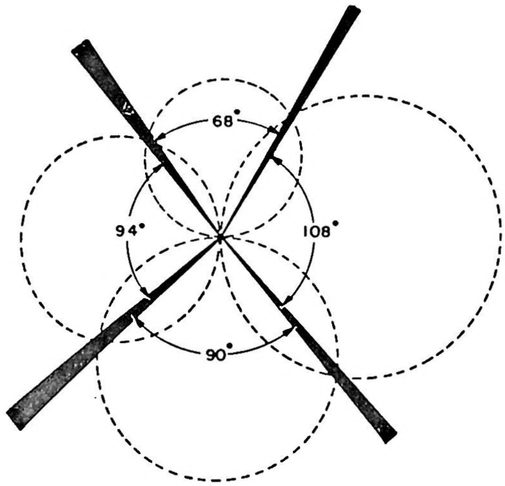


Figure 11-126AW. Radio-Range Beam Pattern with Unequal Energy Fed to Each Antenna

patterns and subtracts from the other two. Since receiving antennas as well as transmitting antennas have a definite field pattern, improper design of the receiving antenna can cause errors in the course. A simple vertical whip antenna introduces the least amount of error, but it must be quite long to ensure ample signal strength. A long antenna offers wind drag to the aircraft, and is subject to icing and bending during flight; it also introduces errors if the plane is not level, as during a turn or when gaining or losing altitude. Many different types of antennas have been used in an effort to obtain a completely directionless pattern, but none of them have been completely satisfactory.

11-978. The low-frequency radio-range receiver is very simple; this is the principal reason for the extension use of radio range. The receiver is essentially an ordinary communications receiver which is capable of being tuned through the radio-range band. The indicator is a speaker or headset connected to the output of the receiver. As mentioned previously, with simultaneous voice and radio-range signals available, both the voice and radio-range signals are present and must be separated by the receiver. Two filters are incorporated in the receiver for this purpose. One filter passes frequencies near 1020 cps, and makes them available to the pilot when the switch is in the radio-range position. When the switch is in the voice position, the other filter passes all audio frequencies except those near 1020 cps, so that voice modulation can be heard. Some of the problems associated with low-frequency radio-range operation are eliminated, or at least reduced, when uhf is employed. The greatest advantage of the uhf radio range is that a large number of frequencies are available, so that interference between radio-range stations is negligible. Furthermore, the radio waves at these frequencies are not reflected or refracted by the ionosphere; hence, several stations can operate at the same frequency if they are separated by a few hundred miles, and bent courses and multiple courses are almost completely eliminated. Another important advantage of uhf is that the shorter wavelengths allow the use of relatively small antennas, and the air hazard presented by the large towers necessary with lf radio-range equipment is eliminated. Uhf radio-range equipment has some disadvantages, however, the most important of which is the limited range obtainable at these frequencies. This is not a disadvantage in heavily populated areas where airfields are close enough together to allow overlap between radio-range beams, but in sparsely settled sections extra stations may be required to cover areas where there are no airports. Another disadvantage of the uhf radio range is that the signals produced can be almost completely blocked by a building, mountain, or other obstruction along the way. For example, if the antennas are placed too near an airport hangar, the hangar may completely block the signal in one direction or cause false indications. A mountain, which is higher, can do the same thing at a much greater distance, thus impairing the usefulness of radio range. However, proper location of the radio-range antenna array can usually eliminate this problem.

11-979. MARKER BEACONS. As a pilot flies directly over a radio-range station, the signal disappears because the antenna arrays used for radio range are vertically polarized and therefore radiate no energy directly upward. This cone of silence can be used as a means of fixing the location of the aircraft directly over the radio-range station, and in bad weather to indicate to the pilot when to let down to the airport below. However, because it is sometimes difficult to locate the cone of silence, and also because false cones of silence are often present behind mountains, more positive identification of the cone is required. For this purpose a beacon (Z marker) operating at 75 megacycles is used. The beacon transmitter has an antenna array which produces a cone of signal in the cone of silence of a radio range. The cone of signal is modulated by a 3000-cycle signal which is coded to provide identification of the station and the particular marker. Another type of beacon, known as a fan marker, operating at the same frequency and also modulated by a 3000-cycle signal, serves to inform the navigator when he has reached the point where the beams from two radio ranges cross. To enable the navigator to distinguish between the fan-marker and Z-marker beams, the former is keyed so as to identify the particular legs that meet at the marker location. The field pattern of the fan marker is elliptical, so that even when the aircraft is off the course, the pattern is wide enough to provide an indication. Other fan markers, used for identification of vital points along the way, operate at the same frequency; therefore, the same receiver can be used for all beacons. The other markers are modulated at different frequencies, to provide identification of particular points. An inner marker, which is modulated at 1300 cycles, identifies the edge of the airport, and an outer marker, modulated at 400 cycles, is usually located about 5 miles from the airport. These marker beacons are used to aid aircraft in landing. Until a few years ago, radio range was the most widely used navigation aid. It has since been replaced at many airports in the United States with the omnidirectional radio-range method, but four-course radio range is still used at most airports as both a backup navigation method, and also for those aircraft that do not carry omnirange equipment.

11-980. During the 1920's the British developed a rotating beacon navigation arrangement known as Orfordness. This navigational method was designed specifically for use by merchant and naval vessels to enable them to obtain reasonably accurate bearings over comparatively long ranges. Low frequencies are used for this purpose. The transmitter output is connected to a rotating loop antenna, which is used to produce a rotating figure-eight field pattern. Each time one of the nulls of the figure-eight pattern is directed toward north, the transmitted signal is interrupted. The loop antenna is rotated at the rate of 1 revolution per minute, or 6 degrees per second. The navigator records the time lapse between perception of the interruption and perception of the signal null. This time (in seconds) multiplied by 6 equals the number of degrees between north and the direction of the craft from or to the beacon. To simplify measurement, a stop watch calibrated in degrees is used and azimuths are read directly. Since the figure-eight field pattern of the transmitting antenna (rather than the pattern of the receiving antenna) is used, the Orfordness beacon navigation method may be considered as the opposite of the radio direction finder. The errors encountered are very similar for both methods; however, since distinction between the true and the 180-degree ambiguous null is not possible with the Orfordness beacon, its effectiveness is reduced. This lack of effectiveness is somewhat overcome, however, by the ease of taking bearings and the small amount of equipment required in the craft. Because more elaborate shielding and balancing equipment can be used at a fixed ground installation, polarization error is also considerably less than with mobile direction-finder equipments. The Orfordness beacon was used quite extensively by the British until more accurate navigation methods were devised.

11-981. During 1940 and 1941, the Germans used a navigation arrangement known as Electra (the forerunner of Sonne), which operated very much like the previously discussed radio range. This method, like radio range, was used mainly for the homing of aircraft, but could be used to obtain a position fix at points where the beams of two equipments crossed. The field pattern illustrated in figure 11-126AX (the one generally used by the Germans in their Electra arrangement) is produced by three antennas lying along a straight line (however, other configurations are possible). The center antenna is spaced three wavelengths from either of the end antennas. The two end antennas are fed 180 degrees out of phase with each other, and the center antenna is fed 90 degrees out of phase with each of the end antennas. Equal-strength signals are applied to the end antennas, while the signal applied to the center antenna is four times as great. If the phase of the signals applied to the end antennas is suddenly reversed, the field pattern shown in figure 11-126AX will reverse, placing the double-ended lobe at the left. In Electra operation, this reversal process occurs twice each second. The pattern shown in the figure is held for 5/6 second, at which time the pattern is reversed and held for 1/6 second, after which the cycle is repeated. The result of the reversal of the field pattern is illustrated in figure 11-126AY, in which the 5/6-second pattern is shown by the solid line. Notice that the side lobes of the two patterns overlap to form beams as in the radio-range method of operation. While in the area of a solid lobe, the navigator will receive a series of dashes (5/6-second duration); while in the area of the dotted lobe, he will

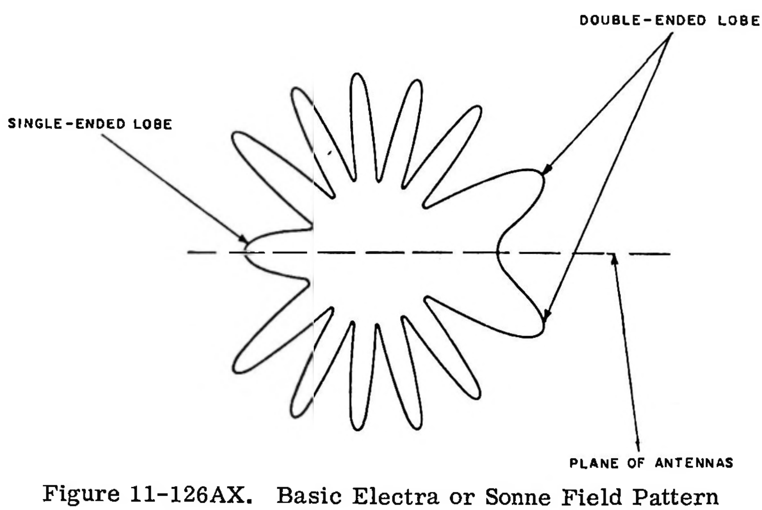


Figure 11-126AX. Basic Electra or Sonne Field Pattern

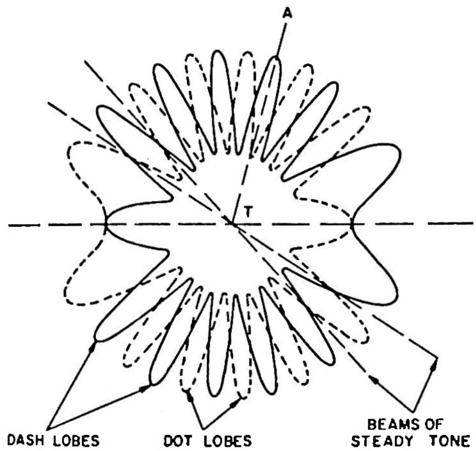


Figure 11-126AY. Electra or Sonne Dot-Dash Field Pattern

receive a series of dots (1/6-second duration); and while on the beam, he will receive a continuous tone. As can be seen from figure 11-126AX, the Electra method used a large number of possible legs (beams). The Electra method was improved by the Germans during World War II until an omnidirectional arrangement, which operated in a manner somewhat like that of the Orfordness beacon, was developed. This method was named Sonne by the Germans, and was referred to as Consol by the British. Instead of using a watch for timing, as required by the Orfordness beacon, this method used rotating field patterns from which the navigator could obtain timing data. All of the information provided thus far for the Electra method also applied to the Sonne method; and, in addition to the reversal of the field pattern to obtain the dot-and-dash sequence, the Sonne arrangement employed slow phase shifting of the two patterns until the dash pattern in effect replaces the dot pattern, and vice versa. The entire phasing process required 1 minute. Consider the effect of the phasing action upon the solid field pattern shown in figure 11-126AY. With the exception of the two ends of the pattern which are in line with the plane of the antennas, each signal minimum begins to expand and each signal maximum begins to contract. This expansion and contraction continues until at the end of 1 minute the former minimum points become the maximum points, and vice versa. In addition, the solid pattern formation changes place with the dotted pattern formation. The result of this causes the beams of steady tone to shift (those in the top half of the pattern shown in figure 11-126AY move counterclockwise, while those in the lower half move clockwise). The degree of shift places each new beam in the position previously occupied by the beam to the left. This phase-shifting sequence is used by the navigator to determine his exact bearing to the transmitter with respect to the beams of steady tone, which are present before the phasing is accomplished. The 1-minute phase-shifting period is followed by a 1-minute period of time used for identification, after which time the entire cycle begins again. The identification period consists of 1 second of no transmission, a 56-second transmission of identification signals and tone from the center antenna only, followed by a 3-second period of no transmission, which informs the navigator that the next cycle is about to start. Assume, for explanation purposes, that a navigator is located on line TA (shown in figure 11-126AY) at the start of the phase-shifting period. After the 3-second silence period, the navigator being located in a dash area will receive a series of dashes, then the steady tone, and finally a series of dots before the identification period begins. By counting the number of dashes (or dots if orginally in a dot area) preceding the steady tone, the navigator can determine his exact position, relative to the original position of the adjacent beams. The entire sequence of operations involved in navigation using the Sonne method is as follows: First the station bearing is determined to within approximately 20 degrees by using a radio direction finder, celestial navigation, or dead reckoning, to establish the location to a sector (within adjacent beams). Next the navigator counts the number of dashes or dots which occur between the end of the 3-second silence period and the time when a steady tone is heard. Then the exact bearing is determined by using a chart which compares the bearing to the station with the number of dashes or dots heard while in the particular sector. To use the Sonne method, the navigator requires only a standard type of receiver and a set of charts. The method has the advantage of providing accurate bearing information throughout 360 degrees. Furthermore, since the operating frequency of Sonne was in the vicinity of 200 kc, it provided long-range navigation (approximately 2000 miles over water and 1000 miles over land). The accuracy of azimuths is claimed to be about 1.7 degrees during the day and 2.3 degrees at night. The Sonne arrangement is subject to inaccuracies in azimuth due to weather and terrain conditions, a failing which is prevalent in all equivalent-type navigation methods. Its major disadvantage, however, is the relatively long time required to obtain a bearing (up to 2 minutes, or one complete cycle of operation).

11-982. The radio-range equipment previously discussed was an up-to-date direction-finding method when it was introduced in the 1930's. Since that time, however, a good deal has happened to air travel. There are now many times the number of aircraft, and they move at many times the speed of former aircraft. These two facts alone are enough to antiquate the direction-finding methods using low-frequency and the four-course range. The modern counterpart is a range with no limit on the number of courses that an aircraft may take in approaching or leaving the ground station. It is also vhf equipment, and therefore eliminates some of the course errors (such as false beams) so prevalent in the old methods.

11-983. <u>OmniDirectional Range</u> (ODR), or simply <u>omnirange</u>, is a radio range which provides the navigator with a visual track in any direction leading toward or away from the ground station. An infinite number of courses are therefore available to aircraft equipped for this method of navigation. The courses are called <u>radials</u>, since they radiate from the station as spokes radiate from the hub of a wheel. The ground station effectively transmits a different signal over each of these radials. The azimuth information is generated at the station and is transmitted to the aircraft, where it is then translated by the airborne equipment into a indication which is usable by the navigator. The ground station is basically composed of a transmitter and a five-element antenna equipment. The operating frequency is within the band of 112 to 118 mc, a frequency range which provides essentially line-of-sight propagation. The effective range of the equipment with an aircraft flying at an altitude of 500 feet is approximately 30 miles, and at 20,000 feet, approximately 200 miles. The practical range of the equipment, however, extends beyond this for greater altitudes. To enable low-flying aircraft to successfully use the range, stations are generally located within 50 miles or less of each other. Because of this relatively close spacing, many stations can be received, even at medium altitudes; therefore, many different frequencies must be used to avoid interference. Since adjacent operating frequencies are separated by 100 kc, 59 channels (total) within the 112 to 118-mc band are available in a specific area. The power output of a ground omnirange station, when using the AN/FRN-12 transmitter equipment, is 200 watts.

11-984. The method by which the omnirange equipment effectively provides an unlimited number of courses is very similar to that of the Oxfordness beacon arrangement previously discussed. The major difference is that in the omnirange method the determination of a position relative to the ground station does not require the use of a stop watch, but rather is accomplished automatically, the bearing being indicated directly in degrees on a meter. This method operates on the principle of phase comparison between two audio signals. The difference in phase is a function of the directional location of the receiving equipment with respect to the transmitter; theoretically, therefore, a different course exists for each azimuthal position around the station. The five-element antenna arrangement of the ground station radiates two different signals: one signal from one element of the antenna, and the other signal from the remaining four elements. The signal-element antenna produces an omnidirectional field pattern, and the resulting signal is referred to as the reference-phase signal. The other four elements of the antenna produce a figure-eight field pattern, and this signal is referred to as the variable-phase signal. The reference-phase signal consists of a 30-cycle signal which frequency-modulates a 10-kc subcarrier (9960-cycle subcarrier in some equipments), which, in turn, amplitude-modulates the rf carrier. The variable-phase signal is simply the unmodulated rf carrier. However, the variable-phase figure-eight field pattern is rotated at a rate of 30 times a second and effectively produces, at any fixed point in space, a 30-cycle signal which varies in amplitude. The 10-kc subcarrier frequency is incorporated to enable the receiving equipment to distinguish between the two signals, which are received simultaneously. The rotating figure-eight pattern is obtained by using a capacitive goniometer to feed the signal to the four elements of the antenna. Figure 11-126AZ illustrated how the field patterns from the four antenna elements combine to form a figure-eight pattern which is of constant field strength but which rotates as the goniometer is rotated. At 0 degrees one pair of elements is fed all the signal, producing a figure-eight pattern; at 45 degrees both pairs are fed equal amounts of signal that are 90 degrees out of phase, producing a figure-eight pattern 45 degrees removed from the 0 degree pattern; and at 90 degrees the other pair is fed all the signal. Since the goniometer turns smoothly, the figure-eight pattern does likewise. The goniometer is rotated by an 1800-rpm (30-rps) synchronous motor. Thus, a rotating pattern is produced, and when the carrier is demodulated at a receiver, a 30-cps component will be present. Still with reference to the demodulated received signal, when the receiver is located due north of a ground station, the reference-phase and variable-phase signals will be in phase. This condition is obtained by producing the frequency-modulated subcarrier mechanically by means of the same shaft that turns the goniometer rotor. A simplified block diagram of the transmitting equipment is shown in figure 11-126BA. The synchronous motor rotates both the goniometer and the ac generator,

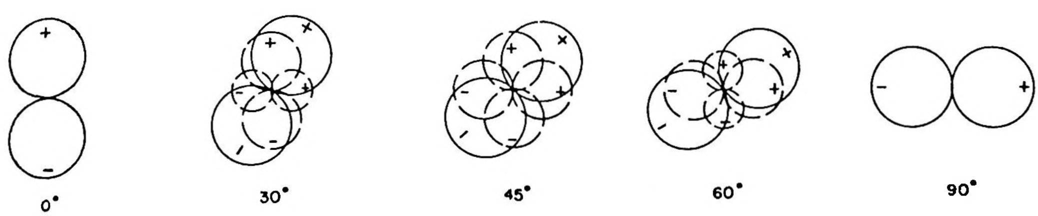


Figure 11-126AZ. Rotating Figure-Eight Field Pattern Shown During 90 Degrees of Rotation

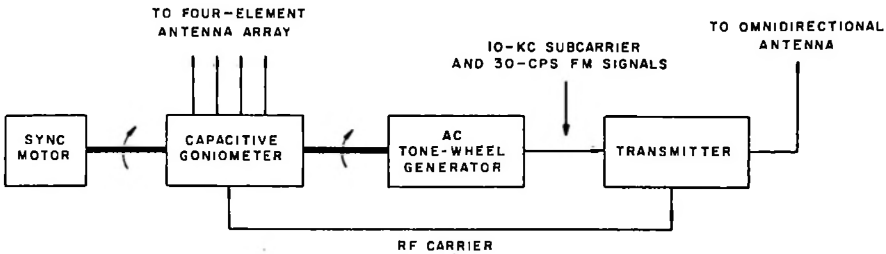


Figure 11-126BA. Simplified Block Diagram of Omnirange Transmitting Equipment

to ensure that the radiated signals will be synchronized. The ac generator is a tone-wheel type of pickup which has nonuniform spacing of teeth on the wheel to provide the 10-kc subcarrier containing the 30-cycle frequency modulation. This signal modulates the rf carrier in the transmitter and is then applied to the omnidirectional antenna. The rf carrier is also fed to the goniometer section. At the aircraft, the function of the receiving equipment is to determine the phase difference between the two 30-cps modulation components. This phase angle, in terms of degrees, is the bearing of the aircraft relative to the station. Figure 11-126BB shows a comparison of the phase of the reference and variable signals as they would be observed in the audio sections of receivers located at various azimuth positions relative to the ground station. The t_0 reference in the figure indicates that all waveforms shown are occurring simultaneously. The receiving equipment is composed basically of a conventional superheterodyne receiver, followed by a filter, a detector, and phase-comparer circuits, as shown in figure 11-126BC. Upon leaving the receiving section, the reference-phase and variable-phase signals are separated by filters. The reference-phase signal is passed through the 10-kc filter circuit to the fm detector circuit. Here the 30-cycle audio signal is restored. It is then passed on to a phasing network and afterward to a phase-comparer

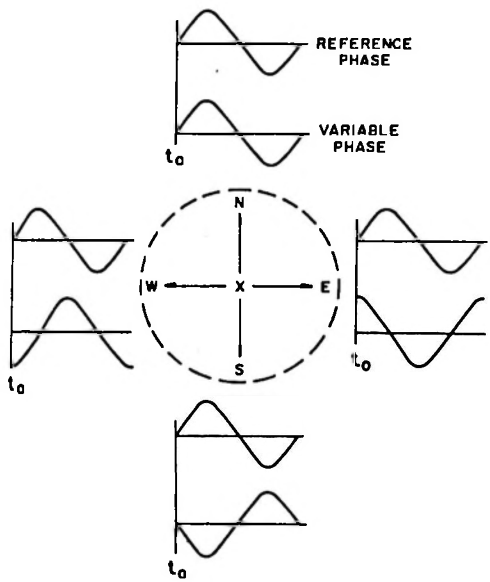


Figure 11-126BB. Relationship of Reference and Variable-Phase Signals at Various Azimith Positions About the Transmitter

circuit. The variable-phase signal is blocked by the 10-kc filter and accepted by the 30-cycle filter. It is then also applied to the phase-comparer circuit.

11-985. The intelligence obtained from the signals by virtue of the operation of the phasing network and the phase-comparer

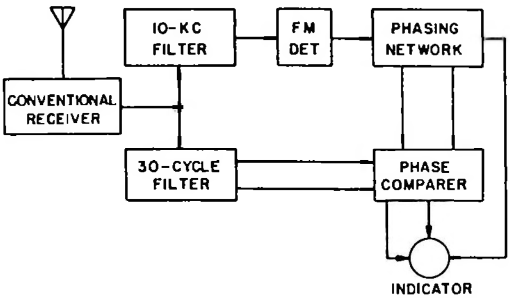


Figure 11-126BC. Simplified Diagram of Omnirange Receiving Equipment

circuit falls into three distinct categories: the radio bearing or course information, the "to" or "from" station information, and the heading indication. Several types of meters are used to display this information; the most generally used type is shown in figure 11-126BD. This meter is also used for Instrument Landing System (ILS) approach, for which the horizontal pointer (as well as the vertical pointer) is used. The vertical pointer is also used for omnirange, and indicated by position (left or right) on which

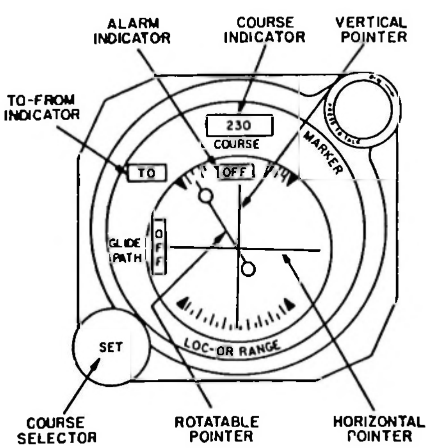


Figure 11-126BD. Typical Omni-Range Receiving Indicator

side of the selected course the aircraft is located. When in the center as shown, it indicates an on-course heading. The number at the top of the meter is the course heading to or from the station in degrees from north. Any course can be selected by the control at the lower-left. The rotatable pointer indicates the angular difference between the selected course and the magnetic heading of the aircraft. A difference in these readings indicates that a cross wind is causing the aircraft to "crab" in order to stay on course. The upper-left "to" indication will change to "from" when the station has been passed and the signal is being received from the reverse direction. The "off" indication, shown over the vertical pointer, is an alarm. The appearance of this indication, which is normally not visible, indicates that some part of the arrangement is malfunctioning and that the meter indications are unreliable.

11-986. Another type of meter is known as the <u>Radio Magnetic Indicator</u> (RMI). This indicator has a rotatable dial face that indicates the direction of flight relative to magnetic north, and a pointer that indicates the azimuth of the station being received. The pointer indicates the azimuth on the dial face so that it can be compared, at a glance, with the aircraft heading. The information presented is not as complete as that presented by the previously described meter; for this reason this meter, when included, is used as an auxiliary unit. The basic information given by omnirange can be presented with either the aircraft or the ground station as the point of reference. In figure 11-126BE the aircraft is the point of reference, and the direction to the ground station is 80 degrees. If the ground station is used as the point of reference, the direction of the aircraft from the ground station is 260 degrees (180 degrees plus the original 80 degrees). The "to" or "from" sense is selected by the equipment operator in the

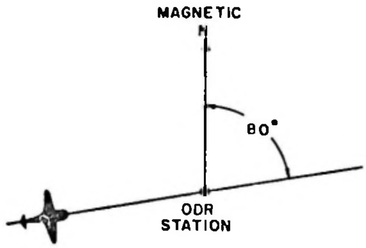


Figure 11-126BE. Direction of Aircraft to an Omnirange Station

aircraft, and serves to establish a reference point for the course line. For example, if the "to" reading is 120 degrees, the "from" reading is 300 degrees. Since the established D F practice uses the aircraft as a reference, it is common to express omnirange bearings in the "to" position. To obtain a fix for position determination of the aircraft, triangulation from two or more ground stations, as shown in figure 11-126BF, is necessary. This method is similar to that used in locating a transmitter by means of radio direction finders.

11-987. In addition to the 30-cps component, the reference carrier is also modulated by a 1020-cps code signal, which identifies the station. If desired, the same carrier can also be voice-modulated for communications purposes.

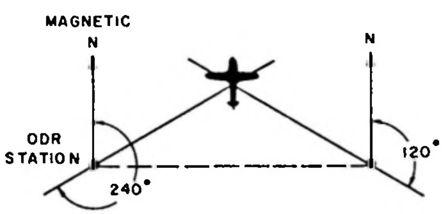


Figure 11-126BF. Use of Two Omnirange Stations to Obtain a Fix

11-988. GONIOMETER. The term goniometer, in direction finding, is applied to a device used to couple two or more input circuits, usually connected to antennas, to an output circuit such as a radio receiver. This is done in such a manner that the degree of coupling varies with the rotation of a shaft. The coupling between one input circuit and the output circuit increases, while the coupling between the other input circuit and the output circuit decreases. When properly connected, a well constructed goniometer provides an output, at each position of its shaft, identical to that which would be produced by a single figure-eight-pattern antenna oriented to the corresponding position. Thus, the goniometer provides an equivalent for rotation of the antenna and makes it possible to use large fixed-antenna arrangements which would, in themselves, be too cumbersome for an operator to rotate.

11-989. INDUCTIVE GONIOMETER. The most common form of goniometer is the inductive goniometer, which usually consists of two fixed windings, arranged at right angles to each other, and enclosing a third winding which is rotatable by means of a shaft; see figure 11-126BG. When the two fixed windings are connected to two identical antennas having figure-eight patterns and arranged at right angles, the magnetic field

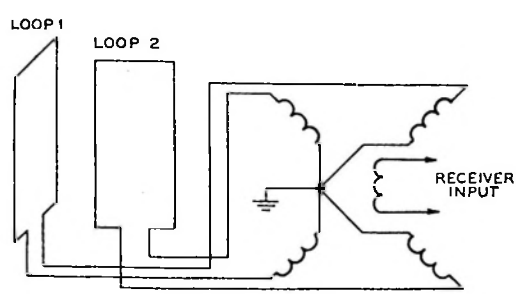


Figure 11-126BG. Basic Goniometer Circuit within the goniometer will have a direction, with respect to the fixed windings, corresponding to the direction of arrival of the signal with respect to the fixed antennas. As the internal winding (search coil) of the goniometer is rotated, its output will vary from maximum to minimum twice per revolution, exactly as would the output of one of the antennas if it were rotatable. The positions of minimum output, or nulls, are used to determine the bearing exactly the same way as if one rotatable antenna were used. The goniometer may be rotated by hand, thus providing for manual null-seeking, or it may be continuously rotated by a motor drive and employed with an automatic visual bearing indicator.

11-990. CAPACITIVE GONIOMETER. Another form of goniometer is the capacitive goniometer, which consists of two fixed sets of capacitor plates enclosing a rotatable set of plates. Operation of this type is similar to the operation of the inductive goniometer except that an electric field rather than a magnetic field is established within the goniometer. In practice, the capacitive goniometer is normally used at frequencies below 100 megacycles, since it is difficult to construct accurate and efficient inductive goniometers for these frequencies.

11-991. REQUIREMENTS. In order to minimize errors, it is necessary to construct goniometers with extreme precision. The basic requirements for accuracy in a goniometer coupling arrangement are as follows:

a. The fixed elements must be electrically identical.

b. There must be complete absence of coupling between the fixed elements.

c. Accurate positioning of the fixed elements at the same angle as the antenna is necessary.

d. Coupling between the rotating element and the fixed elements must vary with the shaft rotation angle in accordance with the same law as the variation of antenna response with the azimuth angle.

11-992. AZIMUTH INDICATORS. The azimuth indicator may be employed independently or in conjunction with a radar set. When operated independently of a radar set, one operator is stationed at the vhf section, and another at the hf section. The operator at the vhf section maintains communication contact with the aircraft, determining which frequency the direction finder is to operate, and reads the azimuth indications. When the azimuth indicator is used in conjunction with a radar set, bearing indications are read on the miniature azimuth indicator by the operator stationed at the vhf section. This operator relays the azimuth information orally to the operator at the radar plan position indicator. The following procedure is employed for both uses of the azimuth indicator:

a. The radar operator informs the operator at the vhf section which DF channel to select. The channel should be identified by the letter "A", "B", etc, for the frequencies of the vhf section, and by the number "1", "2", etc, for the frequencies of the hf section.

b. The direction-finder operator at the vhf section places the REMOTE-VHF-HF selector switch in position for operating at the vhf or hf section, according to the desired frequency band.

c. If the frequency is in the vhf band, the operator at the vhf section selects the proper channel by turning the LOCAL CHANNEL SELECTOR. If the frequency is in the hf band, the operator at the vhf section informs the operator at the hf section to switch to the proper channel.

11-993. The azimuth indicator unit transforms the sine-wave output frequency of the antenna equipment into a properly timed pulse which triggers the azimuth indicator dial, thus indicating the bearing of an aircraft.

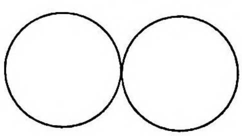
11-994. DF OPERATIONAL TEST. The two main checks to be made on a direction-finder unit provide information which permits you to determine whether transmitted signals at each channel frequency are being received by the direction-finder equipment, and the accuracy of the azimuth indications. The results of these checks depend upon whether the equipment is operated from the vhf-hf or radar station. When operated with a radar set, it is necessary to synchronize the vhf, hf, and radar scope indications. Take a suitable signal source (you can use either the fixed transmitters employed for normal hf and vhf communications or the target transmitters available on the direction-finder unit) and determine the exact azimuth of the transmitting antenna with relation to the receiving antenna array. Also determine the azimuth of the target transmitters. The azimuth of the fixed transmitters can be established by using the direction-finder equipment, once it has been aligned.

11-995. VHF CHECK. Transmit a signal on the frequency of the channel under test. The signal source should be at least 100 feet from the vhf antenna, and the exact azimuth in relation to the vhf antenna must be known. The indicator dial should point to the correct azimuth, and the audio signal of the test signal source, plus the 30-cycle component imposed by the vhf goniometer, should be heard in the speaker or headset. This test should be made on all channels, tuning the frequency of the test signal source to the frequency of the channel under test.

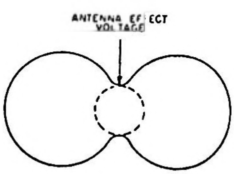
11-996. HF CHECK. Transmit a signal on the frequency of the channel under test. The signal source should be at least 500 feet from the central antenna of the hf antenna array, and the exact bearing in relation to the hf antenna array should be known. The indicator dial should point to the correct bearing, and the audio signal of the test signal source, plus the 30-cycle component imposed by the hf goniometer, should be heard in the speaker or headset. Repeat this check for all channels, each time tuning the test signal source frequency to the frequency of the channel under test.

11-997. BALANCE TEST.

11-998. SYMMETRY. One basic assumption in calculating the directivity pattern of an ideal Adcock antenna is that the two spaced elements, and their associated circuits up to the junction point, are electrically identical. This means that the same output voltage will be produced by a signal of given strength arriving at either antenna element, and that the phase relation between the output voltage and the voltage induced by the incident radio wave will be the same in either case. For a crossed-Adcock or a crossed-loop system, using a mechanical or electronic goniometer to simulate rotation of the antenna equipment, it is further assumed that the two crossed members are electrically identical and differ exactly 90 degrees in their orientation. In direction finders using aural-null or visual-null indication, sharpness of the null in the directivity pattern is important for the sake of bearing readability. Ideally, there should be no output when the direction-finder antenna has its null aligned with the direction of signal arrival, but perfection is never quite attained. If the gain of the direction-finder receiver is made high enough, there will be some output even at the null, which thus becomes a minimum rather than zero. For weak signals, this minimum output is chiefly noise. For strong signals the minimum is established chiefly by antenna effect. In either case, the directivity pattern of a loop or Adcock antenna is changed from the ideal figure-eight pattern with sharp nulls, as shown in part A of figure 11-126BH, to one with rounded nulls, as shown in part B. As far as the nulls are concerned, similar results can be produced by any means which adds another voltage 90 degrees out of phase with the ideal DF antenna voltage and which has a maximum where the ideal voltage has a null. Against a background of silence, a normal human ear can detect sounds which deliver less than a billionth of a microwatt to the eardrum, but it can scarcely detect a change less than 2 decibels; under ordinary circumstances, the minimum change detectable by the ear is 3 decibels. For this reason, and for convenience in mathematical analysis, the null width of a direction finder is defined as the angular interval between azimuths at which the receiver output rises 3 decibels above its minimum (null) value. For steady signals, the apparent azimuth can be determined within 10 or 20 percent of the null width, depending on the skill of the operator and the type of indication used.



A
POLAR DIAGRAM
LOOP ANTENNA WITH
NO ANTENNA EFFECT



B
POLAR DIAGRAM
LOOP ANTENNA
WITH ANTENNA EFFECT

Figure 11-126BH. Polar Diagram of Loop Antenna With and Without Antenna Effect

11-999. NULL RATIO. Measurement of null width is sometimes difficult, particularly when the null width is 1 degree or less. In such cases the null sharpness may be specified by means of the null ratio F_n, which is the ratio of output voltages at the azimuths of maximum and minimum response. Null width can be calculated from null ratio, and vice versa, if the shape of the ideal directivity pattern is known. For example, if the ideal pattern is a figure eight, as in a simple loop or Adcock antenna, the effective height varies with azimuth according to the following relation:

$$h = h_m \sin \phi \qquad (1)$$

where h_m is the maximum effective height. With antenna effect adding a constant voltage in quadrature, corresponding to an effective height h_n, the resultant effective height becomes:

$$h = \sqrt{(h_m \sin \phi)^2 + h_n^2} \qquad (2)$$

At the null, $\phi = 0$ and $h = h_n$. At an azimuth half the null width away, $\phi = \pm \phi_n$, where (by definition) the antenna response is 3 decibels greater, $h = h_n\sqrt{2}$; hence,

$$h_n = h_m \sin \phi_n \qquad (3)$$

At the maximum, $h = \sqrt{h_m^2 + h_n^2}$. Accordingly, the null ratio is:

$$F_n = \sqrt{1 + (h_m/h_n)^2} = \sqrt{1 + 1/(\sin \phi_n)^2} \qquad (4)$$

For reasonably sharp nulls, this equation may be written as an approximation as:

$$F_n = h_m/h_n = 1/\phi_n \text{ radians} = 57.3^\circ/\phi_n^\circ \qquad (5)$$

Thus, a 1 degree null width ($\phi_n = 0.5^\circ$) requires a null ratio of 115. A null ratio of 2 will make the null width 70 degrees (ϕ_n = 35.23 degrees). Equation (4) or (5) may be used for a coaxial spaced-loop antenna receiving vertically polarized signals, if ϕ_n is replaced by $2\phi_n$, to allow for the fact that the nulls are twice as sharp (half as wide) in the four-lobed pattern as in a figure-eight pattern.

11-1000. BEARING ACCURACY AND SENSE CHECK.

11-1001. VHF BEARING ACCURACY CHECK. At least once a week, or whenever the operator suspects that bearing indications are incorrect, the following check should be made:

a. Place the vhf section in operation.

b. Place the vhf antenna or test oscillator at least 100 feet north of the vhf section.

c. Turn on the oscillator.

d. Note the bearing indication. Set the indicator scale so that the dial points to 0 degrees.

e. Setting up the test oscillator to transmit from the east, south, and west, should produce bearing readings of 90 degrees, 180 degrees, and 270 degrees, respectively.

11-1002. HF BEARING ACCURACY CHECK. The accuracy of the hf bearing indications is checked in the following manner:

a. Place the hf section in operation.

b. Place a transmitter at least 500 feet north of the central antenna.

c. Turn on the transmitter.

d. Note the bearing indication, and set the indicator scale so that the dial points to 0 degrees.

e. Set up the transmitter to transmit at spacings of 45 degrees about the antenna array (that is, at 45 degrees, 90 degrees, etc). The bearing indications should be accurate within ±5 degrees.

11-1003. RECIPROCAL BEARING ERROR AND SENSE CHECK.

11-1004. DESCRIPTION. The directivity pattern of a loop or Adcock antenna may be changed from a symmetrical figure eight to a lopsided one, or even a heart-shaped curve, by combining voltage from an auxiliary antenna with the output of the Adcock antenna. Usually this combination is intentional, for the purpose of determining the sense of propagation of a radio wave along its line of direction, but sometimes a small amount of sense voltage may be introduced accidentally when it is not wanted. The two nulls are displaced through equal and opposite angles. The average of the apparent signal azimuths determined from the two nulls (directly from the azimuth when the direct null is toward the signal source, and by addition or subtraction of 180 degrees from the azimuth read when the reciprocal null is toward the signal source) remains unchanged. It is still the true azimuth of the signal source (if the direction finder is otherwise accurate) in spite of the dissymmetry of the directivity pattern. Neither of the apparent azimuths is correct; each differs from their average by an angle which may be called the <u>reciprocal error</u>, and they differ from each other by twice this angle.

11-1005. MAGNITUDE. Consider an antenna with ideal figure-eight-pattern directivity, whose effective height varies with azimuth according to equation (1). The addition of a constant voltage in phase with that of the main antenna, corresponding to an effective height h_n, will make the resultant effective height as follows:

$$h = h_n + h_m \sin \phi \qquad (6)$$

Instead of having nulls at $\phi = 0$ degrees and 180 degrees, as it would without h_n, $h = 0$ at $\phi = -\phi_n$ and at $\phi = 180$ degrees $+ \phi_n$, where:

$$\sin \phi_n = h_n/h_m \qquad (7)$$

To a first approximation, valid for small reciprocal error (ϕ_n), this equation may be rewritten as:

$$\begin{aligned} \phi_n \text{ radians} &= \phi_n^{\circ}/57.3^{\circ} \\ &= h_n/h_m \end{aligned} \qquad (8)$$

11-1006. Note the similarity between equations (3) and (7), and between equations (5) and (8); in either case there is the same relation between ϕ_n and the ratio of effective heights, although the meaning of ϕ_n is different. The undesired pickup represented by h_n may produce either displacement, as in the previous case, of the nulls, or a combination of both, depending on the phase relation between desired and undesired components of antenna voltage. For this reason, null-clearing devices must be constructed with care, to insure that the voltages they introduce will be exactly in quadrature with the desired loop or Adcock voltage; otherwise, the process of balancing out null rounding might actually introduce reciprocal error.

11-1007. Precision in a direction finder is the quality which makes its indication sharply defined, so that you can determine the apparent azimuth of an incoming signal within very close limits. Accuracy is the quality which makes the observed azimuth very nearly the same as the true geographical azimuth of the signal source. Equipment operational precision can be had without accuracy, but the accuracy of a direction finder can never be better than its precision. For example, if its azimuth scale were displaced 10 degrees, an otherwise accurate direction finder so precise that apparent azimuths could be determined to the nearest 0.1 degree would have in every reading an error between 9.95 degrees and 10.05 degrees. Although two of the various factors which affect precision are beyond equipment control (strength and steadiness of the incoming signal), you can control the majority of factors. These factors include azimuth scale design and placement, indicator sensitivity, sharpness of the directivity pattern, and symmetry of the directivity pattern. Accuracy is affected by additional factors, including the spacing and symmetry of elements in fixed antenna equipments; these are directly controllable. Another factor is polarization of the incoming waves, but you can only partly control its influence. Still other factors are the DF site and the propagation conditions along the path of the incoming radio waves; over these factors the equipment designer has little control.

11-1008. AZIMUTH PATTERNS.

11-1009. APPEARANCE. Readability is a primary requisite in azimuth scales, or in any other scales. The scale must be reasonably large, clearly marked, and well lighted. If the interval between graduation is 1/8 inch or more, the operator can interpolate visually, reading halves, quarters, or perhaps even tenths of a division; however, for speed and minimum human error, it is preferable to provide enough graduations to make interpolation unnecessary. A psychological factor is the position of the azimuth scale with respect to the operating controls. This scale should be placed primarily for your convenience so that you do not shift attention from one item to another and thus waste time and effort.

11-1010. ECCENTRICITY. Mechanically, the most important characteristics of an azimuth scale are accurate and uniform spacing of graduations, and concentricity of scale and pointer axes. Suppose that the scale was originally adjusted to read correctly at all positions, and then shifted off-center sufficiently to cause an error of 1 degree in the azimuth reading. Thereafter, when the scale (or pointer) is turned in the normal manner, the azimuth indication will be in error by an amount varying cyclically from +1 degree to —1 degree and back again, the positive and negative maximums being 180 degrees apart, with points of zero error midway between. Since there is one maximum in each half of the azimuth range, this type of variation is known as semi-circular error. Any semicircular error introduced by the azimuth scale can be corrected by a slight mechanical adjustment, the accuracy of which can be tested with two straight edges. If each is aligned with a different pair of scale divisions 180 degrees apart, the two straight lines so defined should intersect on the axis of rotation or so close to it that the off-center distance is a negligible fraction of the distance between adjacent scale divisions.

11-1011. ELLIPTICITY. In the case of ring-shaped scales, ellipticity is introduced if the ring is squeezed or stretched in one direction more than another. In disk-shaped scales, such mechanical distortion is unlikely, but ellipticity of the graduation pattern may occur in the process of photoengraving if the scale is tilted while the graduations are being projected onto it. As a rule, the only remedy is to install a new azimuth scale, since stretching a distorted scale back into the proper (circular) shape requires a high degree of skill and delicate workmanship. If ellipticity is its only fault, an azimuth scale will have zero error for readings along both the major and minor axes of the ellipse, but maximum error midway between, where the graduations are displaced toward the major axis. Since the error reaches a positive or negative maximum (alternately) in each quadrant, this type of variation is known as quadrantal error.

11-1012. The directional characteristics of antennas are generally confined to planes; for example, the usual DF application is generally concerned with the horizontal plane, or the plane of azimuths. An antenna is ordinarily directional in other planes as well. Most radio waves have complete symmetry in but one plane, but this is not always true. Since the supposed direction of propagation is determined on the basis of the special dissymmetry of the amplitude and phase characteristics of the wave, the general problem of azimuthal direction finding is to make the wave dissymmetry in three-dimensional space appear as directional dissymmetry in one plane only. This requires that the azimuthal directivity pattern have the same shape for any angle of incidence. To obtain the directional characteristics of an antenna working in a particular electromagnetic field, use the appropriate equation describing the induction of voltage in the antenna, assume a fixed voltage condition in the antenna for a reference point, and determine the angle of azimuth, ϕ, which yields this condition. In most cases, the reference antenna voltage point used to determine an azimuth is the null condition, but sometimes other voltage conditions are used, such as the 0.707 maximum output. occasionally, the peak output voltage point is used.

11-1013. The antenna null voltage point is generally selected as an azimuth reference point because, in the vicinity of the null, the voltage is changing very rapidly for small changes of angle. A disadvantage of the null reference point method is that when "on azimuth", the voltage is small in the practical case (theoretically zero), and the signal may be difficult to hear. With the rapid advancement in instantaneous indicators, however, more adequate use is being made of the maximum available signal to help produce the azimuth patterns. However, even this problem is complicated by such considerations as bandwidth and circuit noise; thus, it cannot be predicted that working on the maximum side of the antenna will increase the sensitivity of the equipment to any great extent. For example, if an aural null direction finder is properly designed and operated, the gain of the receiver will be limited by the noise which can be tolerated by the ear when the antenna is in the null position. Under these conditions, with a high-Q antenna and input circuit, and negligible pickup by extraneous circuit elements outside the antenna equipment (both of which taken together yield a high maximum-to-null ratio), you only have to rotate the antenna a degree or two, one way or the other from the null, to produce a signal output in the headset that is 6 decibels or more above the noise level. As a consequence, it cannot be conclusively proved that working more and more toward the maximum side of the antenna will yield a greater DF sensitivity. Nevertheless, it is true that until external noise limits the operation of the set, a greater signal sensitivity can be obtained by working as near the antenna voltage maximum as possible. Signal sensitivity and DF sensitivity are different things. The former is concerned with obtaining the greatest over-all signal-to-noise ratio, while the latter is concerned with obtaining the greatest rate of change in signal-to-noise for a given amount of antenna rotation. The 0.707 maximum emf antenna reference point is sometimes picked when two crossed antennas (oriented at 90 degrees with respect to each other) are used, because such a voltage point is obtained when a signal is incident at 45 degrees of azimuth to the plane of each antenna. Under these circumstances, the induction in both of the antennas is the same, and azimuths are taken by rotating the antenna equipment and observing that the voltage increases in one while it decreases in the other. The maximum or near maximum voltage reference point is generally used with special antenna equipments which have either single or double lobular patterns.

11-1014. INSTRUMENT-LANDING EQUIPMENT.

11-1015. GENERAL.

11-1016. Since the movement of aircraft from place to place is accomplished above the earth's surface, complete navigational facilities for aircraft must include landing methods to ensure safe descent to the earth at the end of a flight. The simplest type of landing arrangement is that which permits the pilot to see the landing runway in sufficient time to maeuver the aircraft to a landing. Obviously, such a method is limited to visual flight conditions and would seriously limit the utility of aircraft if it were the only method available. To supplement or to replace the pilot's ability to see his way to a landing, a number of instrument landing equipments have been developed. This section will describe in detail the currently used operational landing equipments; references will be made to older or obsolescent equipment only insofar as they display principles of the current methods. The expression Instrument Landing System (ILS) is generally understood to mean an instrument-approach equipment. Although current equipments will permit completely blind landings and have been used for that purpose, the landing equipment is generally used by the pilot only to facilitate his approach and descent to a point where he can see the landing runway, from which point he completes a visual landing. ILS is a term used by the Civil Aeronautics Administration to denote a specific arrangement of electronic aids to an instrument approach used in the common method of air traffic control. ILAS is used by the military to denote the same kind of arrangement. In landing an aircraft under visual flight conditions, a pilot establishes a path through space by observing the distance to the point of touchdown, the altitude of his aircraft, and its lateral position with respect to the center line of the landing runway. From this information, he can adjust the attitude of his aircraft. An instrument landing equipment, therefore, should present the same information to the pilot. In other words, since the pilot must know his position in a three-dimensional space, a landing method must supply the three dimensional elements of information. In addition, it must meet the following requirements:

a. It must permit landings as safe as those accomplished under visual flight conditions.

b. It must be operable at all times of day and night, during any weather conditions.

c. It must present the landing information to the pilot in a simple, easily understood form.

d. It must have the pilot's confidence.

To accomplish each of the preceding requirements, various equipments have been proposed and developed. These methods may be conveniently classified according to the techniques used, such as radio, radar, television, and various combinations.

11-1017. Over the past 30 years or so, much work has been done in the development of radio equipments for instrument-landing purposes. That work has resulted in a standard landing equipment which is in common use today at all of the major airports. The following discussion will describe the major developments of the past and will conclude with a description of the current common method. During the late 1920's, the Bureau of Standards developed the first instrument-landing method which contained all of the elements found in the present-day arrangement. Although the actual techniques have since been changed, the basic principles of the Bureau of Standards methods were sound enough to serve as a model for all developments to this day. In fact, the contemporary terms localizer and glide path were coined to describe components of the Bureau of Standards method. The Bureau of Standards arrangement required three ground transmitters and two airborne receivers, as follows:

a. A localizer transmitter, which supplied signals by which the pilot determined his position to the right or to the left of the runway.

b. A glide-path transmitter, which supplied a curved path extending from the ground up into space along which the pilot guided his aircraft down to a landing.

c. A marker transmitter, which supplied a signal to make the edge of the runway.

d. A localizer and marker receiver.

e. A glide-path receiver.

11-1018. The localizer transmitter was of the visual-range type, operating at 278 kc, the carrier being modulated at frequencies of 65 cps and 86.7 cps. The transmitter was located at the stop end of a runway so that its course extended directly down the middle of the runway, with a predominance of either 65 or 86.7-cps sidebands on opposite sides of the course. The receiver operated two vibrating-reed indicators, one of which was resonant at 65 cps, and the other at 86.7 cps. The relative amplitudes of vibration of the reeds indicated the relative amounts of sideband energies and, therefore, the heading of the aircraft with respect to the course. To fly the course, it was necessary for the pilot to position his aircraft so that equal amplitudes of vibration were produced by the two reeds. The glide path was radiated from a directive antenna array which was excited at 93.7 mc by a 500-watt transmitter. The path flown to a landing was an isopotential one, defined as the locus in space of an infinite series of points where the glide-path transmitter produces signals of equal field strength, as shown in figure 11-126BI.

11-1019. Figure 11-126BI represents a portion of the vertical radiation pattern of a transmitting antenna in the presence of the reflecting earth, when the antenna is at the origin of the graph and the earth's surface is assumed to be a plane. The assumption is justified when the horizontal distance, or range, involved is less than 50 miles. Each of the lobes represents one of an infinite number of contours of constant field intensity or, in the case of the received signal, one of an infinite number of isopotential con-

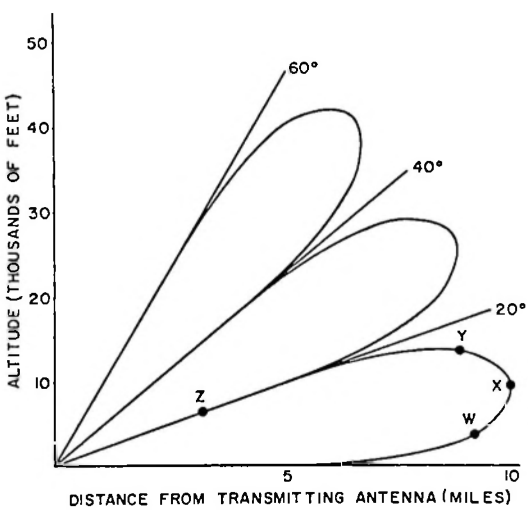


Figure 11-126BI. Vertical Radiation Pattern of Antenna in Presence of Reflecting Earth

tours. Practically, as far as the Bureau of Standards method was concerned, only the lowest lobes were of any consequence, because of the low altitude at which flight was accomplished in the 1920's. (At 10 miles, point X in the figure is approximately 10,000 feet high.) It can be seen from figure 11-126BI that the received-signal intensity is a function of both range (horizontal distance) from the antenna and altitude (vertical distance) above the earth's surface. It can thus be inferred that an aircraft moving toward the antenna, as from point X, would have to ascent along a glide path represented by a line drawn between points X and Y of the figure (XY) or to descent along (XW) as it approached the antenna if a receiver in the aircraft were to receive a constant-intensity signal. The descending path represents the glide path of the Bureau of Standards method. The pilot would obviously reject an ascending path such as XY; similarly, too-steep a path, such as YZ, would be abandoned as unsafe. The airborne glide path receiver used in the Bureau of Standards method operated a field-strength indicator which the pilot observed. Having selected any initial value of field strength as a reference and by maintaining a constant reading on his indicator, the pilot then guided the aircraft along the glide path to a landing. The fan marker used by the Bureau of Standards was a boundary marker, which indicated by an aural signal to the pilot the edge of the airport runway. The marker transmitter operated on the same frequency as the localizer transmitter. Thus, no additional receiving equipment was required for reception of the marker. Continuous development of the Bureau of Standards method caused modifications to the localizer modulation in order to permit the acceptance of the 60-cps and 500-cps frequencies. The localizer receiver was then modified to include narrow-band filters in order to separate the sideband components and to apply their respective outputs to a zero-center differential meter, which was incorporated with the glide-path indicator into one instrument common to both receivers, as shown in figure 11-126BJ. The left-right indication of the vertical needle indicated the heading of the aircraft with respect to the middle of the landing runway. The horizontal needle in figure 11-126BJ, the glide-path indicator, deflected up or down according to the position of the aircraft above or below the glide path. When both needles were perpendicular to each other, the aircraft was on-course and on the glide path. This instrument has evolved into the cross-pointer indicator commonly used today.

11-1020. During the early 1930's, development work on airway radio aids was continued by interested Government and commercial agencies. The major problems connected with landing methods that were investigated during this time were as follows:

a. The low radio frequency of the localizer required large transmitting antenna arrays and subjected the course to bending.

b. The glide path was rather inflexible; hence, it was impossible to alter the steepness of the path without also making the landing point too close to the edge of the runway, although certain runways required steep, controllable paths. The obvious answer to the short-comings of the localizer was to raise the frequency, which was done. In addition, the modulation frequencies were changed to 90 cps and 150 cps, and mechanical modulators were introduced. These innovations were described by the Civil Aeronautics Administration in 1939.

11-1021. Figure 11-126BK shows the horizontal radiation pattern of the localizer used in the CAA instrument landing method. Point X represents the position of the transmitting antenna. The two large lobes contain sidebands of 90 cps and 150 cps. The sideband energies merge toward the interior of the pattern, as indicated by the smaller

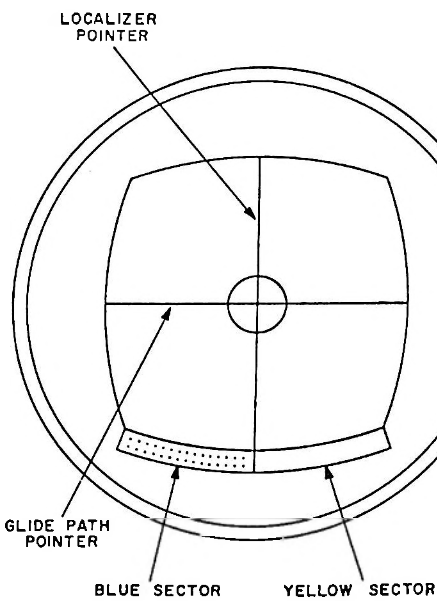


Figure 11-126BJ. Cross-Pointer Instrument for Bureau of Standards Landing System

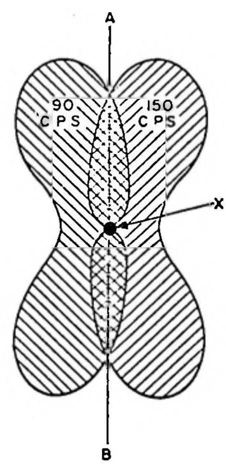


Figure 11-126BK. Horizontal Radiation Pattern of CAA Equisignal Localizer crosshatched lobes. Along the center line, AXB, the amounts of 90 cps and 150 cps sideband energies are equal. Either AX or BX represents the localizer course, which is known as an equisignal course (i.e., the locus of a series of connected points where the field strengths from both energy lobes are equal). To follow the course, a pilot flew his aircraft until the vertical needle of a cross-pointer instrument was straight up and down. Deviation from the course deflected the needle to the right or to the left. The CAA localizer transmitters operated at various frequencies in a narrow band centered at 110 mc. The CAA glide path was a controlled isopotential type, which will not be described further since equisignal glide paths have replaced the isopotential paths. The CAA arrangement employed two fan marker transmitters, one located at the boundary of the airport and the other two miles from the boundary. The transmitters operated at 75 mc and were modulated at 400 and 1300 cps, respectively, to light corresponding lamps in the aircraft cockpit. The CAA arrangement also included a series of calibrated instruments indicating the quantity and quality of the signals being radiated and a set of visual and aural alarms to indicate equipment malfunctioning. In 1938, the CAA introduced an equisignal glide path which was variable between 6 degrees and 11 degrees. The output of this glide-path transmitter, which operated at 700 mc, was applied to two horn radiators. The glide-path angle was varied by varying the tilt angle of the horns. Refer to figure 11-126BL, which shows the vertical field pattern used to produce the glide path. The upper lobe was

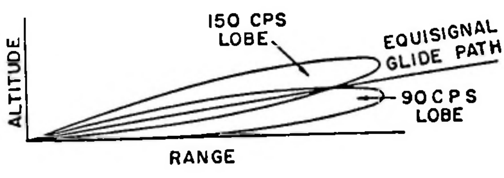


Figure 11-126BL. CAA Equisignal Glide Path

modulated at 90 cps, and the lower at 150 cps. In effect, the pattern is that of the equisignal localizer pattern turned through 90 degrees.

11-1022. Prior to World War II, much of the effort expended by airlines, electronics companies, and Government agencies in developing electronic navigational aids was ineffective because of a lack of coordination. One result of this lack of coordination was that a variety of imcompatible methods came into existence. The lack of uniformity precluded the use of standardized airborne equipment, which, in turn, meant that many pilots were denied the use of the electronic aids. The military requirements of World War II necessitated the adoption by the military of standardized devices. Such a device was the Army Air Force's SCS-51 Instrument Landing Approach method. Similarly, the increased density of air traffic after the war made it necessary for all interested parties to adopt a common set of navigational aids. Among the aids proposed in 1948 by the Radio Technical Commission for Aeronautics was the Instrument Landing System (ILS). This method identical with the SCS-51, was adopted by the CAA, and thus became the standard approach arrangement for all military and civilian aviation activities. ILS is a radio-beam method designed to provide a pilot with precise guidance from an elevated point in space down to a runway. As in earlier landing methods, two beams are used. They are supplied by a localizer transmitter and a glide-path transmitter. The vertical plane of the localizer intersects a horizontal plane that slants from a touchdown point on the runway upward at an angle of approximately 2.5 degrees and beyond the approach end of the runway for a distance of 25 to 60 nautical miles. Fan markers are placed along the course to provide accurate fixes, as shown in figure 11-126BM.

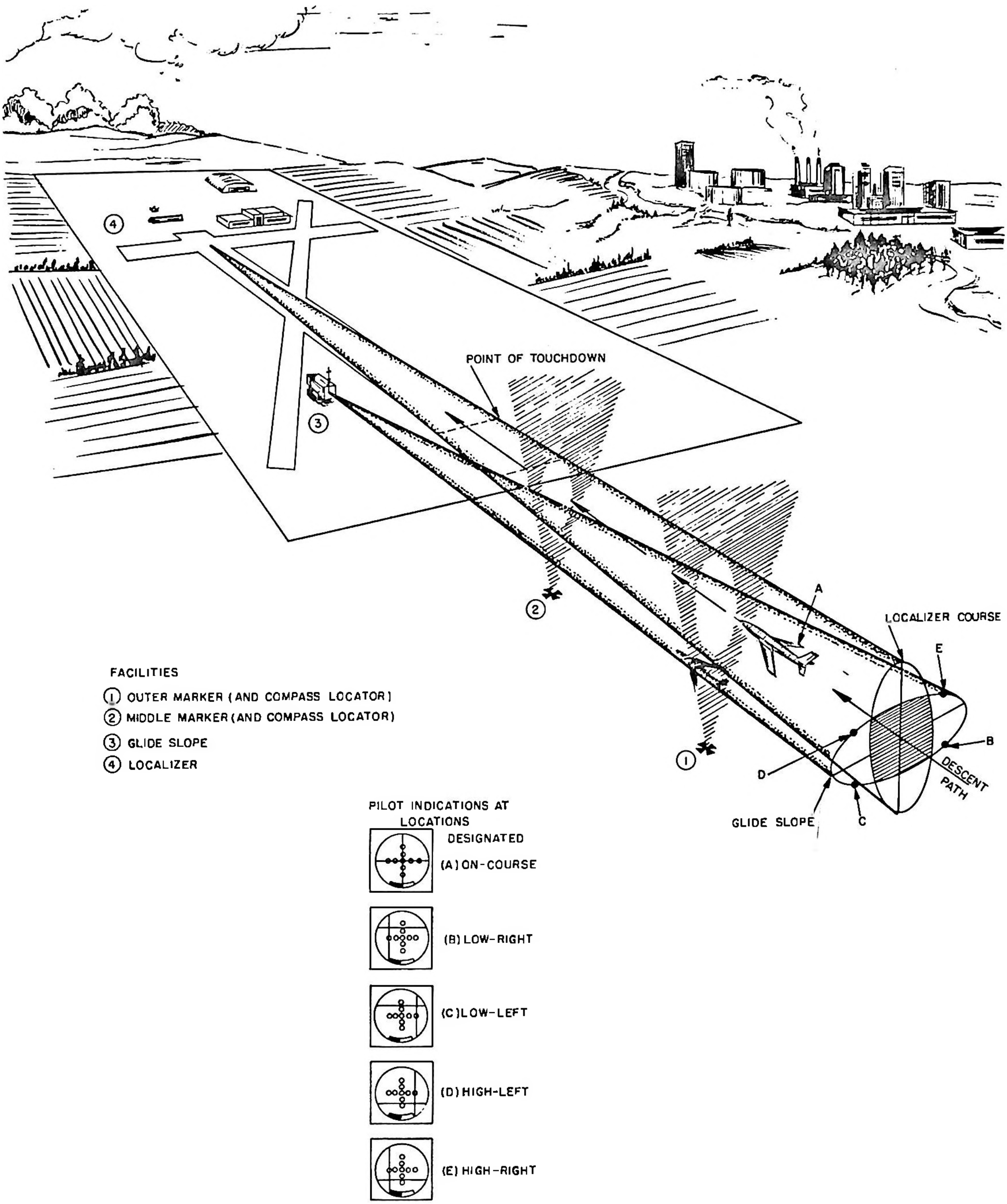


Figure 11-126BM. Instrument Landing Facility

11-1023. Under normal operating conditions, the equipment is functioning in remote operation. This signifies that when one transmitter is placed in "full" operation, the other is placed in "standby" operation. Control lines existing between the remote control point and the equipment permit the equipment to be shut down or started up from the remote point. Also in remote operation the alarm and changeover circuits of the control-indicator can change over from an operating transmitter which breaks down to a standby transmitter in less than a minute if a defect in transmission should occur. The localizer equipment furnishes lateral guidance along the extended center line of the instrument runway. The localizer equipment consists of four major parts: the transmitting equipment, the course (or directional) antenna array, the clearance antenna array, and the field monitors. The glide slope facility furnishes vertical guidance along the approach course as defined by the localizer. The glide slope equipment consists of three major parts: the transmitting equipment, the antenna assemblies, and the field monitors. The localizer is essentially a transmitter and frequency radiating arrangement. The equipment radiates a localizer course by the simultaneous transmission of two separately modulated signals. One of the signals is radiated by a highly directional antenna, known as the course (or directional) array, and forms the desired lateral guidance area. The other signal is radiated by a smaller antenna array, known as the clearance array, and serves the following two purposes:

a. It suppresses false course signals generated by the directional array.

b. It complements the function of the course antenna array by setting up a "back course" and by providing localizer information in the area not covered by the directional array.

11-1023A. ANTENNAS.

11-1023B. GENERAL. There are five antennas involved in the make-up of instrument landing equipment.

11-1023C. The localizer antenna, as shown in figure 11-126BN, is a horizontal array of V-shaped antennas with directors, which radiates two horizontally polarized overlapping field patterns. The field patterns contain predominant modulation products at an audio frequency, and are orientated in such a manner that the intersection of the frequency fields forms equisignal planes along the extended center line of the runway.

11-1023D. The glide slope antenna consists of two antennas mounted one above the other. The lower antenna is a broad-band, horizontally polarized, half-loop antenna and reflector screen placed at such a height that it has a relatively high angle of radiation. The upper antenna consists of a double stack of broad-band horizontally polarized V-shaped antennas with dipole reflectors placed at such a height that it has a relatively low angle of radiation; see figure 11-126BO.

11-1023E. The compass locator antenna is a flattop T-shaped antenna, as shown in figure 11-126BP. The top of the T is a single wire mounted at a predetermined height above the ground.

11-1023F. The marker beacon antenna, as shown in figure 11-126BQ, is made up of two dipole elements and a vertical steel mast.

11-1023G. The course-centering antenna is a half-wave dipole with a parasitic director and a parasitic deflector. This antenna is located on the center line of the runway, approximately 275 feet in front of the localizer antenna array, as shown in figure 11-126BR.

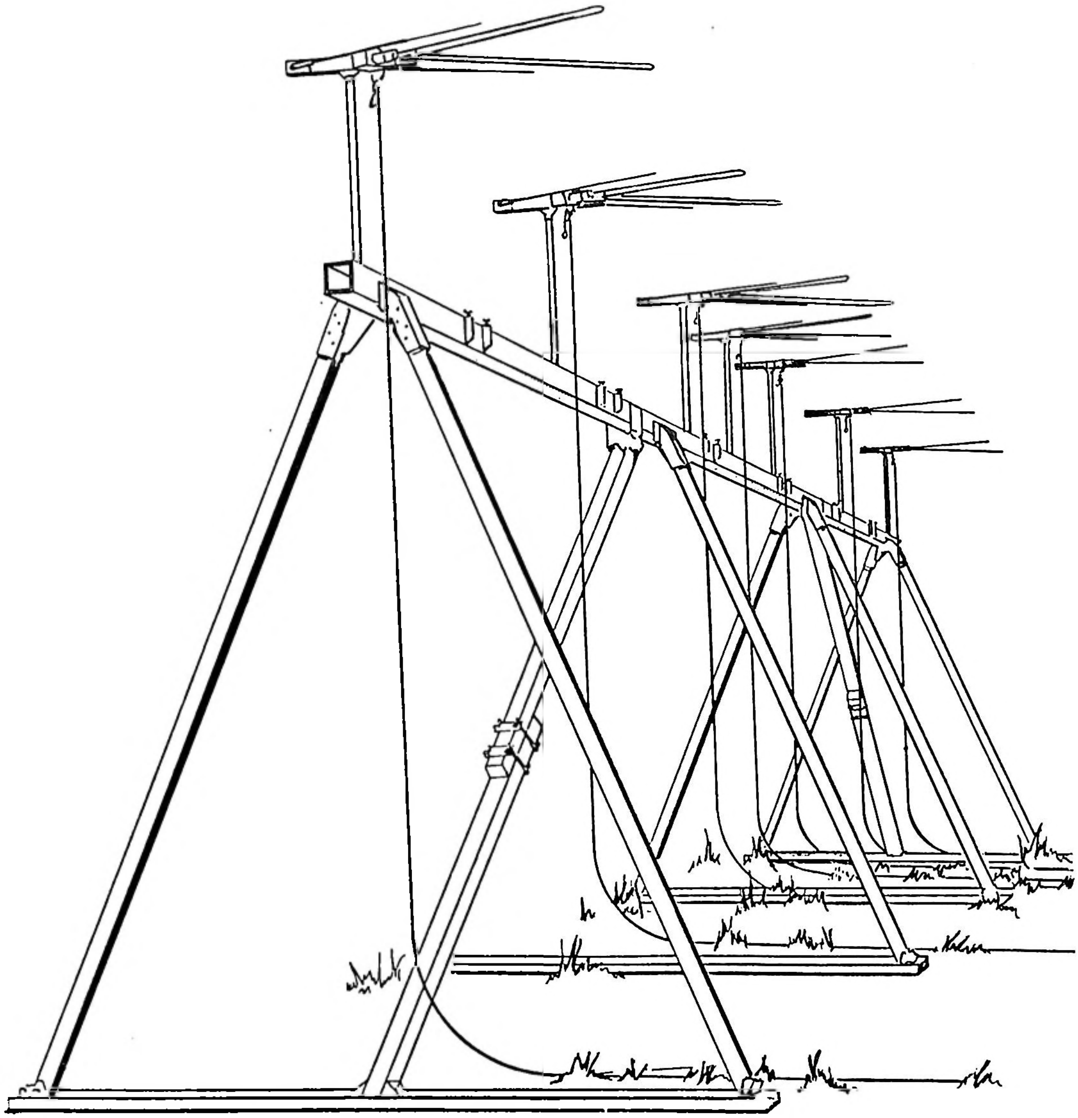


Figure 11-126BN. Horizontal Array of V-Shaped Antennas

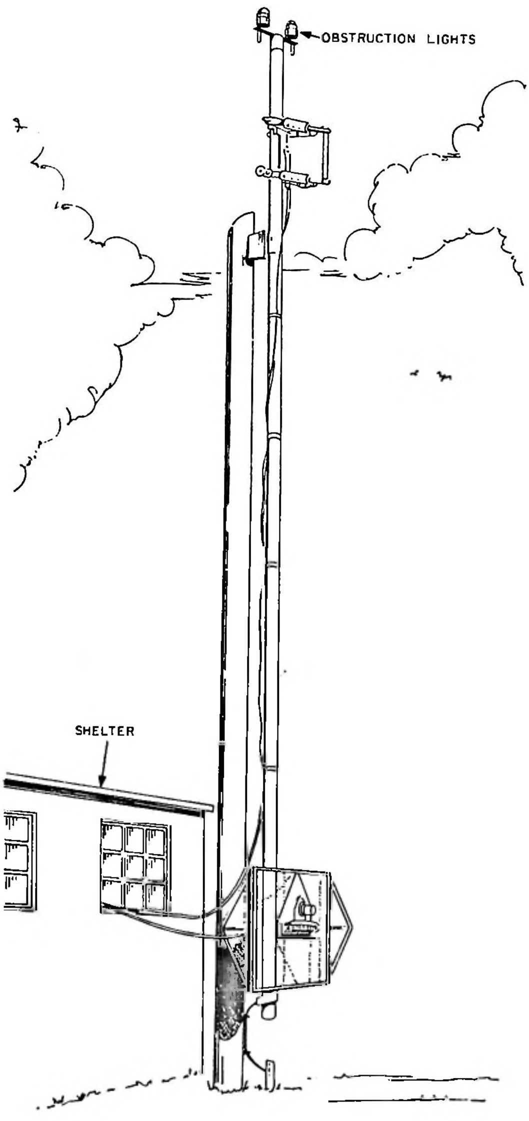


Figure 11-126BO. Glide Slope Antenna Installation

11-1023H. ANTENNA BRIDGE TEST. The antenna test procedure is given below. This procedure is referenced to figure 11-126BS.

a. Connect dummy loads, with a variable standing wave ratio indicator, to jacks J1, J2, and J3 of the antenna bridge, as indicated in part A of figure 11-126BS.

b. Connect jack J4 to the slotted line and make the other connections indicated in part A of the figure.

c. Using a crystal calibrator, calibrate the signal generator.

d. Measure the variable standing wave ratio on the slotted line.

e. Find a voltage minimum along the slotted line. Adjust the signal generator output for a convenient reference reading on the detector. Then leave the signal generator output at this level for the following steps.

f. Reconnect the equipment to the bridge as indicated in part B of the figure.

g. Find a voltage minimum point and compare the output voltage obtained at this point with that obtained in step e.

h. Reconnect the equipment to the bridge as indicated in part A of the figure with the following exceptions: Disconnect the dummy load from J2 and connect it to J4. Connect the output of the slotted line to J2. (This shifts the bridge connections by 90 degrees.) Then repeat steps d through g.

11-1023I. CABLE ASSEMBLY TEST AND ADJUSTMENT. The cable assembly must be tested electrically on a slotted line. This procedure applies to both old and new cable assemblies. If a cable assembly cannot be made to meet the requirements stated below, it should be rejected and a new cable assembly fabricated. The procedure for cutting the cable assembly to the proper length and performing the proper electrical test is given below.

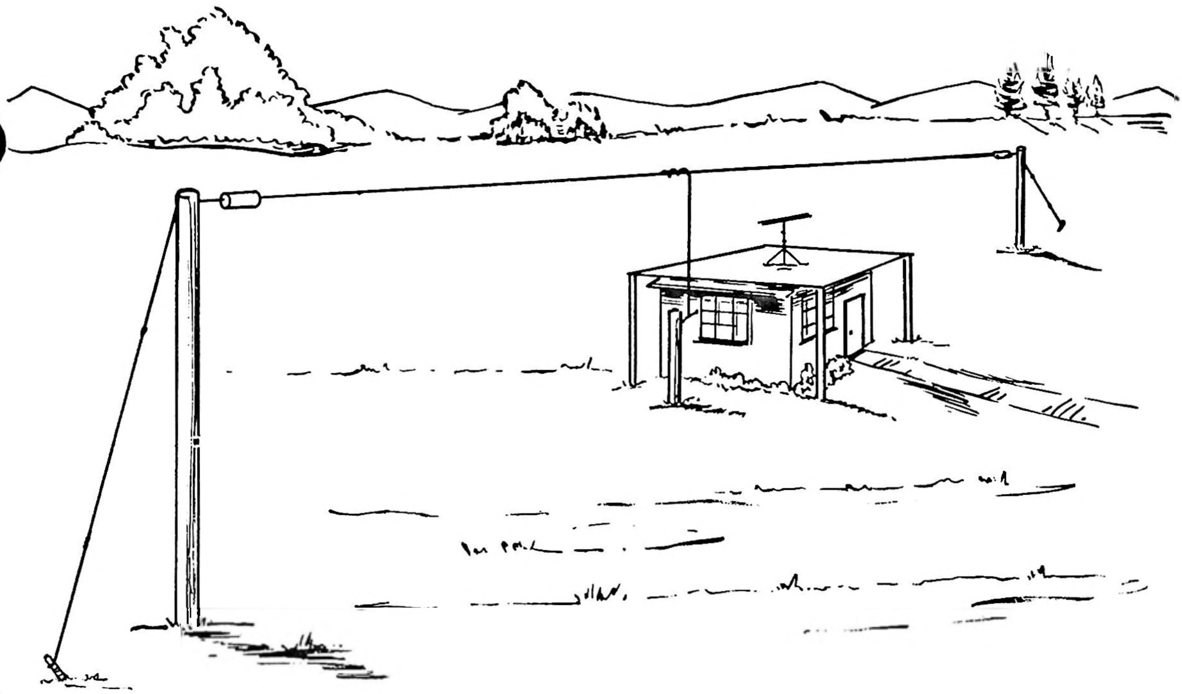


Figure 11-126BP. Compass Locator Antenna

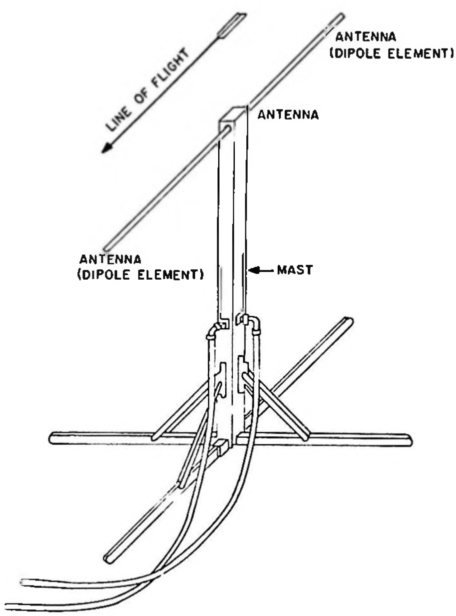


Figure 11-126BQ. Marker Beacon Antenna

a. Measure the cable for the desired length.

b. Connect the test equipment for slotted line measurements, leaving the slotted line open-circuited.

c. Use a crystal calibrator to adjust the signal generator to the appropriate frequency.

d. Accurately determine and record the distance of the first null from the open end of the slotted line. This distance will be referred to as d1 (ref).

e. Accurately determine the distance of the second null from the open end of the slotted line. This distance will be referred to as d2 (ref).

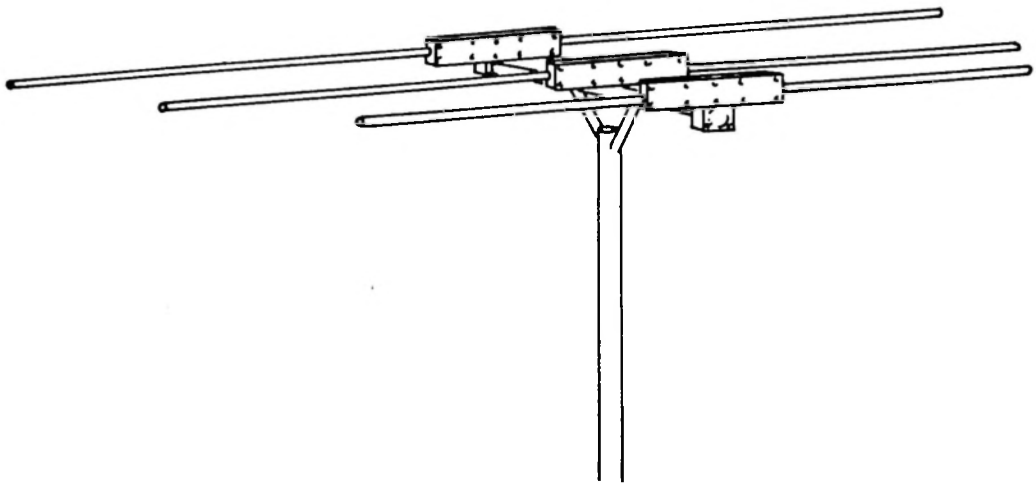


Figure 11-126BR. Course-Centering Antenna

f. Calculate the degrees/cm of the slotted line from the following formula:

Degrees/cm of a slotted line =

$$\frac{180}{d2\ (ref) - d1\ (ref)}$$

g. Compute the cable tolerance, T_1 from the following formula:

$$T_1 = \frac{5}{degrees/cm}$$

h. Connect the cable under test to the end of the slotted line, leaving the cable open-ended.

i. Determine d1 and d2 under this condition. If d1 and d2 fall within T_1 centimeters of d1 (ref) and d2 (ref), the cable is satisfactory. If it does not fall within T_1 centimeters of d1 (ref) and d2 (ref), refer to the cutting instructions in the next steps.

j. If d1 falls on the load side of d1 (ref) and d2 falls on the load side of d2 (ref), the cable is electrically too long and may be cut to the proper length. If d1 falls on the signal input side of d1 (ref) and d2 on the signal input side of d2 (ref), the cable is too short and must be discarded.

k. The length to be cut from a cable that is electrically too long can be determined from the following formula:

Length of cable to be cut (in.) =

$$\frac{degrees/cm\ (slotted\ line)}{15.2\ degrees/in.\ (cable)} \times d - \left[d\ (ref)\right]$$

where d— [d (ref)] is the average value of d1—d1 (ref) and d—d2 (ref) in cm.

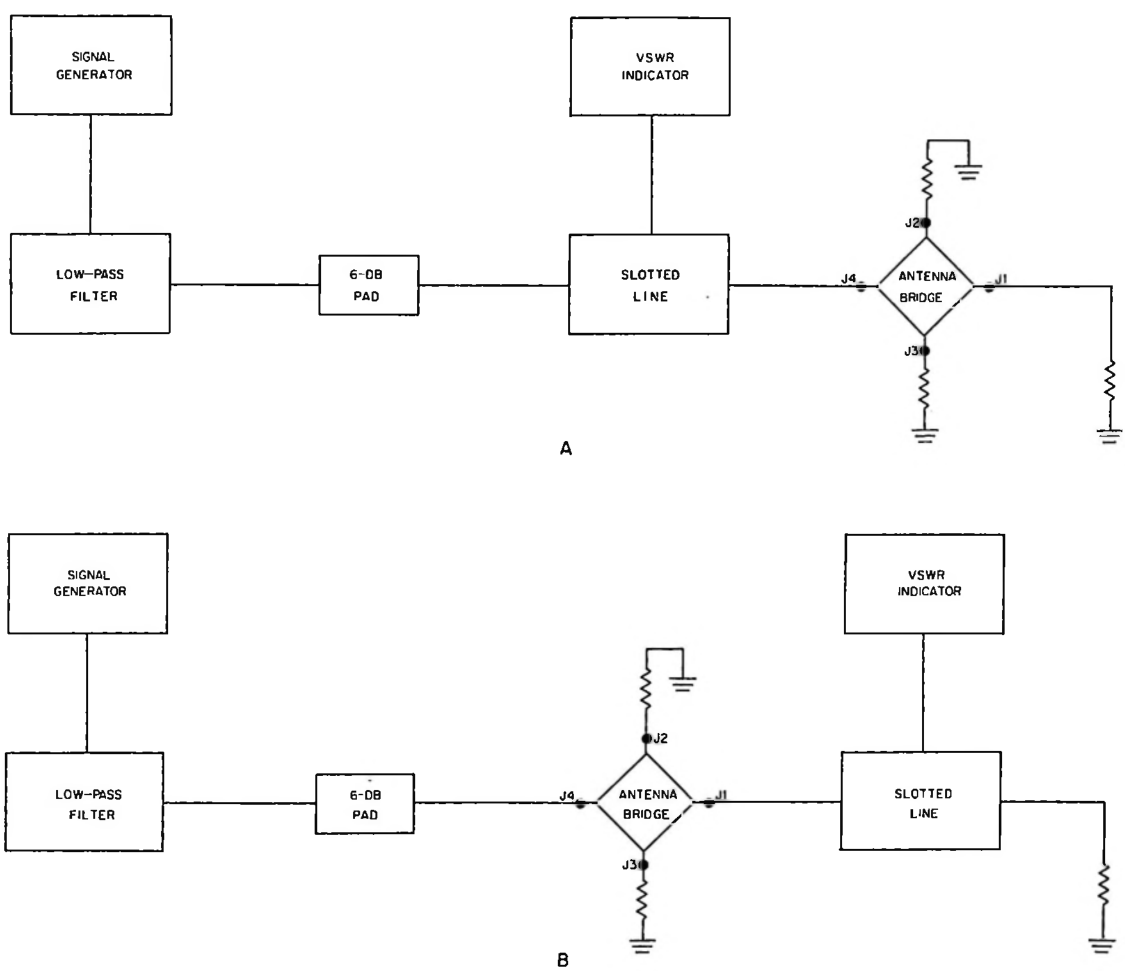


Figure 11-126BS. Antenna Bridge Test Setup

l. Remove the connector from one end of the cable, including the unsoldering of the center pin. (The amount of cable cut should be measured from the end of the center conductor.) Then cut the cable and reassemble the connector to the cable.

NOTE

It is advisable to make the first cut one-half the value calculated. Then repeat step i. If the nulls move the proper amount, the length of the proper amount, the length of the second cut can be calculated from the formula. If the nulls do not move the proper amount, again cut the cable one-half the calculated amount and repeat this procedure until the proper length is obtained.

m. Repeat the procedure given in steps i through l for two other cables.

n. Use a crystal calibrator to adjust the signal generator to the proper frequency, and check d1 and d2 for each cable in a set of three antenna cables. All d1 and d2 distances measured should be within T_1 centimeters of each other. Replace any cable which does not meet this requirement.

o. Repeat step n at other desired frequencies.

p. Check the physical length of one set of antenna cables. Replace any cables that do not meet the required length.

q. Terminate each cable with a dummy load and perform the slotted line measurements. Replace any cable that does not meet the requirements.

11-1023J. FLUTED LINE CHECK. The fluted line is checked as follows:

a. Connect the test equipment for slotted line measurements, leaving the slotted line open-circuited.

b. Use a crystal calibrator to adjust the signal generator to the proper frequency.

c. Accurately determine and record the distance of the first null from the open end of the slotted line. This distance will be referred to as d1.

d. Accurately determine the distance of the second null. This distance will be referred to as d2.

e. Connect the fluted line to the open end of the slotted line, and record the nearest point of minimum voltage from the open end of the slotted line to d1. This reading will be referred to as d3.

f. Determine the electrical length of the fluted line from the following formula:

$$L_e = \text{electrical length in degrees} =$$

$$360 \text{ degrees} \pm \frac{180\ (d3-d1)}{d2-d1}$$

g. The electrical length of the fluted line should be 360 $\pm$10 degrees.

11-1023K. MODULATOR.

11-1023L. This unit modulates the frequency output of the transmitter with 90-cps and 150-cps audio components, and separates the resultant signals for distribution to the antenna equipment.

11-1023M. CROSS-MODULATION ADJUSTMENT. Perform the following adjustments to minimize cross modulation between the 90-cps and 150-cps modulation components for both the course and clearance transmitter sections.

a. Place the oscilloscope into operation; then make the necessary connections and perform the necessary adjustments to obtain the coarse 150-cps foldover audio pattern or as a visual presentation. This oscillogram can be used in adjusting the minimum cross modulation in the course transmitter section. Figure 11-126BT is a sample of the desired pattern. The peaks of this waveform are all of equal amplitude. Adjust the course section control for cross modulation to obtain the desired waveform.

b. Adjust the course transmitter power amplitude plate control for a maximum indication on the grid meter in the proper switch position. Then repeat step a to readjust for minimum cross modulation.

c. Make the oscilloscope connections and perform the necessary adjustments for

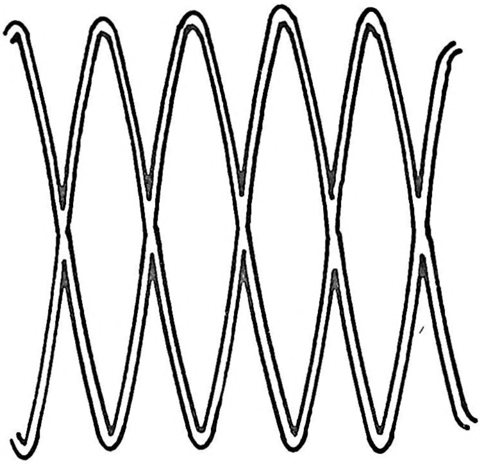


Figure 11-126BT. Normal 150-Cycle Foldover Pattern, Audio Presentation

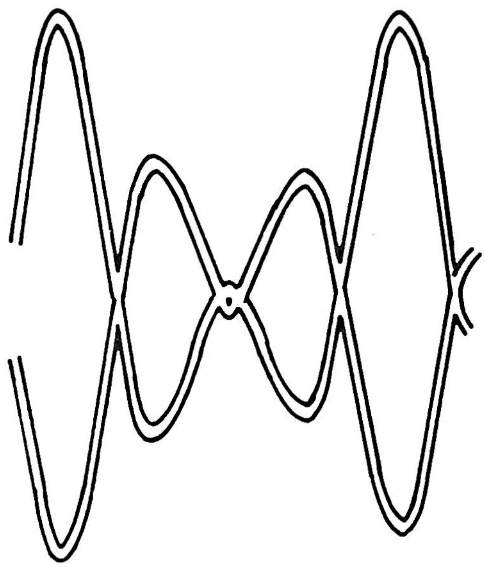


Figure 11-126BU. Normal Carrier Foldover Pattern, Audio Presentation

displaying the clearance carrier foldover audio pattern as a visual presentation. This oscillogram can be used in adjusting for minimum cross modulation in the clearance section. A sample of the desired pattern is shown in figure 11-126BU. Note here that the zero point of the center of the folded carrier trace is coincident with the zero axis of this pattern.

d. Readjust the clearance power amplifier plate control for a maximum indication on the grid meter in the proper switch position. Then repeat step c to readjust for minimum cross modulation.

11-1023N. CONSTANT FREQUENCY CONTROL CHECK AND ADJUSTMENT. The constant frequency control unit should be tested, using the modulator motor as a load. Reference should be made to the equipment service book for aligning and/or adjusting the following:

a. Align the constant frequency control unit as described in the service handbook.

b. Check the adjustment of the overload circuit as described in the service handbook.

c. Check the percent regulation of the unit by turning off the power and connecting a sufficient load between the proper terminals; a 150-volt, 430-ohm, 525-watt rheostat should be sufficient. Set the rheostat to approximately 35 ohms.

d. Turn the power on and measure the ac voltage across the load. This is the high output load voltage.

e. Adjust the load to a higher resistance (approximately 325 ohms), and note the output voltage. This is the low output load voltage.

f. Calculate the percent regulation from the following formula: Percent regulation = 100 x (low load output voltage) - (high load output voltage) all divided by the high load output voltage.

g. The percent regulation should not exceed +3 percent or −6 percent.

11-10230. POWER OUTPUT AND CARRIER ADJUSTMENTS. Make the following adjustments at the course and clearance transmitter sections to obtain the proper carrier power output:

a. Connect a wattmeter to the DIRECTIONAL CARRIER-ANTENNA jack.

b. Turn on the transmitter and retune the course power amplifier stage by adjusting the load on the power amplifier plate for a peak grid meter reading in the proper switch position.

c. Make the connections and perform the adjustments necessary to display the course carrier foldover audio pattern as a visual presentation. The desired configuration is a trace whose peaks just begin to flatten. To obtain the desired pattern, slowly adjust the course transmitter section output coupling loop for a carrier output where the top of the carrier pattern just starts to flatten.

d. Repeat the course section power amplifier tuning and cross-modulation adjustments. (This is done because it is necessary to swing out the transmitter-modulator panel in most units to adjust the output coupling loop, and the closing of the panel could jar the previous settings.)

e. Check the course output coupling loop control setting. After tuning, it will probably be necessary to reset the control to again obtain the condition where the peaks of the course carrier foldover pattern just start to flatten.

f. Repeat steps c through e until no further adjustment of the course output coupling loop control is necessary.

g. Record the course carrier output power indicated by the wattmeter.

h. Disconnect the wattmeter and replace the cable.

i. Connect the wattmeter to the CLEARANCE CARRIER-ANTENNA jack.

j. Start up the transmitter and retune the clearance power amplifier stage by adjusting the load on the power-amplifier plate for a peak grid-meter reading in the proper switch position.

k. Make the connections and adjustments necessary to display the clearance carrier foldover audio pattern as a visual presentation. The configuration desired is the same as that described for the course carrier foldover pattern in step c. To obtain the desired pattern, slowly adjust the clearance transmitter section output coupling loop.

l. Repeat the clearance section power amplifier tuning and cross-modulation adjustments.

m. Check the clearance output coupling loop control setting. (This is done because in most units it is necessary to swing out the transmitter-modulator panel to adjust the output coupling loop, and the closing of the panel may jar the previous settings.)

n. Repeat step m until no further adjustment of the clearance output coupling loop control is necessary.

o. Record the clearance carrier output power indicated by the wattmeter.

p. Calculate 80 percent of the smaller of the two power output readings recorded in steps g and o.

q. Referring to the wattmeter scale, reduce the clearance transmitter output coupling (by adjusting the clearance coupling loop control) to obtain a power output reading equal to that calculated in the preceding step.

r. Recheck, and readjust if necessary, the clearance power amplifier tuning and cross-modulation adjustment.

s. Disconnect the wattmeter and replaced cable.

t. Connect the wattmeter to the DIRECTIONAL CARRIER-ANTENNA jack.

u. Referring to the wattmeter scale, reduce the course transmitter output coupling (by adjusting the course coupling loop control) to obtain a power output reading equal to that calculated in step p.

v. Recheck, and readjust if necessary, the course power amplifier tuning and cross-modulation adjustment.

11-1023P. PERCENT MODULATION ADJUSTMENTS. The percent modulation adjustments are as follows:

a. Obtain the 150-cycle audio pattern on the oscilloscope. Adjust the setting of the M150 percent modulation control for a low reading (approximately 25). Then slowly increase the setting of this control until a condition of 100-percent modulation is obtained. This condition is observed on the oscilloscope when the bottom peak of the 150-cycle pattern just begins to flatten. This should occur for a dial reading of between 70 and 100.

b. In the event that 100-percent modulation is obtained before the 70 mark, a condition of overmodulation exists. Correct this condition by moving the shorting bar slightly away from the stator blades. Recheck the point at which 100-percent modulation occurs. Repeat this procedure until a flattening occurs at the proper point.

c. In the event that the pattern shows no flattening by the time the 100 mark on the dial is reached, a condition of undermodulation exists. Correct this condition by following the procedure in step b, except that the shorting bar should be moved slightly <u>toward</u> the stator blades.

d. Repeat the adjustments of steps a through c (as applicable) for the 90-cps trough.

e. Repeat all adjustments made on the modulator thus far until no readjustment of any control necessary to achieve the desired results.

11-1023Q. SIDEBAND PATTERN ADJUSTMENTS. The sideband pattern adjustments are as follows:

a. Adjust the M150 control for 100-percent modulation.

b. Connect a wave analyzer to the 150-cycle jack. Tune the wave analyzer to 150 cycles and adjust it so that a meter reading of 100 is obtained. This normalizes the wave analyzer so that its meter reading can be used to determine percent modulation.

c. Set the M150 control for a meter reading of approximately 95 on the wave analyzer. Obtain a visual presentation of the audio foldover sideband pattern on the oscilloscope, and observe it for coincidence of the major peaks, as shown in figure 11-126BV. If it does not show coincidence,

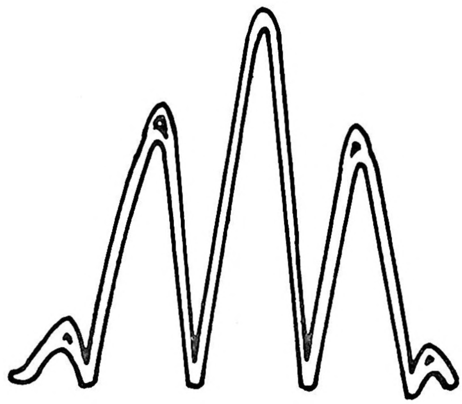


Figure 11-126BV. Normal Sideband Foldover Audio Pattern, Visual Presentation

vary the M90 control until coincidence of the major peaks is obtained. The resulting reading of the M90 dial should be within 40 divisions of the 150-cycle control adjustment.

d. Recheck the transmitter tuning, the setting of the cross-modulation control, the power output, and the phaser settings.

e. Repeat any necessary procedures until all conditions contained in these procedures are met with no readjustment of controls.

11-1023R. MODULATION BALANCE. The purpose of the modulator alignment procedure is to obtain coincidence of major peaks of the foldover sideband pattern, as shown in figure 11-123BV, at the same time that a reading of 0 is obtained on the control indicator modulation balance meter.

a. Perform the procedures given in paragraphs 11-1023O, 11-1023P, and 11-1023Q.

b. Adjust the M90 and M150 percent modulation controls until the modulation balance meter on the control indicator reads zero. This should occur between dial settings of 20 and 80 on both controls.

c. Observing the foldover sideband pattern, reduce the percentage modulation of the signal that results in coincidence of the sideband pattern.

d. Observe the modulation balance meter reading. The reduction in percentage modulation of one signal results in a modulation balance meter reading that indicates an increase in the other. The level of that signal whose percentage modulation was reduced is too high; the blade spacing of this signal must be increased relative to the blade spacing of the other signal.

e. If the 150-cycle signal was reduced in the previous step, increase the 150-cycle blade spacing by inserting the next larger-sized shim. If the 90-cycle signal was reduced in the previous step, decrease the 150-cycle blade spacing by replacing the existing shim with the next smaller-sized shim.

f. Repeat steps b and c until the difference between the peaks of the sideband pattern is no more than 10 percent for a control-indicator meter reading of zero and with the percent modulation dials set between 20 and 80.

11-1023S. CONTROL INDICATORS.

11-1023T. GENERAL. The control indicator receives monitoring signals from the position, width, and clearance field monitors. As long as all monitor signals indicate normal localizer or glide slope transmission, the control indicator provides a normal indication. When the monitoring signals indicate an abnormal localizer or glide slope transmission, the control indicator shuts down the main transmitter and indicates abnormal operation. It then switches operation to the standby transmitter. The control indicator employs three bridge circuits: one in the position monitoring circuit, one in the width monitoring circuit, and one in the shape monitoring circuit.

11-1023U. PATH POSITION, WIDTH, AND SHAPE ADJUSTMENTS. When all antennas are connected and radiating, proceed with the following adjustments:

a. On the modulator front panel, set the path shape control to 70. Note that the control-indicator modulation balance meter deflects to the 90-cycle (left) side of the meter when its switch is in the shape channel position. Vary the modifier phaser on the modulator front panel to obtain a maximum modulation balance meter deflection on the 90-cycle (left) side. If the meter needle was off-scale by the adjustment, draw it back on-scale by decreasing the setting of the path shape control slightly.

b. Lock the modifier phaser control at the position where it yields maximum 90-cycle (left) side deflection of the modulation balance meter.

c. Slowly decrease the setting of the modulator path shape control until a zero rating is obtained on the control-indicator meter.

d. Recheck the transmitter tuning and cross-modulation adjustments.

e. Set the modulation balance switch on the control indicator to the position channel position and check the modulation balance meter to determine whether it still indicates zero. If the indication is slightly off zero, carefully readjust the height of the position field monitor until a zero indication is obtained.

f. Set the modulation balance switch to the width channel position and note whether the meter deflects to the 100-microampere indication. If the reading is slightly off, carefully adjust the modulator path width control until the proper reading is obtained.

g. Set the control-indicator modulation balance switch to the shape channel position and observe whether the meter is reading zero. If the meter needle is slightly off the zero reading, carefully adjust the modulator path shape control until a zero indication is obtained.

h. Lock all controls.

11-1023V. ALARM CALIBRATION. This procedure is given to check the functional operation of the monitoring equipment in order to see that alarms will be activated for deviations in the transmitted signal. The alarms are factory-set and will be checked for functional operation by introducing error signals of sufficient magnitude to trigger the various alarms. The five alarms to be checked are as follows:

a. Path position alarm

b. Path width alarm

c. Path shape alarm

d. Rf level alarm

e. Modulation level alarm

In addition, it is necessary to check the time interval that an alarm indication must exist before the control indicator will initiate a transfer.

11-1023W. PATH POSITION ALARM CALIBRATION. The path position alarm is adjusted so that it will be triggered by any change in the glide slope exceeding 6 percent above or below its normal height. In order to obtain the proper error signal, one above or one below the normal path position, the position field monitor must be raised or lowered from its normal height by 6 percent.

SECTION V

TELEVISION EQUIPMENT TESTING

11-1024. GENERAL.

11-1025. Television can be defined simply as the transmission and reception of visual images by means of electrical signals from one point to another. All television methods are based on the use of the following three basic elements: (1) a pickup device (transducer) to convert an optical image to an electrical signal, (2) a means of transmission of the electrical signal to distant points, and (3) a reproducing device to reconvert the electrical signal into an optical image at the distant location. Pickup devices presently in use are the image orthicon, vidicon, image dissector, iconoscope, and the flying spot scanner. Transmission devices include coaxial lines, low or high power transmitters, or microwave relay links. The reproducing device generally is a standard home-type television receiver, a television monitor, or some form of electro-optical projection system. The first large-scale application of television was in commercial broadcasting. The technical requirements imposed upon the terminal equipment for commercial broadcasting are fairly well standardized, because the objective of each television station is the same, that is, to produce a program suitable for transmission to home receivers. In closed-circuit television, however, the technical requirements vary widely from one application to another. Transmission is by wire line or point-to-point radio. Technical standards do not necessarily have to conform to the specifications governing broadcast television. The requirements for picture quality may be much lower than for broadcast television, for example, when it is desired merely to read a meter at a distance; or they may be much higher, for example, when highly detailed map information is to be transmitted. Another difference between commercial and closed-circuit television is that a greater variety of equipments and facilities are employed in closed-circuit television. Installations vary from a simple camera-monitor combination to highly complex combinations of equipment including many cameras, large distribution and switching equipment, and numerous viewing monitors. Furthermore, closed-circuit television equipment must operate under widely varying environmental conditions which cannot always be controlled.

11-1026. TELEVISION FUNDAMENTALS.

11-1027. SCANNING METHODS.

11-1028. Television transmission of a picture from one point to another utilizes a process in which light, reflected from an object or scene, is converted to electrical impulses of varying magnitude. This is accomplished by means of synchronized scanning, in which the picture is examined by the camera and reproduced by the viewing monitor, point by point, in a regular pattern. This process is carried out so rapidly that the entire picture is scanned many times a second, and the eye sees it as a single complete image.

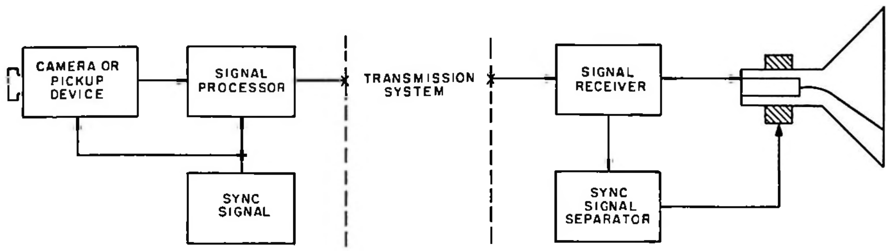


Figure 11-127. Basic Television System

11-1029. A basic television system is shown in figure 11-127. An image of the scene is focused on the camera pickup tube. An electron beam in the pickup tube scans the optical image and produces an electrical signal which varies in amplitude with the amount of light falling on each point on the image. A synchronizing signal is then added to the electrical picture signal from the camera and the composite video signal is transmitted to the display monitor. Within the monitor, the synchronizing signal causes the electron beam to scan the kinescope (television cathode-ray tube) faceplate in synchronism with the camera scanning beam. The intensity of the kinescope beam is varied in accordance with the picture signal and the image appears on the face of the kinescope. The time required for one vertical scan of the picture in television systems in the United States is 1/60 of a second or a multiple or submultiple thereof. The rate of 60 cycles per second was chosen because in the United States most of the power sources are 60-cycle systems, and synchronization (or even near synchronization) with the power frequency reduces the visible effects of hum and simplifies the problem of synchronizing film projectors with scanning. The number of scanning lines determines the maximum ability of the system to resolve fine detail in the vertical direction. The number of scanning lines is also related to resolution ability in the horizontal direction in that, for a given video bandwidth and frame time, horizontal resolution is inversely proportional to the number of scanning lines. Therefore, as the lines are increased in number, the bandwidth of the system must also increase in the same ratio to maintain the same resolution in the horizontal direction. If, as is usually the case, it is desired to maintain approximately equal values of horizontal and vertical resolution, the bandwidth requirements increase as the square of the number of scanning lines. The present system of 525 lines was chosen for broadcast television as the most suitable compromise between channel width and picture resolution. This number of lines has also been found satisfactory for, and is used in, most closed-circuit television systems.

11-1030. NON-INTERLACED. The simplest scanning method is called non-interlaced or sequential (sometimes called "progressive".) This scanning method uses an electron beam which moves very rapidly from left to right on an essentially horizontal line and travels slowly from the top to the bottom of the picture. When the electron beam reaches the end of a line, a blanking voltage is applied which shuts off the beam. This period of time is known as the horizontal retrace period or "flyback" time. Similarly, when the beam reaches the bottom of the picture, the beam is blanked out and reappears at the top of the picture.

11-1031. INTERLACED. An important variation of the scanning method discussed above is interlaced scanning, which is used

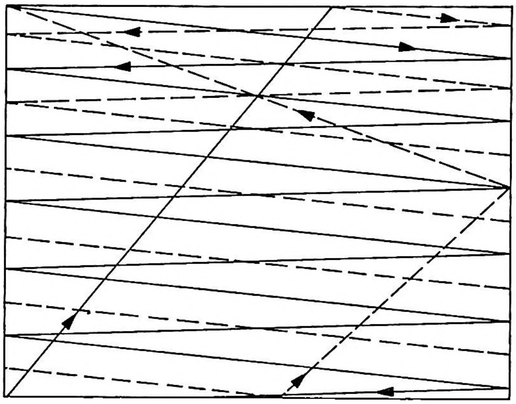


Figure 11-128. Interlaced Scanning

in broadcast television and most, though not all, closed-circuit equipment. With interlaced scanning, it is possible to reduce the video bandwidth by a factor of two without reducing the resolution or seriously increasing flicker. In the standard two-to-one method of interlacing, alternate lines are scanned consecutively from top to bottom, after which the remaining alternate lines are likewise scanned. This principle is illustrated in figure 11-128, which shows interlaced scanning with 13 scanning lines. In this kind of scanning, each of the two groups of alternate lines is called a "field," while the complete set of lines consisting of two consecutive fields, is called a "frame." Interlacing is accomplished by making the total number of lines in a frame an odd integer. Thus the number of lines in each of the two fields is a whole number plus a half. This results in consecutive fields which are displaced in space with respect to each other by half a line, thus producing interlacing of the lines. In the actual method used for broadcast and most closed-circuit television, the total number of lines is 525, the total per field is 262-1/2, the vertical scanning frequency is 60 cycles per second, the number of complete pictures (frames) per second is 30, and the horizontal scanning frequency is 15, 750 cycles per second (60 x 262-1/2).

11-1032. SIGNALS.

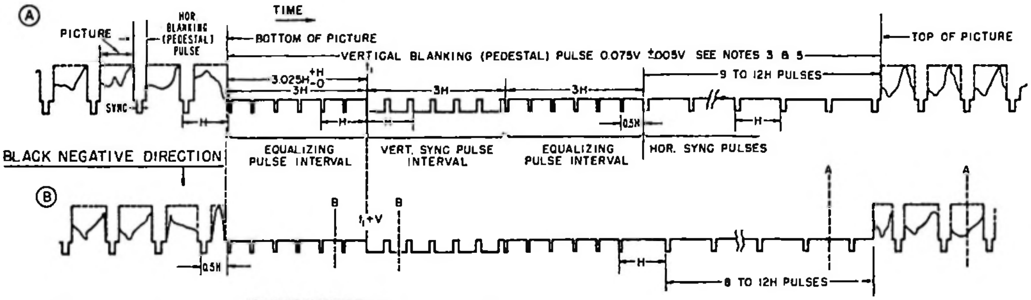
11-1033. GENERAL. The standard television signal comprises four elements: the picture information generated during active scanning time, picture blanking pulses, picture average dc component, and picture synchronizing pulses.

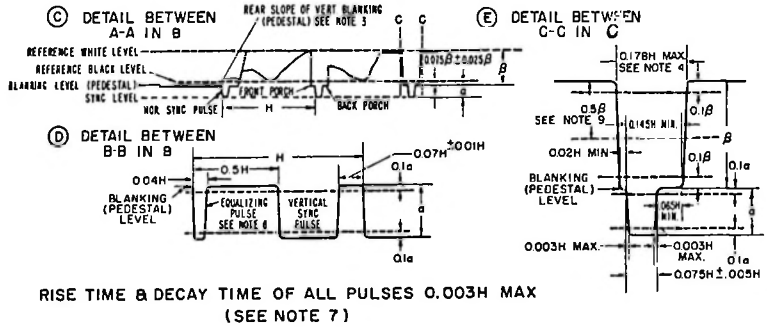
11-1034. PICTURE INFORMATION. The basic part of the signal, the picture information, is a series of waves and pulses generated during the active line scanning of the pickup or camera tube. As a scanning line traverses the face of the pickup tube, it is modulated in amplitude in proportion to the brightness variations in the scene it is scanning. The signal produced varies in amplitude proportionally with the brightness of the scene. For commercial broadcasting the amplitude variations are such that the maximum video amplitude produces black; the minimum video amplitude produces white. Ordinarily the maximum and minimum video amplitude values represent 75 percent and 15 percent, respectively, of the maximum carrier voltage.

11-1035. PICTURE BLANKING PULSES. During retrace time, in order to prevent undesirable signals from entering the picture, blanking pulses are applied to the scanning beams in both the camera tube and the receiver kinescope. Camera blanking pulses are used only in the pickup device. They serve only to close the scanning aperture in the camera tube during retrace periods and never actually appear in the final signal sent to the receiver. (Most low-cost closed-circuit or industrial camera chains do not use a special camera blanking pulse; the same pulse used to trigger the scanning circuit and blank the kinescope is used to close the camera scanning aperture.) The function of the kinescope blanking pulses is to suppress the scanning beam in the kinescope during both vertical and horizontal retrace time. They are simple rectangular pulses, somewhat wider than the corresponding camera blanking pulses used only in broadcasting and similar applications. They have a duration slightly longer than the actual retrace periods in order to trim up the edges of the picture and to provide a clean, noise-free period during retrace. The blanking signal shown in (B) of figure 11-129 contains pulses for the removal of visible lines during both horizontal and vertical retrace periods. The horizontal pulses recur at intervals of 1/15, 750 of a second (monochrome broadcasting). At the bottom of the picture they are replaced by vertical blanking pulses which are similar to the horizontal pulses, except that they are of much longer duration (approximately 15 scanning lines) and have a periodic recurrence of 1/60 of a second. It is important to note that the blanking pulses (and synchronizing pulses) are added at a relatively high-level point in the transmitter and are, therefore, considered noise-free. The importance of noise-free blanking and synchronizing pulses should not be underestimated. They determine the stability of the viewed picture or determine the degree to which a picture will remain locked-in on a kinescope under even the most adverse transmission conditions. This point is especially important when considering the use of television for closed-circuit applications. The extreme environmental conditions which may be encountered can seriously degrade the picture signal, making it difficult to synchronize or lock-in a picture unless the original blanking-to-picture signal-to-noise ratio is high. Details of the horizontal blanking pulse are shown in (E) of figure 11-129; details of the vertical blanking pulses are shown in (A) and (B) of figure 11-129. Limitations of the vertical pulses are not determined so much by circuit considerations as by the requirements for satisfactory operation with intermittent television film projectors.

11-1036. PICTURE AVERAGE DC COMPONENT. An important feature necessary in a video signal, if a television picture is to be transmitted successfully with the necessary fidelity, is the dc component of the picture signal. This component is a result of slow changes in light intensity. The loss of the dc component is a result of ac or capacitive coupling circuits and is evidenced by the picture signal tending to adjust itself about its own ac axis. The dc component is returned to the video signal by means of a dc restorer or inserter circuit.

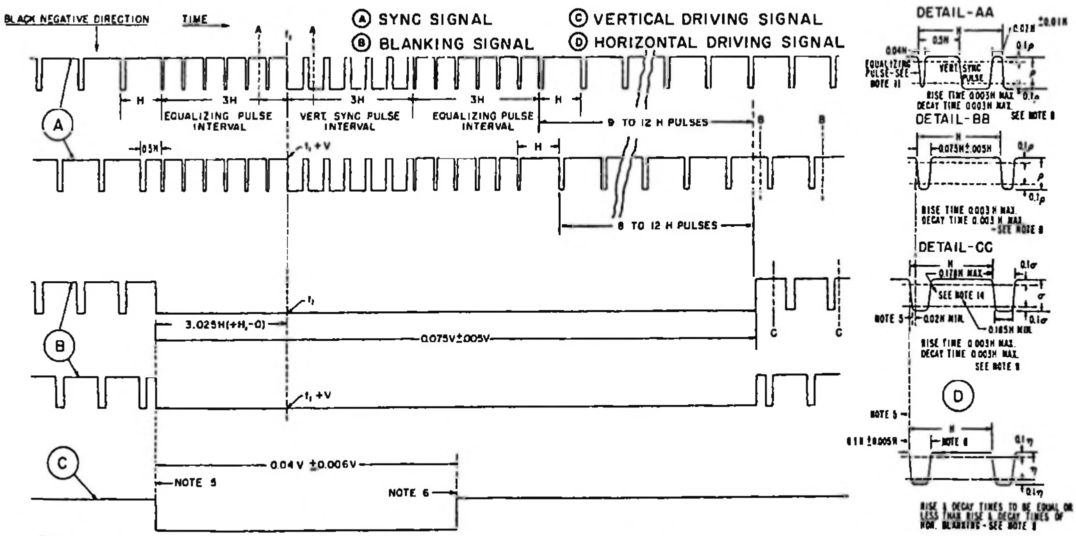
11-1037. SYNCHRONIZING. Synchronization of the scanning beams in the camera and the receiver must be exact at all times in order to provide a viewable picture. To accomplish this, synchronizing information is provided in the form of electrical pulses in the retrace intervals between successive lines and between successive pictures. The retrace periods (which are as short as circuit considerations permit) are areas in which synchronization pulses may be inserted without interfering with the picture. Synchronizing pulses are generated at the program origin end of the television system, in the equipment that controls the timing of the scanning beam in the pickup tube. They become a part of the complete signal which is transmitted to the receiver or monitor. In this manner, scanning operations in both ends of the television system are always in step with each other. In general, synchronizing signals should provide positive synchronization of both horizontal and vertical sweep circuits. They should be separable by simple electrical circuits to recover the vertical and horizontal components of the composite synchronizing signal, and be able to combine simply with the picture and blanking signals to produce a standard composite television signal. Most televi-





NOTE:

1. H-TIME FROM START OF ONE LINE TO START OF NEXT LINE.
2. V-TIME FROM START OF ONE FIELD TO START OF NEXT FIELD.
3. LEADING AND TRAILING EDGES OF VERTICAL BLANKING (PEDESTAL) SHOULD BE COMPLETE IN LESS THAN 0.1H.
4. LEADING AND TRAILING EDGES OF HORIZONTAL BLANKING (PEDESTAL) MUST BE STEEP ENOUGH TO PRESERVE MIN. & MAX. VALUES OF DURATIONS UNDER ALL CONDITIONS OF PICTURE CONTENT.
5. ALL TOLERANCES AND LIMITS SHOWN IN THIS DRAWING ARE PERMISSIBLE ONLY FOR LONG-TIME VARIATIONS.
6. EQUALIZING PULSE AREA SHALL BE BETWEEN 0.45 AND 0.5 OF THE AREA OF A HORIZONTAL SYNC PULSE.
7. ALL RISE AND DECAY TIMES TO BE MEASURED BETWEEN 0.1 AND 0.9 AMPLITUDE REFERENCE LINES.
8. THE OVERSHOOT ON BLANKING (PEDESTAL) SIGNAL MUST NOT EXCEED 0.02β AT THE BEGINNING OF THE FRONT PORCH AND MUST NOT EXCEED 0.05β AT THE END OF THE BACK PORCH. THE OVERSHOOT ON SYNC SIGNALS MUST NOT EXCEED 0.05α. ANY OTHER EXTRANEOUS SIGNALS ON THE BLANKING (PEDESTAL) SIGNAL MUST NOT EXCEED .02β.
9. FOR SETTING ASPECT RATIO, THE HORIZONTAL BLANKING (PEDESTAL) DURATION SHOULD BE SET TO 0.175H AT THE 0.5β POINT.
10. RISE TIME AND DECAY TIME OF HORIZONTAL BLANKING MUST NOT EXCEED 0.003H.



NOTE:

1- H-TIME FROM START OF ONE LINE TO START OF NEXT LINE.
2- V-TIME FROM START OF ONE FIELD TO START OF NEXT FIELD.
3- LEADING AND TRAILING EDGES OF VERTICAL DRIVING AND VERTICAL BLANKING SIGNALS SHOULD BE COMPLETE IN LESS THAN 0.1H.
4- ALL TOLERANCES AND LIMITS SHOWN IN THIS DRAWING ARE PERMISSIBLE ONLY FOR LONG TIME VARIATIONS.
5- TIMING ADJUSTMENT, IF ANY, MUST INCLUDE THIS CONDITION.
6- THE VERTICAL DRIVING PULSE DURATION SHALL BE 0.04V ±0.006V. THE HORIZONTAL DRIVING PULSE DURATION SHALL BE 0.1H ±0.005H.
7- THE TIME RELATIONSHIP AND WAVEFORM OF THE BLANKING AND SYNC SIGNALS SHALL BE SUCH THAT THEIR ADDITION WILL RESULT IN A STANDARD RETMA SIGNAL. THE TIME RELATIONSHIP MUST BE ADJUSTABLE IN ORDER TO SATISFY THIS RELATIONSHIP FOR THE CONDITION WHERE THE BLANKING SIGNAL IS DELAYED WITH RESPECT TO THE SYNC SIGNAL OVER THE RANGE FROM 0.0H TO 0.05H.
8- THE STANDARD RETMA VALUES OF FREQUENCY AND RATE OF CHANGE OF FREQUENCY FOR THE HORIZONTAL COMPONENTS OF THE SYNC SIGNAL AT THE OUTPUT OF THE PICTURE LINE AMPLIFIER SHALL ALSO APPLY TO THE HORIZONTAL COMPONENTS OF THE OUTPUT SIGNALS FROM THE RECOMMENDED SYNC GENERATOR.
9- ALL RISE AND DECAY TIMES TO BE MEASURED BETWEEN 0.1 AND 0.9 AMPLITUDE REFERENCE LINES.
10- THE TIME OF OCCURRENCE OF THE LEADING EDGE OF ANY HORIZONTAL PULSE "n" OF ANY GROUP OF TWENTY HORIZONTAL PULSES APPEARING ON ANY OF THE OUTPUT SIGNALS FROM A STANDARD SYNC GENERATOR SHALL NOT DIFFER FROM "nH" BY MORE THAN 0.0008H WHERE H IS THE AVERAGE INTERVAL BETWEEN THE LEADING EDGES OF THE PULSES AS DETERMINED BY AN AVERAGING PROCESS CARRIED OUT OVER A PERIOD OF NOT LESS THAN 20 NOR MORE THAN 100 LINES.
11- EQUALIZING PULSE AREA SHALL BE BETWEEN 0.45 AND 0.6 OF THE AREA OF A HORIZONTAL SYNC PULSE.
12- THE OVERSHOOT ON ANY OF THE PULSES MUST NOT EXCEED 5%.
13- THE OUTPUT LEVEL OF THE BLANKING SIGNAL & THE SYNC SIGNAL MUST NOT VARY MORE THAN ±3% UNDER THE FOLLOWING CONDITIONS:-
   A. THE A.C. VOLTAGE SUPPLYING THE SYNC GENERATOR MUST BE IN THE RANGE BETWEEN 110V AND 120V AND MUST NOT VARY MORE THAN ±5V DURING TEST.
   B. A PERIOD OF 5 HOURS CONTINUOUS OPERATION SHALL BE CONSIDERED ADEQUATE FOR THIS MEASUREMENT, AFTER SUITABLE WARM-UP.
   C. THE ROOM AMBIENT MUST BE IN THE RANGE BETWEEN 20 DEG. & 40 DEG. C. AND MUST NOT CHANGE MORE THAN 10 DEG. C. DURING THIS TEST.
14- ADJUSTMENT MUST BE POSSIBLE BETWEEN MINIMUM AND MAXIMUM LIMITS SO THAT ASPECT RATIO CAN BE SET TO THE NORMAL VALUE.

Figure 11-129. Standardized RETMA Waveform, Showing Video Pulses sion systems, even the most basic closed-circuit system, produce synchronizing information which conforms to the basic requirements, although the degree of conformance varies widely. Diagrams (A) and (B) of figure 11-129 show in considerable detail how the synchronizing signal waveform is added to the picture information and the blanking signals to form a complete composite picture signal ready to be transmitted for commercial use. Note that the duration of the horizontal synchronizing pulses is considerably shorter than that of the blanking pulses. Vertical synchronizing pulses are rectangular in shape, but are of much greater duration than the horizontal pulses, thus providing the necessary means for frequency discrimination. Their general waveshape, amplitude, and tolerance limits are shown in diagrams (A) and (B) of figure 11-129. Synchronization presents one of the most difficult of problems. It is the one in which the largest number of failures occurs which result in loss of proper interlacing. Discrepancies in either timing or amplitude of the vertical scanning of alternate fields cause displacement, in space, of the interlaced fields. The result is nonuniform spacing of the scanning lines, which reduces the vertical resolution and makes the line structure of the picture visible at normal viewing distance. The effect is usually called "pairing." To obviate the pairing problem and to maintain continuous horizontal synchronizing information throughout the vertical synchronization and blanking interval, another series of pulses called "equalizing" pulses, have been added before and after the vertical synchronizing pulses. The time between the last horizontal sync pulse and the first equalizing pulse changes from a full horizontal line interval to a half horizontal line interval every other field. This is caused by the ratio between 15, 750 cps and 60 cps, which produces the necessary difference between fields to provide interlaced scanning. Since the horizontal oscillator is adjusted to the frequency of the horizontal sync pulses, it is triggered only by every other equalizing pulse or serration of the vertical sync pulse.

11-1038. OTHER STANDARDS.

11-1039. Although commercial broadcasting and many closed-circuit installations adhere closely to the previously described RETMA standards, there are some noncommercial as well as closed-circuit installations that use synchronizing signal specifications which are considerably less rigerous. These are worthy of note and will be outlined briefly here. The television systems to be discussed fall into four general categories, as listed below.

a. System I: Random interlace, no special synchronizing pulses.

b. System II: Odd line interlace, no special synchronizing pulses.

c. System III: Odd line interlace, modified synchronizing pulses.

d. System IV: Special slow-speed scan technique.

11-1040. SYSTEM I. This is the simplest television method, since it provides no special synchronizing pulses and no fixed relationship between the horizontal and vertical scanning raster. "Lock-in" or synchronization information at the receiver or monitor is obtained from the horizontal and vertical blanking pulses contained in the video signal as shown in part A of figure 11-130. Usually, at the camera control location, sufficient blanking signal or "setup" is added to the video signal to provide an adequately long and steep transition at both vertical and horizontal frequencies to provide picture lock-in. Some receivers and monitors may have difficulty, however, in synchronizing on this meager information. In the pres-

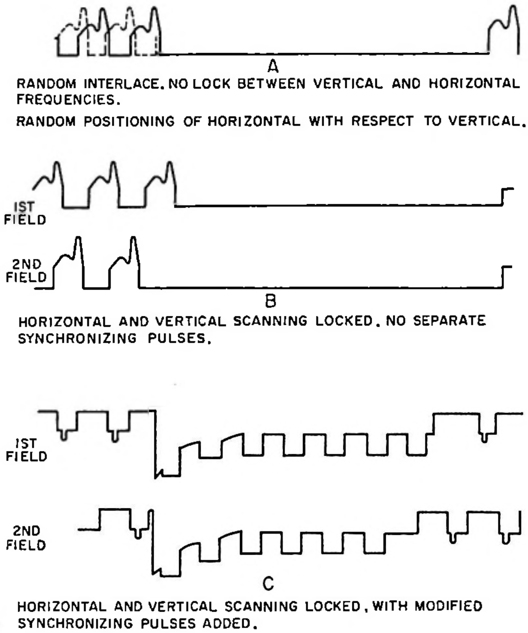


Figure 11-130. Non-Standardized Television Waveforms

ence of electrical noise, which is a possible condition when the camera and viewing device are separated by a great distance, the lock-in of such a signal becomes extremely difficult. Note, in A of figure 11-130, that there is no horizontal synchronizing signal during vertical blanking time. The horizontal frequency circuit in the monitor, even though it has some tendency to keep on the correct frequency, will be essentially free-running during this period. This may not be detrimental during blanking or retrace time; however, when horizontal synchronization information returns, the receiver horizontal circuits may have trouble synchronizing to the new information. Probably the most undesirable characteristic of television System I is lack of sufficient resolution caused by the lack of interlace. Good interlace is not possible because absolute frequency relationship between the horizontal and vertical frequencies is lacking. The nominal vertical frequency is usually 60 cps (locked to 60-cycle power line), while the horizontal frequency (usually established by a free-running oscillator) is nominally 15.75 kilocycles. Thus, there is no direct relationship between the two frequencies (horizontal frequency should be an odd multiple of one-half the field rate), as is required for satisfactory interlace. The advantages of such a system, of course, are reduced cost and greater simplicity of circuits. However, marginal resolution capabilities, incompatibility, marginal stability, and general reduction in performance of this system seriously limit its application and use.

11-1041. SYSTEM II. This system has a distinct advantage over System I, since a definite relationship exists between the horizontal scan frequency and the vertical field rate. This system, whose waveform is shown in B of figure 11-130, can effectively use the 2:1 odd-line interlace technique. It thereby provides a considerable improvement in the resolution capabilities of the system. In theory, the vertical resolution should be double that of the previous system; in practice, however, the improvement in resolution is somewhat less, since it is difficult to obtain perfect interlace (no pairing). System II, like System I, does not provide a special synchronizing signal and is subject to the same synchronizing or lock-in limitations as were discussed previously. These limitations become a particularly important problem in the more elaborate installations, where a series of cameras and monitors separated by wide distances might be used. For this reason, installations using these systems are usually limited to smaller, less complex applications, where stability and reliability may not be an important factor.

11-1042. SYSTEM III. This method provides further advantages over the previous two systems but has a considerable number of limitations when compared with the RETMA system. In this system, a form of special synchronizing signal has been added to the video waveform, as illustrated in part C of figure 11-130. Note that the synchronizing signal has been added to the tip of each horizontal blanking pulse. The synchronizing pulses also continue through the vertical blanking interval, providing synchronizing information for the monitor horizontal frequency-locking circuits at all times. These circuits, therefore, are no longer free-running during the vertical blanking interval. Addition of the special synchronizing pulse information greatly improves the lock-in ability of the composite video signal under more adverse conditions of noise and spurious signals.

11-1043. SYSTEM IV. A television system finding increasing usage is one utilizing the slow-speed scan technique. This represents a radical departure from nominal scanning standards. This technique permits a scene which contains a limited amount of action or movement and a great deal of redundancy to be picked up and transmitted successfully from one location to another. It affords fair resolution and fidelity, over relatively economical narrow-band transmission facilities. For example, a "slow-speed" camera located in a bank, a message center, or even a newspaper office, can scan printed information and transmit it to a distant location over ordinary telephone line facilities. Some methods are able to transmit pictures having more action, such as a person talking, with reasonable clarity; such methods, however, require somewhat more bandwidth. Most slow-scan methods presently used are of an experimental nature; the results, however, are extremely encouraging and indicate considerable potential in this field. Most slow-speed scan systems use a much slower scanning rate than present telecasting standards, with a correspondingly narrower bandwidth. Broadcasting systems transmit a picture every 1/30 second, with a 4-megacycle bandwidth. Slow-scan systems transmit a picture in from 1/10 second to 2 seconds, with a video bandwidth ranging from approximately 250 kilocycles to as low as 500 cycles. Slow-speed scan systems are practical where time is available for transmission. For example, the information contained in a 5-minute commercial television program requires several hours of time to be transmitted with comparable detail by the average slow-speed scanning technique. The advantages of the slow-speed scanning system are greatly simplified equipment and relatively inexpensive transmission facilities as compared to the complex relay systems required for broadcast television. The disadvantages are that the scene content is limited to relatively immobile objects, resolution is marginal, and the system is incompatible with standard television systems. Rather complex scan conversion equipment is required to make the two systems compatible. Except for certain special applications, slow-speed scan systems are inferior in performance and cannot be used successfully where a higher degree of resolution and detail is required.

11-1044. CLOSED-CIRCUIT TELEVISION.

11-1045. A closed-circuit television system consists basically of the equipment necessary for generation, distribution, and display of video signals, and includes a parallel audio or sound system. Adjuncts to the video equipment provide for camera mounting, lighting, signal stabilizing, testing, etc. The audio system provides for program sound, intercommunication, and talkback.

11-1046. VIDEO SIGNAL GENERATING EQUIPMENT.

11-1047. GENERAL. In any closed-circuit

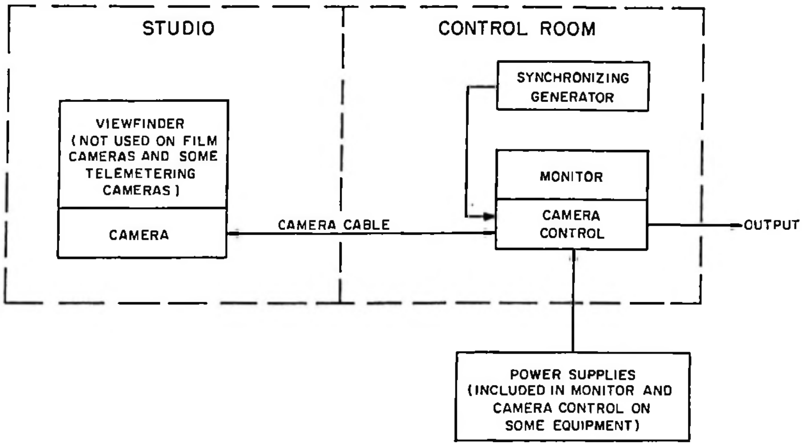


Figure 11-131. Camera Chain, Simplified Block Diagram

television system, the following basic equipments are required: (1) synchronizing generator, (2) pickup cameras, (3) camera control equipment, (4) monitors, and (5) projectors and multiplexers (for film pickup). The pickup camera is the basic unit since it converts the optical image to an electrical one. The picture source may be a live program or telemeter scene, or film; one of several film projectors may be selected by the multiplexer. The synchronizing generator furnishes synchronizing information which is used throughout the system. Monitors are used to check the quality of the video waveform.

11-1048. SYNCHRONIZING GENERATOR. The synchronizing generator furnishes the horizontal and vertical drive pulses and the synchronizing information required by the scanning circuits of the cameras and receivers in the complete video signal generating system. Hence, the synchronizing generator is often referred to as the heart of the television system. The synchronizing generator provides four output signals which are used throughout the television system. Two of these signals, the synchronizing and blanking pulses, appear in the video signal; the other two, the horizontal and vertical drive pulses, are used locally for synchronization of the studio equipment. The synchronizing generator is usually located at or near the point of program origin.

11-1049. PICKUP CAMERAS. The pickup camera and those control and auxiliary units which must be duplicated for each camera constitute a camera chain. Figure 11-131 is a block diagram which illustrates the arrangement of the major units in a camera chain. Pickup cameras can best be described by classifying them according to use as studio-type cameras, film cameras, and telemetering or remote monitoring cameras.

11-1050. Studio-type cameras, as the name implies, are those which are specifically designed for permanent installation in a studio or similar location. They are of high-quality design and are used where high quality picture production is desired. Although the studio-type camera is sometimes used for high quality telemetering or remote-monitoring applications (where subject matter is presented independently of television and television is used as a means of repeating the visual information at remote points), their primary use is for programming (where a program is produced for and by television). Studio-type cameras usually have features that distinguish them from film cameras and telemetering cameras. These include such items as intercommunication facilities, tally lights, a lens turret for multiple lens mounting, a supporting cradle and operator's handle for pan and tilt control, iris control, smooth and positive mechanical focus control, and a high-quality viewfinder.

11-1051. Film cameras are used to obtain a high-quality, high-resolution television picture from film, either motion picture film or slides. Film cameras have a wide contrast range and are capable of operation for long periods of time without adjustment. Controls for the film camera are generally similar to those for "live" studio cameras. There are three types of film pickup cameras. They differ chiefly in the type of pickup tube or system used, for example, iconoscope pickup tube, the flying spot scanner, and the vidicon pickup tube.

11-1052. The iconoscope pickup tube obtains the image from a film projector which focuses a picture on a photosensitive mosaic plate. The varying illumination of the optical image causes electron emission and the buildup of stored charges on the mosaic (cesium-silver "islands" on the face of the mica plate). When a scanning beam from the electron gun moves across the mosaic, the stored charges are released from the islands, in sequence, and are capacity-coupled to a metal signal plate. The charges then appear across the output load impedance as voltages which are proportional to the intensity of light in the original optical image. While the iconoscope tube is still widely used in film cameras, it has several inherent limitations; edge flare, black-level instability, poor shading, and a relatively poor signal-to-noise ratio all tend to limit its performance.

11-1053. The flying spot scanner employs a modification of the mechanical disc-scanning used in the early days of television. This method uses a phosphor-coated crt and a photoelectric cell. In the flying spot kinescope, magnetic deflection coils cause an electron beam to scan the phosphor-coated face of the tube. This generates a moving spot of light which represents the scanning raster. The spot of light is then passed by an extensive lens assembly through the film, and through another lens assembly onto the photoelectric cell. The variations in light intensity, introduced by the varying transmission coefficients over the area of the film transparency, are converted into proportionally varying electrical signals by the photoelectric cell; the result is the required electrical image. The flying spot scanner produces adequate pictures, but the signal-to-noise ratio is relatively low because of the limited light from the kinescope and the inherent limitations of the photoelectric cell. The flying spot principle is relatively simple, and camera design is correspondingly inexpensive. However, its use for motion picture film pickup requires a complex projector design. (This is because of problems resulting from the lack of storage in the photocell which makes it impractical to use an intermittent-type projector.)

11-1054. The vidicon pickup tube is ideal for use in a film camera. It has an excellent signal-to-noise ratio, stable black level, and improved shading. It is small and simple, and it performs well with the light supplied by the average film projector. A photosensitive element takes on a pattern of positive charges as a result of electron emission caused by an optical image. The beam from the electron gun strikes this element and discharges it to zero potential. Unlike the orthicon, however, the video information in the vidicon is taken directly at the photosensitive element. Excess electrons, instead of returning to the rear of the tube, are collected on the screen in front of the photosensitive element and returned to the cathode indirectly. The video information, then, is the voltage developed through the load resistor when current flows as a result of the target element returning to zero. The photosensitive element is designed to have low lateral conductance so that adjacent picture areas do not discharge into each other during one frame time.

11-1055. CAMERA CONTROL UNITS AND MONITORS.

11-1056. Cameras are designed to have minimum size and weight consistent with their application. Those functions not housed in the camera itself are incorporated in a separate housing called the camera control unit. Typical functions performed by the camera control unit are: (1) to provide control of the electrical performance of the camera pickup tube (electrical focus, beam current, etc), (2) to synthesize the picture signal which is to be delivered to the video switching equipment (adjust signal level, provide dc restoration, add blanking signals, provide gamma correction and aperture compensation, add shading, if necessary, and provide signals for monitoring), and (3) to provide proper termination for the cables connecting the camera with the remainder of the system.

11-1057. In most portable or field applications, the camera control unit includes monitoring facilities. Although picture and waveform monitoring facilities are closely associated with the camera control unit, in most studio camera systems the monitors are usually housed separately. In a large installation, separately housed monitoring facilities can be used at other points in the television system where required. The monitors used in conjunction with the camera control unit are precision units used to display a number of details of the video signal. The monitor is of sufficiently high quality to show accurately the composition of the picture signal.

11-1058. PROJECTORS AND MULTIPLEXERS.

11-1059. Closed-circuit television frequently makes use of motion picture film as a source of program material. Television film projectors, therefore, are an integral part of a television system. These projectors produce a sharp, stable optical image from the film and project it with sufficient light intensity on the photosensitive surface of a camera pickup tube. In the reproduction of film, it is usually more economical to operate a number of film and slide projectors with only one pickup camera. This is made possible by the use of multiplexers. A multiplexer provides the optical link necessary to select one of several film, slide, or opaque projector outputs for projection into a film pickup camera. Movable mirrors, driven by motors or relays, change the optical paths to permit one of several optical picture sources to impinge on the camera pickup tube. High quality mirrors and reflecting surfaces are used to keep the light efficiency high and to prevent multiple or distorted images or other deterioration of the image. The mirrors, motors, and control panel, contained in one housing, constitute the multiplexer.

11-1060. VIDEO SIGNAL DISTRIBUTION EQUIPMENT.

11-1061. GENERAL. After a television signal has been produced and processed by the video signal generating equipment, it has to be distributed or transmitted to its ultimate destination. This is the video signal display equipment. This equipment may be a monitor, a receiver, a kinescope recorder, or a combination of these display and reproducing devices. The equipment which distributes and transmits the video signal to the display or recording devices is called video signal distribution equipment. This equipment includes distribution amplifiers, equalizers, rf generators, rf distribution equipment, microwave relay and switching equipment.

11-1062. DISTRIBUTION AMPLIFIERS. In most television systems, low-impedance coaxial cables (usually 75 ohms) are used to distribute timing and drive pulses and video signals. Driving the television signals through these cables is the function of the distribution amplifier. Distribution amplifiers are usually divided into two groups on the basis of the specific function they perform: (1) video distribution amplifiers, specifically designed to handle the picture signals; and (2) pulse distribution amplifiers, used to distribute the various synchronizing and drive pulses from the synchronizing generator. The most important requirement of a video distribution amplifier is to distribute video signals passing through it without introducing appreciable degradation or distortion. Video signals are attenuated in passing through coaxial cables so most distribution amplifiers provide at least 6 db of amplification to compensate for this loss. In portable systems, which are usually less elaborate and have less equipment, distribution amplifiers are not used. Instead, most portable or field systems have output driving amplifiers, similar to those in the video distribution amplifiers, included as an integral part of the camera control or switching equipment. This arrangement obviates the need for additional units.

11-1063. Pulse distribution amplifiers are much like video distribution amplifiers except that they are designed to distribute the higher amplitude drive and synchronizing pulses from the synchronizing generator. Pulse distribution amplifiers include a pulse regenerator circuit. This circuit produces a pulse that is dependent upon the width and timing of the output pulse but is otherwise independent of it. Thus, the regenerator circuit forms a pulse with ideal characteristics regardless of poor risetime, overshoot, tilt, or other defects that may be present in the incoming signal.

11-1064. EQUALIZERS. Coaxial cables, used extensively for interconnecting television systems, have relatively poor amplitude-versus-frequency response. Long or improperly terminated cables, therefore, may cause serious high-frequency attenuation and accompanying phase distortion. Such distortion will cause reduced resolution, ringing, smearing, or streaking in the picture. For this reason, devices called video equalizers are inserted in the coaxial lines.

11-1065. RF GENERATORS. Distribution of a television signal (0—6 mc) within the immediate area is usually accomplished by coaxial cable, using the standard video bandwidth of 6 megacycles. When the picture signal, however, is to be distrubuted to remote locations, the use of rf generators and rf distribution systems is required. Rf generators are simply low-power television transmitters, usually tuned to one of the commercial vhf television channels. Instead of feeding an antenna, however, the generator feeds a coaxial line which then carries the vhf signal to one or more regular television receivers.

11-1066. Rf generators are of two general types. The first type uses a single television channel and can accept video signals only. It simply converts video frequencies to radio frequencies to operate a single receiver; audio facilities are not provided. The second type accepts both video signals and the associated audio signals. This type usually can operate on any of the standard vhf television channels.

11-1067. MICROWAVE EQUIPMENT. Distribution of television signals by coaxial cable over distances of 5 miles or more is difficult and costly. Cables cause considerable attenuation and have other undesirable characteristics when used for long-distance transmission. For this reason, a more appropriate system, the television microwave relay, is usually used for transmission over longer distances. A television microwave relay system is essentially a complete television system. It contains a transmitter, a transmitting antenna, a receiving antenna, and a receiver. Such a system is designed to operate at frequencies in the microwave region extending from 1000 megacycles to 15,000 megacycles. Here, with a minimum of transmitter power and highly directional antennas, television signals can be relayed over line-of-sight distances (30 miles or more) without serious attenuation or picture degradation.

11-1068. SWITCHING EQUIPMENT. Switching equipment is necessary in all but the simplest, one-camera television installation. It connects and selects the video output from one of several television signal sources to one outgoing path or distribution system. Switching equipment, in addition to making the signal transfer, also performs several other functions. Synchronization is inserted, and the transfer is smoothly made with fade-ins and fade-outs, lap dissolves, or such special effects as diagonal wipes. Instead of switching, two signals may be superimposed (as when titles are flashed over a picture), or mixed in split-screen montages. Furthermore, there are provisions for previewing the video and monitoring the output line.

11-1069. VIDEO DISPLAY EQUIPMENT.

11-1070. GENERAL. Video display equipment reproduces the original picture or scene viewed by the pickup camera, and converts the electrical signals into an optical image. Although the technique of reproduction may vary slightly in different types of equipment, all video display equipment employs some form of cathode-ray tube. Video display equipment is classified primarily according to picture size, as follows: (1) small screen which employs a direct-view kinescope, with picture sizes of 8, 10, 12, 14, 17, 21 or 24 inches (diagonal measurement); (2) medium screen which uses projection kinescope and optics, with picture sizes of 3 x 4 feet or 6 x 8 feet; (3) large screen which uses projection kinescope and optics, with picture sizes up to 15 x 18 feet. Video display equipment is also divided into two groups, monitors and receivers, according to the type of signal with which they operate (video frequency or radio frequency).

11-1071. TELEVISION RECEIVERS. Television receivers reproduce a picture image from an rf signal. Many closed-circuit television systems are extensive enough to require distribution of the video signal over considerable distances by rf line distribution or microwave transmission. Depending on the quality of reproduction desired, receivers may be used in such a system for the final video display. The receiver units employ rf amplifiers, i-f amplifiers, and detectors to recover the video information from the modulated rf signal. Receivers include the necessary video circuits and kinescope to reproduce the signal directly and can also supply the demodulated video to a series of monitors for display. Receivers, which are used primarily for picture viewing, are all of the small-screen type.

11-1072. TELEVISION MONITORS. Television monitors reproduce a picture image from a video-frequency signal. The signal may be taken directly from the video line or, in a more extensive closed-circuit system, the monitor may receive a demodulated signal from a television receiver. By adding the necessary rf and i-f stages the monitor may also function as a receiver. Small-screen television monitors are high-quality units which give optimum reproduction of video signals. They contain such features as complete dc restoration, wide-band video amplifiers, external synchronization addition (for noncomposite video signals), and magnetic focus and deflection. In addition to filling the studio needs of quality and continuity checks and projection room cueing, the small-screen monitor is used extensively for picture viewing. The size of the picture tube is determined by the particular application. A small classroom or administrative office might use a 24-inch tube; a studio control room might employ a 17-inch tube, but an 8-inch tube is adequate for continuity checks or cueing work.

11-1073. TEST EQUIPMENT.

11-1074. GENERAL. Generally, test procedures and test equipment required to troubleshoot closed-circuit television components are specified in technical manuals provided by manufacturers with their equipment. Test equipment is normally required to: (1) observe the character of the video signal, (2) evaluate the performance of the television equipment, (3) maintain and repair the equipment, and (4) align and adjust the equipment. In addition to the usual tube checkers, voltmeters, ohmmeters, and milliammeters used with electronic equipment, the following test equipments are particularly applicable to television use: (1) cathode-ray oscilloscope, (2) video sweep generator, (3) multiburst generator, and (4) grating generator.

11-1075. OSCILLOSCOPE. A cathode-ray oscilloscope suitable for television application is specifically designed for maintenance, alignment, and adjustment of that equipment. It presents accurately all of the video waveforms that are employed in a television broadcast installation. Appropriate horizontal sweep circuits permit the observation of any portion of the television picture waveform, from complete television picture frames to small portions of individual scanning lines. Calibrated attenuator and sensitivity controls permit variations of sensitivity of the oscilloscope. Wideband vertical amplifiers are provided so that any one of the horizontal lines of a picture may be observed in minute detail. The amplifiers are adjusted for optimum transient response rather than maximum bandwidth. Accurate time markers are used in the measurement of pulse rise time, pulse width, and other pertinent measurements with respect to time. The oscilloscope, therefore, provides an accurate means of measuring both amplitude and time of television signals, as well as a detailed representation of the picture signal.

11-1076. VIDEO SWEEP GENERATOR. The video sweep generator is a convenient device for measuring the amplitude of a given signal as a function of frequency. The output of a fixed rf oscillator, operating at approximately 70 megacycles, is heterodyned against a sweep frequency oscillator, operating between 69 and 80 megacycles. The 0 to 10-megacycle (approximate) video beat frequency is then applied to the circuit or unit to be tested, and the output from the test circuit, after detection, is observed on an oscilloscope. A marker notch is inserted for frequency calibration; this is accomplished by an additional oscillator stage in the sweep generator. Most sweep generators are swept at a 60-cps rate which provides a convenient, instantaneous visual display of the entire sweep characteristic. A more accurate frequency can be obtained with a sweep generator that employs a calibrated cw oscillator. This type provides either sweep or cw operation, over a tunable range of 100 kilocycles to 10 megacycles.

11-1077. MULTIBURST GENERATOR. The multiburst generator, like the video sweep generator, measures the amplitude-versus-frequency response of a television system. Although it is somewhat limited for test setup and maintenance of individual components or circuits, the multiburst generator is very effective for systems operational checks. The generator applies test signals of standard television frequencies directly into the television system to provide a spot check of the system response; no circuits need be disabled. The test signal consists of bursts of various frequencies superimposed on a pedestal, with standard blanking and synchronizing pulses added. All bursts are adjusted for equal amplitude before the test signal is applied to the input of the television system. Changes in burst amplitude, as observed on a waveform monitor (at the system output, or at any convenient check point), indicate frequency response deviations for the particular frequency or frequencies. As indicated above, this is simply a spot check, and does not necessarily isolate the cause of a malfunction.

11-1078. GRATING GENERATOR. A uniform distribution of picture detail on the kinescope of a receiver or monitor requires a uniform velocity of scanning by the deflection circuits. If the scanning characteristic deviates from the standard, picture detail is compressed over part of the area and expanded over the balance of the area of the kinescope. The grating generator provides a convenient means for checking and adjusting the linearity of television deflection circuits. It generates a timing signal synchronized by standard television synchronizing pulses and injects this signal into the video circuit under test. The pattern produced on the kinescope has the appearance of a grating (grid). The frequency which generates the vertical bars is a multiple of the horizontal scanning frequency and that which generates the horizontal bars is a multiple of the vertical scanning frequency. Equal spacing between vertical bars over the width of the raster indicates the linearity of scanning by the horizontal deflection circuit. Equal spacing between horizontal bars over the height of the raster indicates the linearity of scanning by the vertical deflection circuit.

11-1079. TESTING.

11-1080. GENERAL. A closed-circuit television system can be composed of a single camera and a receiver for the simplest installation, or of many cameras, monitors, receivers, distribution equipment, and special purpose equipment, for a more complex system. Hence, testing and trouble shooting a system necessarily require individual procedures, as outlined in the pertinent technical manuals. However, by the very nature of a television system, it is relatively easy to isolate trouble to the defective equipment component. Most systems have more than one receiver (monitors included); by observing the chain, if one receiver is defective, the others will operate satisfactorily. If all receivers fail to provide a picture (or audio) but the camera monitor operates satisfactorily, the distribution equipment is at fault. If no receiver or monitor provides a picture, the camera or associated equipment is at fault. In addition, many circuit defects may be isolated by analysis of an improper picture. Tests, like picture resolution, sensitivity, etc, should be made periodically to uncover faults of an atrophying nature.

11-1081. RESOLUTION TEST. The resolution (resolving power) of a television system is defined as the measure of its ability to delineate picture detail and is expressed in terms of a number of lines resolved on a test chart. Figure 11-132 shows a chart used for standardizing resolution measurements. Many stations employ this chart or variations of it for testing resolution and picture distortion, but this chart is not generally broadcast because it is much more detailed than the typical test pattern. A

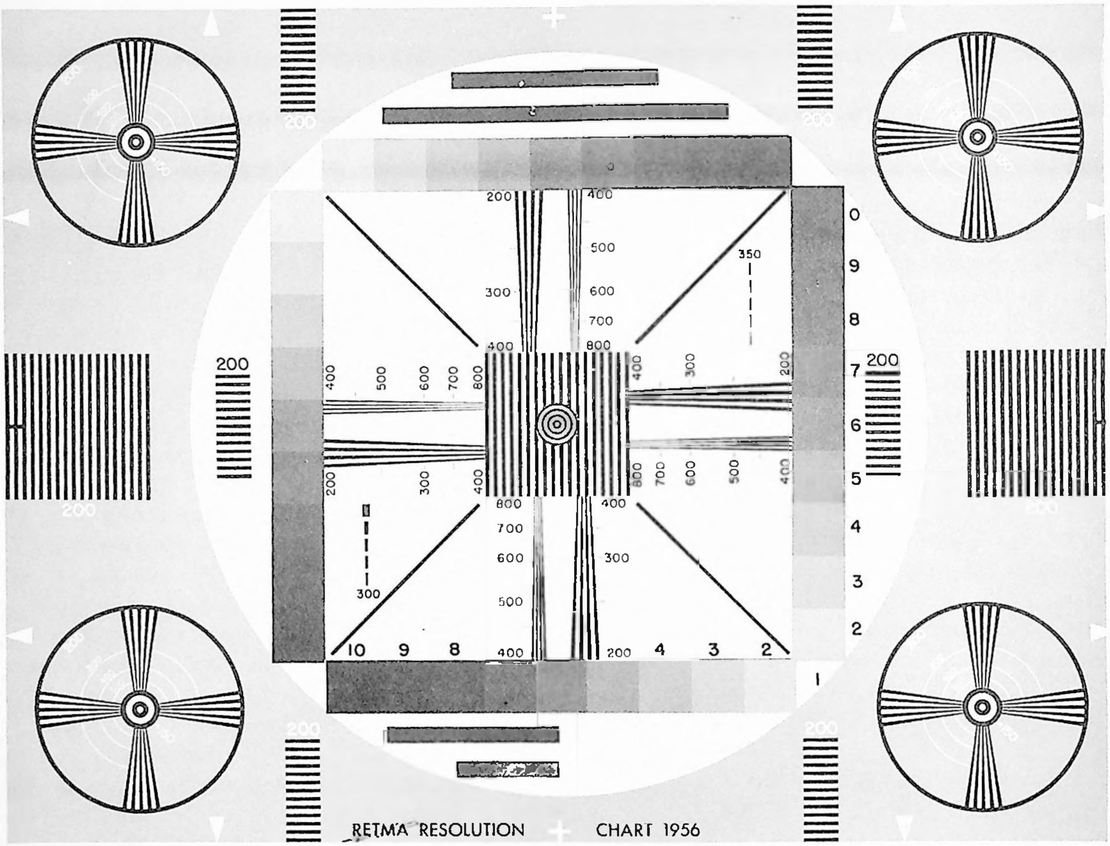


Figure 11-132. Resolution Chart (RETMA)

suitable chart is very helpful in making performance checks. Most test patterns have a series of vertical and horizontal wedges consisting of alternate black and white lines, circles, and vertical and horizontal bars. In addition, the resolution chart shown in figure 11-132 has a gray scale consisting of four wide bars positioned in the form of a square. Each of the four gray-scale bars is composed of 10 logarithmic steps from maximum white brightness to approximately one-tenth of this value (black). Aspect ratio is checked by noting that the four gray-scale bars, in the case of figure 11-132, or circle, grid or other geometric figure on a test pattern, form a perfect square, circle, grid or figure. The wedges located in corner circles permit linearity and resolution to be measured in the four corners of the picture. Larger wedges located near the center are calibrated in lines of resolution; some charts are also calibrated in frequency response. Resolution calibration extends up to 800 lines and can be used with equipment having a bandwidth up to 10 mc. The vertical and horizontal parallel bars are spaced for 200-line resolution and are used to check horizontal and vertical linearity. Resolution is measured by the number of alternate black and white lines of equal width which can be seen separately. The point in the wedges where the lines start to blend to-

Table 11-8. Resolution of Representative Monochrome Closed-Circuit Television Systems

TYPE OF TELEVISION SYSTEM	HORIZONTAL RESOLUTION (Center)	HORIZONTAL RESOLUTION (Corners)	VERTICAL RESOLUTION (Center)	VERTICAL RESOLUTION (Corners)	RELATIVE HORIZONTAL DETAIL RESPONSE AT 300 LINES (Center)
High Performance System	600 lines	500 lines	350 lines	350 lines	1.0 or higher
Medium Performance System	450 lines	300 lines	250 lines	200 lines	0.5
Low Performance System	300 lines	200 lines	175 lines	150 lines	0.0

gether and are just barely perceptible as separate and distinct lines indicates the limit of resolution. By referring to the calibration adjacent to the wedges or bars, an expression of resolution may be read directly in terms of lines. The vertical resolution is determined from the horizontal wedges by the number of separate and distinct lines that can be traced, one above the other. Vertical wedges are provided for determining the horizontal resolution. These are identical to the horizontal wedges and are calibrated in the same manner. They show the number of lines resolvable horizontally in a length equal to the height of the picture or 3/4 of the picture width (4/3 aspect ratio). The horizontal resolution is generally stated as the number of lines it is possible to resolve in a horizontal distance equal to the height of the picture. In the case of the 4/3 aspect ratio, the horizontal resolution is determined from the wedge reading times 4/3. Actual values of resolution and horizontal detail response achieved with systems of varying degrees of performance are summarized in table 11-8.

11-1082. GRAY-SCALE REPRODUCTION. The gray scale on the test chart (10 degrees between black and white) can be used for checking the contrast of the television system. When the television system is operating satisfactorily and is properly adjusted, the individual gradations of the gray scale should be clearly defined. Faithful gray-scale reproduction requires that the gray scale (tonal values indicated in gradations of white through gray to black) of the original scene be reproduced in the same proportion or ratio as in the display. Since equipment factors enter into the gray-scale reproducing characteristics of a television system and inasmuch as the linearity of a system is subject to control, certain closed-circuit television systems require non-linear gray-scale reproduction, especially those systems used to collect data from meters, etc. Therefore, there is no absolute standard of equipment performance because gray scale or contrast is largely a subjective matter similar to picture definition. Depending upon the type of picture information which must be reproduced, the amount of contrast distortion or compression which can be tolerated is rather wide. However, a gray-scale standard for a particular system may be made as an equipment performance standard.

11-1083. GEOMETRIC DISTORTION. Geometric distortion is any defect which causes the perspective or geometry of the reproduced picture to deviate from that of the original scene. It is desirable in most applications to keep distortion below 2 percent of picture height. Greater distortion is immediately noticeable, but it may be tolerated depending on the particular application. Geometric distortion may be produced by scanning circuit malfunction and by improperly designed locus and deflection yokes.

11-1084. Scanning circuits in either the pickup camera or the video display equipment may introduce distortion into the picture on the display kinescope. Nonlinear operation of the sawtooth sweep generator, either horizontally or vertically, results in an apparent compression in the displayed picture. It is generally on the right side of the displayed figures, because of the slowing down of the sweep. If distortion originates in the deflection circuits, it normally takes one of the following four forms:

a. S-curving. An example of S-curve distortion is shown in (A) of figure 11-133. It is the result of a nonuniform axial field in the pickup tube, which imparts nonuniform rotation to the electrons in the scanning pattern. S-curving may also result from improper adjustment of the pickup tube potentials.

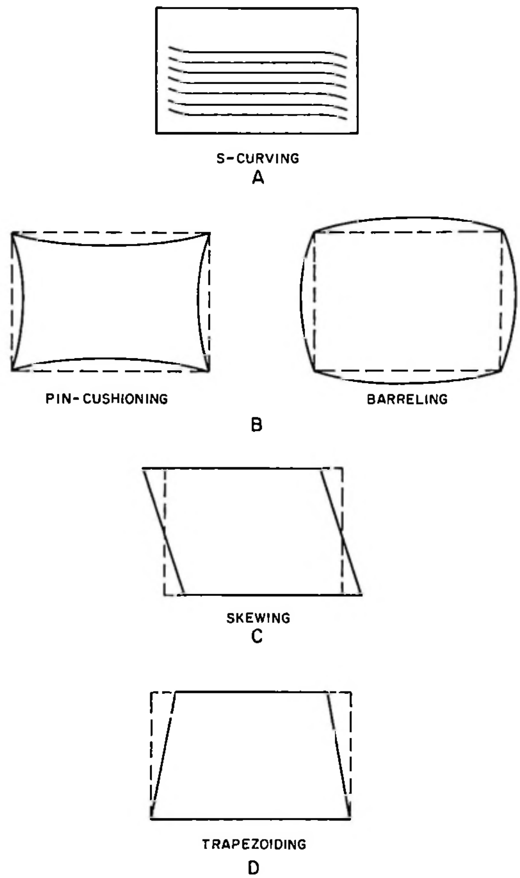


Figure 11-133. Typical Types of Geometric Distortion

b. Pin-cushioning (or barreling). Pincushioning distortion, shown in (B) of figure 11-133 results from improper distribution of the windings in a kinescope deflection yoke. Consequently, it is a fairly common failing in low-cost units employing low-quality components.

c. Skewing. An example of skewing is shown in (C) of figure 11-133. It is produced when the horizontal and vertical deflection yokes (for either the pickup tube or the display kinescope) are not perpendicular to each other.

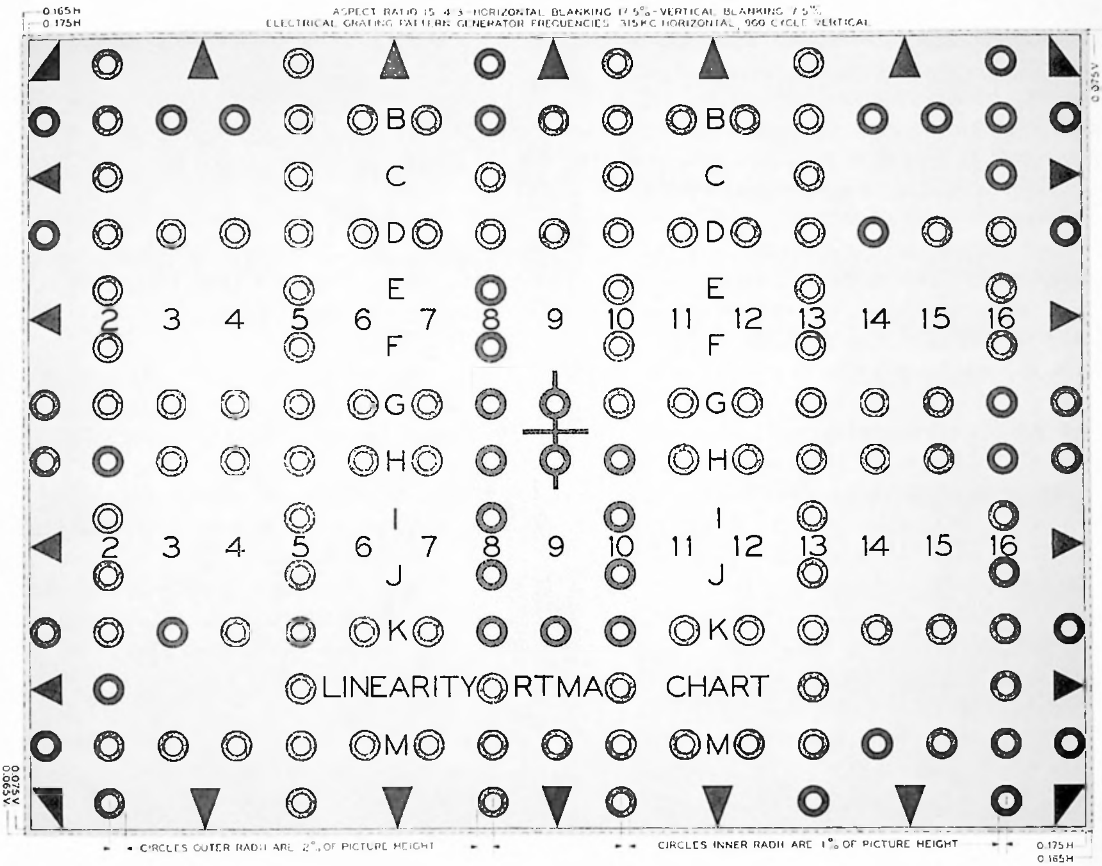


Figure 11-134. Linearity Chart (RETMA)

d. Trapezoiding. Trapezoiding, shown in (D) of figure 11-133, is the result of one set of deflection yokes not being equidistant about the axis of the other (the axes of the horizontal and of the vertical deflection coils should effectively bisect each other). This distortion may be introduced in the pickup camera or in the display equipment.

11-1085. Geometric distortion may be measured directly on a picture monitor. Two patterns are superimposed and compared. One pattern, provided by a pattern or grating generator, is a grating of horizontal and vertical bars. The other pattern is a televised linearity chart (figure 11-134) consisting of a series of dots, circles and intersections. With zero geometric distortion, the dots, circles, or intersections of the linearity chart can be made to coincide with the intersections on the grating pattern (within 2 percent tolerance) by adjustment of camera linearity controls. Any distortion present on the superimposed display is measured in distance units (for picture elements) and time and distance figures (for scanning velocities).

11-1086. SIGNAL-TO-NOISE RATIO. Noise in television is defined as any spurious or random signal which has a degrading effect on the quality of a television presentation and is usually grouped into four separate categories: (1) tube noise, (2) impulse noise, (3) low-frequency noise and hum, and (4) rf interference. Since it is normally random in nature, noise is usually measured in terms of rms voltage and is usually expressed in db as a ratio of the noise to some fixed value or level. The value usually used is the level of the video signal with which the noise is associated; hence it is classed as a signal-to-noise ratio. The signal-to-noise ratio of the output from the pickup camera is usually determined by the noise produced in the first stage of the preamplifier chassis. General information regarding signal-to-noise ratio is given under paragraph heading 11-34.

11-1087. GENERAL SERVICING AND TESTING.

11-1088. GENERAL. Servicing of a television system includes all equipment components, that is, the camera, receiver, antennas, and distribution equipment. However, servicing is simplified to some extent because most troubles can be localized visually or aurally. For example, the television signal can be broken down into three components: (1) raster, (2) picture, and (3) sound. Analysis of defects apparent in any or all of the components can isolate the trouble to a particular equipment circuit or group of equipment circuits. When the trouble has been localized to a section of the television equipment, the defective component can normally be located by signal tracing and voltage measurements or resistance checks.

11-1089. Tubes are a common source of trouble, and should be checked first preferably by substituting a tube known to be good for the suspected one. Tubes that are noisy or microphonic can often by located by gently tapping them and noting any indication of noise in the sound or picture. Power supply rectifiers and power output tubes, since they are driven at high current ratings, are a frequent source of trouble.

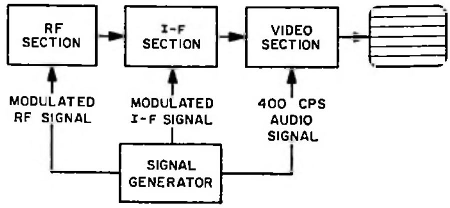


Figure 11-135. Test Setup Using Signal Generator for Signal Injection in Picture Circuits

11-1090. SIGNAL TRACING. Signal tracing is a valuable technique in the servicing of television equipment. Three methods of signal tracing suitable for expeditious location of trouble are: (1) signal injection methods, (2) oscilloscope method, and (3) voltmeter method.

11-1091. The signal injection method is suitable for a system having a normal raster presentation but no picture. For this method a test signal from a standard signal generator having 400 or 1000 cycle modulation is injected into the video circuits. When passed, the test signal appears as a series of horizontal bars on the face of the kinescope. In the case of a receiver, the signal is introduced into each amplifier stage in turn, from the kinescope toward the antenna input until a point is reached where the signal does not pass, thereby isolating the defective stage (see figure 11-135). In the intermediate stages the modulated rf test signal is injected at the video intermediate-amplifier frequency; in the rf section, the test signal is injected at the rf picture carrier frequency.

11-1092. The oscilloscope method is more versatile than the signal injection method described above, but can be used in conjunction with that method when deemed desirable. Normally, however, signal tracing with an oscilloscope is accomplished with a picture signal input. The path of the signal is traced by observing the composite video signal on the oscilloscope screen, from the video detector output to the grid of the kinescope. The oscilloscope can also be used to trace the sync signals from the composite video signal input through the sync circuits. Observing the sync wave shapes can be helpful in cases of hum, poor interlacing, and horizontal afc defects. Also, the deflection voltages can be observed to check linearity and amplitude for horizontal and vertical scanning. When viewing the composite video, the vertical input of the oscilloscope is connected to the circuit being checked and the internal sweep of the oscilloscope is used. Setting the internal sweep frequency to 30 cps provides a two-cycle waveform of vertical scanning fields; setting it to 7875 cps provides the waveform for two cycles of the horizontal scanning lines.

11-1093. A voltmeter can be employed to test the operation of an oscillator or an amplifier using grid-leak bias. During normal operation, a negative voltage is produced at the grid; absence of this voltage is indicative of a failure in this circuit. Therefore, operation of the local oscillator, vertical deflection oscillator, or horizontal deflection oscillator can be checked by measuring the negative grid-leak bias with an electronic voltmeter or other highly sensitive voltmeter. For amplifiers employing grid-leak bias, a negative voltage at the grid is indicative of ac excitation. The horizontal output, sound i-f, and mixer amplifiers are typical of stages normally employing this bias. The voltmeter can also be used to measure the rectified signal output of the video detector and fm sound detector. Absence of this voltage indicates that there is no i-f signal input.

11-1094. COLOR TELEVISION RECEIVERS.

11-1095. GENERAL.

11-1096. PARTS. Whether a television set is a black-and-white set or a color set, it may be considered as consisting of two parts, the receiver assembly and the cathode-ray-tube assembly. You my have serviced many black-and-white television sets and know that it makes no difference which assembly is aligned first. Color television sets are different in this respect because the color tube beams should be converged before the receiver assembly is aligned.

11-1097. CATHODE-RAY-TUBE ASSEMBLY. The cathode-ray-tube assembly for a black-and-white television set is not a very well standardized assembly. Manufacturers have located control coils and control magnets on the neck of the cathode-ray tube in combinations, locations, and positions which suit their own purposes for a particular model. Of all the assemblies mounted on the tube neck for electro-magnetic deflection sets, the deflection yoke, a focusing magnet or coil assembly, and an ion trap are standard. Color television cathode-ray-tube assemblies are more stereotyped than their black-and-white counterparts. Color television picture tubes using three electron guns to produce a color image are quite well standardized with respect to the control mechanisms mounted along the neck of the tube. Refer to figure 11-136.

11-1098. Adjustments performed on a color television cathode-ray-tube assembly, should be performed in a definite sequence. You should follow the recommended procedure for many reasons. For example, faulty convergence adjustments can cause a slight shift in color purity. Unless you inspect the color image closely, your conclusion may lead you to perform a color purity adjustment. This adjustment would only compound the color image faults by introducing other errors that would not allow you to change circuit conditions proportionately with control settings. It is, therefore, advisable to perform the adjustments in sequence, once you have decided

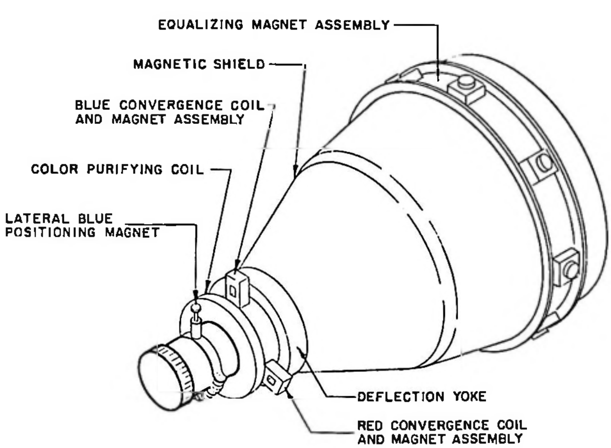


Figure 11-136. Location of Deflection Assemblies

that adjustment is really necessary. The adjustments required to obtain a good color picture are aimed toward:

a. Convergence of the three electron beams at all points within the picture viewing area.

b. Color purity.

c. White balance.

11-1099. RECEIVER ASSEMBLY. In many respects, there is very little difference between the black-and-white and the color television receiver until the last video i-f amplifier is reached. In theory and in general practice, the shapes of the response curves are relatively the same for both types of receiver up to this point. Manufacturers, in the interest of economy and with small sacrifice in picture quality, have limited the video bandwidth to approximately 3 to 3.5 megacycles per second for their black-and-white television receivers. This limitation has the advantage of reducing both the interference problems and the complement of trap circuits required in the receiver. Color television, on the other hand, requires a minimum video bandwidth of 4.2 megacycles to adequately display a color picture; therefore, color equipments require a full complement of accurately tuned trap circuits. Even though closed circuit color television equipments may not use adjacent channel transmissions and the geographical area may not permit adjacent channel broadcasting, the traps of the color receiver must still be accurately tuned. If the upper adjacent channel (video modulations) trap (39.75 mc) is misaligned, interference in the form of a vertical black bar will be observed shifting horizontally back and forth across the viewing area. This bar is the horizontal blanking signal (and other video signals) slipping through the misaligned trap from the upper adjacent channel, and it cannot be locked in synchronization with the present channel in use. If the lower adjacent channel (sound modulations) trap (47.25 mc) is misaligned, interference in the form of diagonally shifting lines can be observed. These lines result when the sound carrier of the next lower television channel slips through the misaligned trap. There are two general rules of thumb that can be applied to the alignment of trap circuits. One rule is that when two traps operate at the same frequency the output trap should be tuned first and the input trap tuned last. The second rule applies when you discover that a trap circuit functions at two different core settings; the best performance will generally occur at the position of minimum core insertion in the coil.

11-1100. A comparison of color and black and white television sets is shown in figure 11-137. The unshaded blocks represent those sections of a television set normally found in both the black-and-white and color television receiver. The shaded blocks represent only those sections found in a color television receiver. From this diagram you can see that the chrominance circuits and the convergence circuits form the two major differences between black-and-white and color television equipments.

11-1101. When a color receiver is being used to receive a black-and-white picture, you can inspect the white areas in the picture and actually see the individual red, green, and blue triads that make up the screen phosphors. It should be apparent that the individual electron beams in the picture tube must always be focused on the correct portion of the triad, even when a black-and-white scene is being televised. Considering the size of the electron guns as compared with the size of the triads, the three beams must be individually deflected toward one another so that they strike the triad properly; this type of deflection is called convergence. Once the three beams are converged, they can be moved as a unit from triad to triad under the influence of the horizontal deflection voltages. As long as the angle of deflection is restricted to a relatively small angle, the three beams will move from triad to triad with very little error. As the angle of deflection increases, the amount of energy needed to keep the beams properly converged must also be increased. The convergence circuits must develop two kinds of control on the three electron beams; one is a steady, continuous control, called static convergence, and the other is a varying control, called dynamic convergence. Static convergence affects the central area of the cathode-ray-tube screen. It is accomplished by applying a dc voltage to the convergence coils to create a basically constant magnetic deflection field in the immediate vicinity of each electron gun. Dynamic convergence affects only the outer regions of the cathode-ray-tube screen. It is accomplished by injecting a parabolic voltage into the convergence coils to create a varying magnetic field in addition to the steady field already present in the coils. The parabolic voltage waveform used to drive the dynamic convergence circuits is derived from the vertical and horizontal deflection circuits.

11-1102. Observation of figure 11-137 shows that the chrominance video and color synchronization signals are taken from the video detector. These circuits, unlike the convergence circuits, need only be activated when a color scene is being televised. When a black-and-white scene is being televised, there is no color burst contained in the video signal. Hence, the keying pulse arriving from the horizontal deflection circuit is insufficient to activate the color synchronization and chrominance circuits. When a color scene is being televised, the color burst pulse is present as part of the video information; it arrives at the input to the chrominance and color synchronization circuits at the same time the horizontal keying pulse arrives. These two pulses, acting together in the same time interval, permit the television set to detect that a color scene is being televised instead of a regular black-and-white scene; refer to figure 11-138.

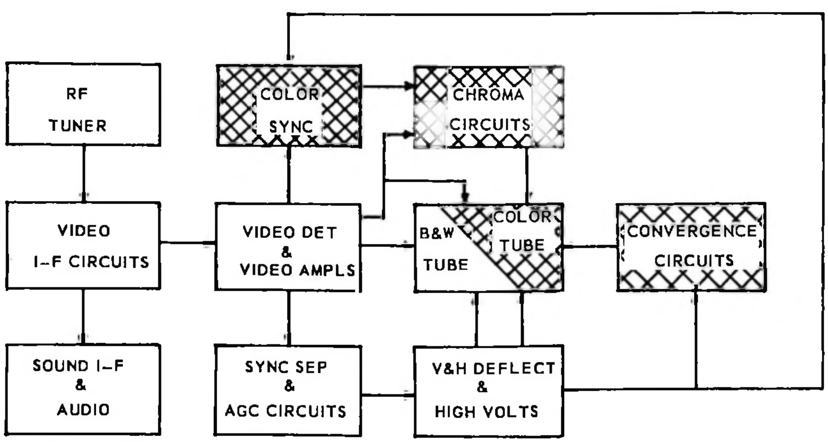


Figure 11-137. Comparison of Color and Black and White Television Sets, Block Diagram

11-1103. The regular synchronization pulses that you are familiar with in black-and-white sets are used to control the stability of the horizontal and vertical deflection circuits. Loss of these synchronization pulses results in the loss of the televised picture. Color synchronization plays no part in the stabilization of the deflection circuits. Loss of color synchronization results in shimmering, shifting, and misplacement of colors, but the colors will always be present. Do not mistake the intermittent or erratic shifting between color and black-and-white as a loss in color synchronization; the color-killer circuit is the only circuit that can produce this particular defect.

11-1104. The first stage in the color synchronization circuits is the burst amplifier. This stage requires the presence of a horizontal keyer pulse and the burst pulse before it can conduct. The output of the burst amplifier consists of a single burst pulse. The keyer pulse, obtained from a winding on the horizontal flyback transformer, occurs during the blanking interval as does the burst pulse, which is situated on the back porch of the blanking pulse; both of these signals occur, therefore, during the interval when the picture is not on the cathode-ray tube.

11-1105. The output of the burst amplifier is applied as one input signal to the phase detector circuit. The other input signal to the phase detector stage is taken from the output of the 3.58-mc reference oscillator. Since these two input signals have the same frequency, any difference between the signals will be a phase difference. The output of the phase detector stage, therefore, is a dc voltage which is proportional to the difference between the local oscillator frequency and the transmitter burst frequency.

11-1106. The dc output voltage of the phase detector is applied to the grid of a reactance tube. The reactance tube is connected to provide circuit action similar to that of a capacitor placed across the tank circuit of the oscillator. As the output of the phase detector varies, the effective capacitance in the oscillator tank circuit also varies; the phase of the reference oscillator will be changed so that it is in synchronism with the burst oscillator located at the station transmitting the television signal.

11-1107. The chrominance circuits must now be considered. The horizontal keyer pulse and the burst pulse, acting together, raise the bias of the color killer out of the plate current cutoff region. The time constant of this circuit is adjusted so that the color-killer tube remains conductive during the entire time the horizontal scan is active.

11-1108. The output of the color-killer circuit adjusts the operating bias of the bandpass amplifier into the linear operating region. The complete video signal, which is continually applied to the bandpass amplifier from the video detector, is now permitted to drive the amplifier. The output of the bandpass amplifier is tuned only for chrominance information; hence, all of the luminance (Y) signal is removed, including the synchronizing pulses; refer to figure 11-142.

11-1109. The X and Z demodulators, shown in figure 11-138, require two input signals each. The X demodulator receives, as one input, chrominance signals from the bandpass amplifier. The other input is the signal taken directly from the 3.58-mc reference (subcarrier) oscillator. These two signals are added together in the X demodulator, and the resultant signal is passed into the color matrices of the receiver. The Z demodulator receives the chrominance information from the same circuit point in the bandpass amplifier as does the previously discussed X demodulator. The second input is the signal from the 3.58-mc reference (subcarrier) oscillator after it has been passed through a quadrature circuit. These two signals are added together in the Z demodulator and then passed into the color matrices of the receiver. The quadrature circuit can be a vacuum tube circuit or a transformer-coupled circuit; the only requirement of the quadrature circuit is to shift the phase of the oscillator voltage applied to the Z demodulator by 90 degrees with respect to the oscillator output voltage applied to the X demodulator.

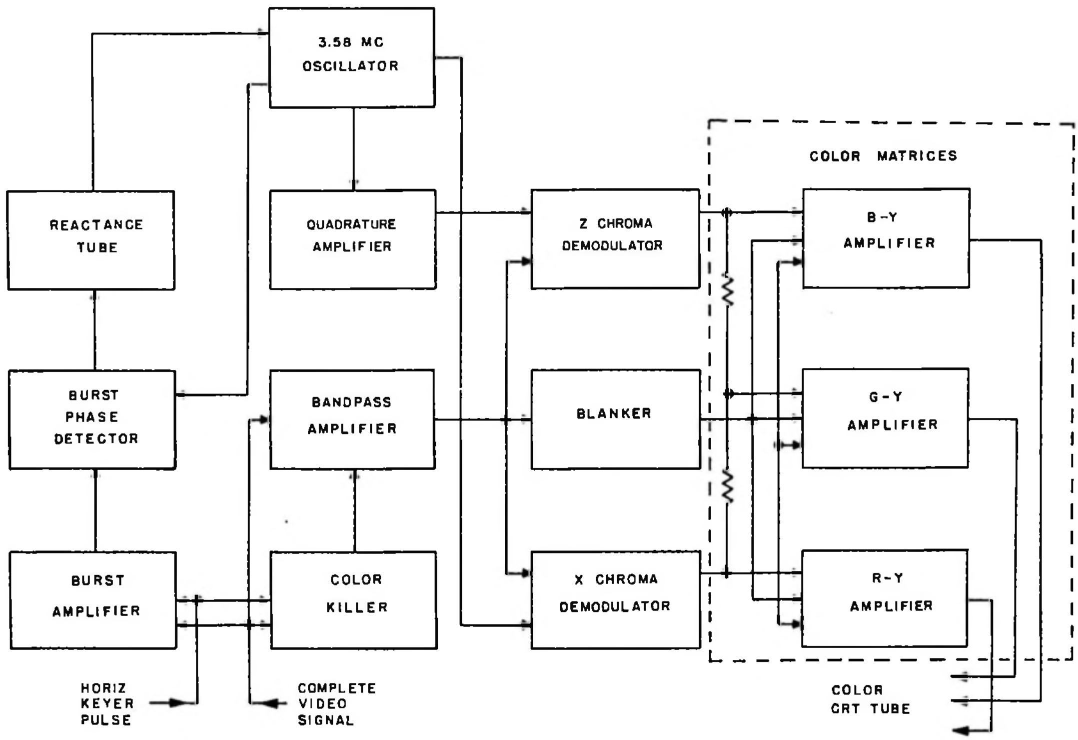


Figure 11-138. Television Chrominance and Color Synchronization, Block Diagram

11-1110. Note that both the X and Z chroma demodulators feed their signals into opposite ends of the matrices. The output voltages of both demodulators distribute themselves across the two resistors shown in figure 11-138, for application to an R-Y, a G-Y, and a B-Y amplifier. The relative distributions of these voltages are such that the voltages interact with each other to ultimately produce signals that are injected into the color cathode-ray tube. The R-Y amplifier output signal is sent to the red gun in the cathode-ray tube, the G-Y signal is sent to the green gun, and the B-Y signal is sent to the blue gun. Do not make the mistake of calling the output of the R-Y amplifier (for instance) red; <u>it is only the chrominance portion of the red signal</u>, and needs the luminance signal to form the complete red signal which excites the screen phosphors. The final mixing of chrominance and luminance takes place at the cathode-ray tube in the latest-design receivers. Older types of color receivers mix the luminance and chrominance signals in the matrix section of the receiver.

11-1111. TROUBLE SHOOTING.

11-1112. TUBE FAILURES. Television sets, whether color or black-and-white, are just as prone to tube failures as any other electronic equipments. Since color television is relatively new, many technicians who are thoroughly familiar with black-and-white television maintenance techniques are reluctant to align or repair a color set. Table 11-9 shows the chrominance stages in a color television set to be checked first when one of the X-marked indications occurs.

11-1113. SIGNAL TRACING. The signal-tracing method of locating defective color television circuits is essentially the same as for other electronic equipment. Connect a color bar generator to the antenna terminals of the receiver. Connect a wide-band oscillator through a low-capacitance probe to the point in the burst amplifier circuit where the complete video signal is normally an input. You can now observe the video signal display; refer to figure 11-139. Inspect the amplitude of the displayed receiver synchronization pulses, and then inspect the peak-to-peak amplitude of the burst pulses; these two signals should have approximately the same amplitude, each with respect to the other, whether from a color bar generator as in figure 11-139 or from a regular televised scene. If the burst pulses are attenuated, poor rf or i-f alignment or trap circuit misalignment is indicated. By transferring the oscilloscope probe to the horizontal keyer pulse input, you can observe keyer pulses similar to those shown in figure 11-140. Calibrate the oscilloscope and measure the pulse amplitude; then check against the value specified in the service manuals. If the amplitude of the keyer pulse is low, the burst amplifier output signal will also be of low amplitude. If the amplitude of this pulse

Table 11-9. Logical Suspect Circuit To Be Tested When X-Marked Indication Occurs

Circuit	Colors				Hum Bars	Loss of Sync	Tinted Rasters
	None	Partial	Weak	Erratic			
Color Killer	X		X	X	X		
Bandpass Amplifier	X		X		X		
Chroma Demodulators		X			X		X
Chroma Amplifiers		X			X		X
Burst Amplifier						X	
Phase Detector	X					X	
Reactance Tube	X					X	
Reference Oscillator	X	X				X	X
Tuner, Video, and Video HF	X	X			X		

is incorrect, it is advisable that you check the keyer winding in the horizontal output circuit. The output signal from the burst amplifier, as shown in figure 11-141, should consist of a single pulse. Disregard any tendency of the pulse to be unsymmetrical; measure the peak-to-peak amplitude of the pulse and compare this value with published data. If the output pulse amplitude is reduced more than 20 percent, the burst amplifier should be inspected for faulty operation.

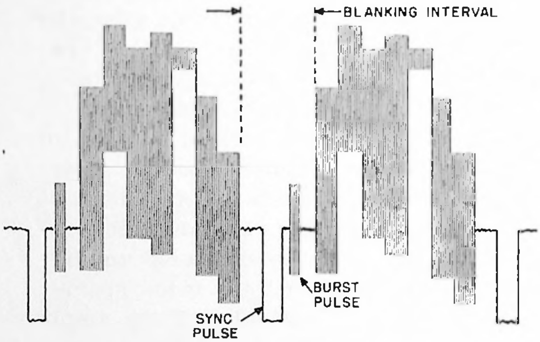


Figure 11-139. Oscilloscope Display of Complete Color Bar Video Signals

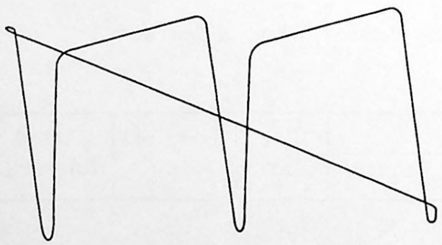


Figure 11-140. Oscilloscope Display of Horizontal Keying Pulses to Color Synchronization and Chrominance Circuits

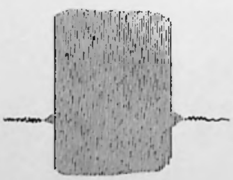


Figure 11-141. Oscilloscope Display of Burst Amplifier Output Pulse

11-1114. Transfer the low-capacitance probe to the input of the phase detector so that you can view the output signal of the burst amplifier, to be certain the pulse has not been altered. Move the oscilloscope probe to view the 3.58-mc oscillator signal as applied to the phase detector tube. Recalibrate the oscilloscope and measure the peak-to-peak amplitude of the oscillator signal to determine whether this signal has the correct amplitude when compared with the available published data.

11-1115. Using the oscilloscope and low-capacitance probe, observe the 3.58-mc signal directly from the output terminals of the oscillator proper; measure the peak-to-peak amplitude of this signal and compare the measured value with the published data. In cases where the 3.58-mc oscillator is not crystal-controlled, form a single-turn loop around the oscillator tube (following tube shield removal, if any), using one test lead ending in an alligator clip. Following the completion of the loop, the clip is attached to the line to hold the loop tight and prevent slippage. Connect the other end of the lead to a frequency meter to measure the oscillator operating frequency. Oscillators that use the LC tank circuits for this kind of application use a tuning slug for frequency adjustment. It may prove advisable to remove the reactance tube prior to tuning the 3.58-mc oscillator frequency.

11-1116. If the color reproduction is weak, check the dc control voltage at the grid of the bandpass amplifier; it is possible that the output from the color-killer stage biases the bandpass amplifier too near plate current cutoff. Using the wideband oscilloscope and the low-capacitance probe, observe the video output waveform at the input of the X and Z chroma demodulators and the blanker circuit. Notice, in figure 11-142, that the video signal to these

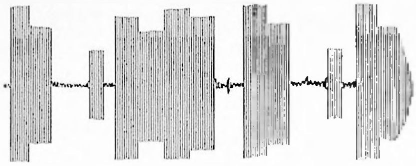


Figure 11-142. Oscilloscope Display of Chrominance Video Signals at Output of Bandpass Amplifier

demodulators contains only chrominance signals. The luminance, blanking and synchronizing pulses have been removed from the complete video signal by means of the bandpass amplifier. Recalibrate the oscilloscope, measure the peak-to-peak amplitude of the chrominance signal waveform, and consult the published data; a signal of low amplitude at this point in the circuit will produce weak colors in the picture.

11-1117. Where the colors are weak in the color bar reproduction, recalibrate the oscilloscope and measure the 3.58-mc injection voltage at the input of both the X and Z chroma demodulators; then compare this value with published data. It is quite possible that the injection voltage to one demodulator is satisfactory, while it is faulty or absent at the other demodulator.

11-1118. The output signals from the X and Z demodulators can be checked directly as long as the color bar generator you are using provides some access to the X/90° and the Z/90° signals, such as separate output terminals or selection by means of a function switch, etc. Otherwise, only an indirect check of the output from these demodulators is possible.

11-1119. Operate the color bar generator controls to obtain a Z/90° output signal. Connect the wideband oscilloscope through the low-capacitance probe to the input of the R-Y amplifier. Only a horizontal sweep should be observed in the oscilloscope display; there should be no vertical deflection. If any vertical deflection is obtained, vary the color phasing control (tint control) to eliminate the vertical deflection. In case the vertical deflection cannot be eliminated by this process, inspect the burst phase detector and the reactance tube circuits for faulty operation or abnormal phase shifts. If no vertical deflection is obtained or vertical deflection has been eliminated, transfer the oscilloscope probe to the input of the G-Y amplifier. You should obtain vertical deflection at this point in the circuit; refer to figure 11-143 for a typical output waveform. Then transfer the probe to the input of the B-Y amplifier. At this point in the circuit you should receive a greatly increased amount of vertical deflection as compared with that obtained at the input of the G-Y amplifier, because you are receiving the full output signal from the Z demodulator. Now operate the color bar generator controls to obtain the X/90° output signal. Any vertical deflection you had previously should disappear at the input to the B-Y amplifier without moving the setting of the color phasing control (tint control). If it does not disappear, faulty operation of both demodulator stages is indicated. If the oscilloscope display is satisfactory at this point, transfer the probe to the G-Y amplifier input, where you will again observe some vertical deflection. Do not expect the same signal amplitude at the input of the G-Y amplifier when using the Z input signal that you obtained at the same point when using the X signal. Transferring the oscilloscope probe to the input of the R-Y amplifier will result in an oscilloscope display of the full output signal amplitude of the X demodulator.

11-1120. If the X and Z color bars are not separately available from the color bar generator, the demodulators must be checked indirectly. Connect the oscilloscope through the low-capacitance probe to the output of the R-Y amplifier. Set the color bar generator to inject a B-Y signal into the receiver, and then adjust the color phase control for zero output from the R-Y amplifier. Transfer the scope probe to the output of the B-Y amplifier; you should observe a large output signal similar to that shown in figure 11-143. Adjust the color bar generator to give you an R-Y signal into the receiver; the output of the B-Y amplifier should be reduced to zero without adjustment of the color phasing control. Now transfer the oscilloscope probe back to the output of the R-Y amplifier; you should obtain full output from this point in the circuit.

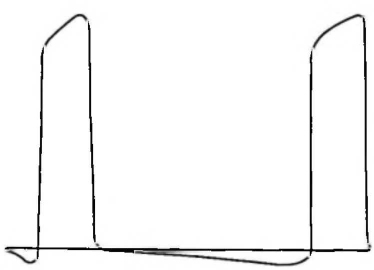


Figure 11-143. Oscilloscope Display of Output Signal from X or Z Chroma Demodulator

11-1121. If only a rainbow generator is available, the peak-to-peak output voltages should be measured by using the oscilloscope at the output of the R-Y, G-Y, and B-Y amplifiers. This data is often contained in receiver service instructions. Phasing and signal amplitudes are so interrelated that if one is correct, the other is equally correct.

11-1122. It should be fairly obvious to you that if a color television set is receiving a black-and-white telecast and the resulting picture is a single color, such as yellow, one or more of the matrix amplifiers is not functioning. In the case of yellow images, the B-Y amplifier is inoperative; in the case of cyan (blue-green), the R-Y amplifier is inoperative; in the case of magenta (red-blue), the G-Y amplifier is inoperative. Removal of the B-Y amplifier when the black-and-white scene is yellow should produce little or no change in the observed picture.

11-1123. SIGNAL INJECTION. Trouble shooting by the signal injection method provides another practical means of quickly locating faulty circuits. The use of the color bar generator as the signal injecting device is very useful, but there are some points that must be discussed before this method is used. Notice that the generator can be set to provide rf, i-f, or video signals. The video signal function will permit you to choose between a positive-going or a negative-going output video signal. When you are making signal injection tests in the video amplifier, you must decide whether a plus video signal or a minus video signal is required at the circuit point in question. No damage will be done to the set if you make a mistake, but the results can be misleading. If the detector provides a positive-going video output signal and you inject a negative-going video test signal, you will lose the horizontal synchronization pulses and blanking pulses, together with the burst pulse situated on the back porch of the blanking pulse. Quite often, the 3.58-mc subcarrier signal must be injected directly into a receiver oscillator. The subcarrier may require some additional amplification before it is sufficient to drive the chrominance demodulators or the subcarrier oscillator. Circuit loading must be carefully watched when injecting the 3.58-mc signal into the oscillator circuit; hence, the best injection point is at the oscillator grid. A rainbow generator is not suitable for signal-injection testing because the color subcarrier frequency of the rainbow generator is 3.56 mc instead of the 3.58 mc that the receiver requires.

11-1124. If there is a poor television display with the color bar generator connected to the receiver antenna terminals, it would be advisable to check the response of the rf and i-f stages. Disconnect the color bar generator from the antenna terminals and reconnect the generator at the video detector output terminals. Set the function switch of the generator to provide you with a complete video signal. Do not forget to check for the plus or minus video requirements at this circuit point; remember that only one of these switch positions will give you the proper receiver operation. If a normal color bar pattern is observed on the television screen, the rf tuner and video i-f amplifiers (including the input to the video detector) should be investigated for misalignment or a defective stage.

11-1125. If the receiver exhibits loss in color synchronization, connect the color bar generator to the output terminals of the video detector. Adjust the generator to provide you with a complete plus or minus video signal so that the receiver will stabilize its deflection circuits. Connect an additional lead from the 3.58-mc generator terminal to the 3.58-mc receiver oscillator. You may have to provide additional amplification by means of an external video amplifier, in order to obtain the signal amplitude necessary to drive the receiver subcarrier oscillator. This will cause the receiver oscillator to be locked in synchronism with the color bar generator. The phasing of the oscillator is now independent of the reactance tube and phase detector. Remember, however, that the phase of the generator oscillator is arbitrary with respect to the demodulators. The tuning range of the color phasing (tint) control may be adequate to lock the color bar pattern at the start of the horizontal sweep under these circumstances. Turn the color phasing (tint) control through its tuning range to determine whether a normal color bar display can be observed on the cathode-ray tube. If the normal pattern can be displayed at some point in the tuning range of the tint control, investigate the phase detector and reactance tube circuits for faulty operation. If the sequence of color bars is correct but appears to start at some point in the television display other than the beginning of the horizontal sweep, switch the generator function control to the chroma position. The portions of the previous picture tube display which contained white will now become black, because the luminance (Y) signal is eliminated from the generator output. Turn the color phasing control through its tuning range once again. If the sequence of colors is correct at some setting of the tint control, but still start at some location in the picture tube display other than the beginning of the horizontal sweep, the phase detector and/or reactance tube is defective. However, the chroma demodulators and matrices are functioning correctly.

11-1126. If the correct sequence of colors cannot be obtained at some setting of the color phasing control, the chroma demodulators and matrices are defective. Operate the color bar generator to its R-Y function, and adjust the color phasing control to obtain the R-Y color bar. If you cannot obtain the R-Y color bar, investigate the X demodulator and matrix for faulty operation. Perform the same test for the B-Y color channel which involves the Z demodulator.

11-1127. The bandpass amplifier can be checked by connecting the output of the color bar generator to the chrominance video input terminal of the X and Z demodulators. Set the generator function switch to the chroma position. If a normal color pattern is now obtained, the bandpass amplifier is defective. If the normal color display cannot be obtained, then continue inspecting the chroma demodulators and matrix sections.

11-1128. MISALIGNMENT SYMPTOMS.

11-1129. GENERAL. Observation of the television picture itself will indicate the condition of the circuits producing the picture. For example, focusing, size, and linearity defects in the picture can be detected quite easily. Furthermore, when you identify one of these types of defects, you have effectively pinpointed a circuit (and its associated controls) that needs repair or readjustment. The defect in the picture can be corrected quite easily by a simple readjustment of the indicated control if the indicated circuitry does not require maintenance.

11-1130. When the television picture is faulty, you must decide whether the trouble is in the video circuits or the deflection circuits. In a black-and-white equipment, operate the channel selector to an unused channel. If you have a complete raster, the difficulty is in the video portion of the receiver; if the raster is defective, the deflection circuits should be investigated for defects.

11-1131. In a color equipment the procedure is somewhat different. Turning the set to an unused channel will result in a dark screen if there is no incoming signal or input noise; "snow" or "confetti" will be observed as large colored dots if there is considerable noise on the unused channel. After the unused channel is selected, turn the contrast control completely counterclockwise and set the brightness control at approximately its mid position. The raster should now be visible, but it will be gray rather than white. If the raster is gray and well formed, not only are the deflection circuits operating correctly, but the convergence and color purity circuits are functioning correctly as well.

11-1132. If there is any reason to suspect the convergence circuits of faulty operation, operate the television set to a channel that will give you a black-and-white display. Faulty convergence will be immediately apparent because the black-and-white images will be bordered with one or more color fringes. However, do not make the mistake of trying to adjust any of the convergence controls. Inspect the black-and-white scene very carefully; look for loss in the width or height in the television raster, and observe the display very closely for nonlinearity. The convergence circuits depend very heavily upon the formation of a linear sawtooth out of the vertical and horizontal deflection circuits. The parabolic waveshape required by the convergence circuits simply cannot be obtained when these deflection voltages are nonlinear or of insufficient width or height. When you are satisfied beyond any doubt that the deflection voltages are not at fault, then proceed to the convergence adjustments.

11-1133. Alignment defects can also be observed in the television picture, in the sound channel, or both. However, incorrect television receiver tuning will invalidate conclusions concerning receiver alignment. As the fine tuning control is rotated, the fine details in the picture should be observed when the sound is the best. If the sound and picture are not their best at the same setting of the fine tuning control, the receiver should be aligned. This condition occurs more commonly in receivers that have separate sound and video i-f stages, but it may also occur in intercarrier receivers. Generally, you should disregard the mottled or spotted condition known as "snow" when tuning for a television signal. Quite often in weak signal areas, the tuning positions for the best sound and minimum snow are separated by a small amount. On the other hand, do not forget that these weak signals can be caused by poor alignment.

11-1134. FINE DETAILS ABSENT. Poor alignment of the television tuned circuits will produce a faulty response curve. When the response curve is not broad enough to include the required high frequencies, the picture quality is reduced. The reduced quality ranges from loss of fine details to a smeared picture in which all of the picture details are stretched toward the right-hand side of the viewing area.

11-1135. TRAVELING GHOSTS. Secondary images in the television picture are known as "ghosts." Some ghosts cannot be eliminated, because of the locality of the television receiver or the orientation of the television antenna. Ghosts that move horizontally as the receiver fine tuning control is varied, however, indicate a poorly aligned receiver.

11-1136. RINGING. A faulty video amplifier or a poorly aligned television receiver can also produce another type of ghost. This type of ghost is the same on all channels of the receiver, and can be identified as follows:

a. Ghosts appear at points in the picture where there is a sharp change from black to white, usually along vertical edges.

b. A ghost is most pronounced near the change in contrast point, gradually diminishing in brightness as the scanning line moves to the right.

c. The spacing between the ghost images is the same.

11-1137. FINE MESH. When a beat note is encountered at points on either side of the correct tuning position of the fine tuning control, alignment of receiver tuned circuits is required, especially the 4.5-megacycle trap circuits. This is usually a 4.5-megacycle beat, and can be observed on the television picture at all points as a fine mesh pattern. A fine mesh pattern can also appear in color television sets as a result of a 920-kilocycle beat note generated between the chroma subcarrier and the intercarrier sound signals.

11-1138. HIGH CONTRAST. Poor alignment in television receivers is also indicated by extremely high contrast in the observed picture. The picture is often grainy and the range of gray picture areas is very limited. This type of picture can be the result of a response curve that has been highly peaked in the high frequency portion of the curve.

11-1139. SOUND BARS. Alignment of the receiver tuned circuits is required when sound bars become apparent in the picture. These sound bars appear as dark horizontal bars across the television picture that come and go in time with the frequency of the accompanying sound. The sound bars in color television sets are not necessarily any particular color; they appear in the same color as that being transmitted at the time, but they tend to be darker than the adjacent area.

11-1140. DISTORTED SOUND. Weak or distorted sound can indicate receiver misalignment. This symptom is detected by the same method you would employ in any other fm detector and af amplifier stage testing.

11-1141. IMPROPER COLOR RENDITION. Improper color rendition can be due to poor alignment of the receiver tuned circuits. A chroma amplifier (or amplifiers) may require alignment if one or more colors is missing from the televised picture. Misalignment of rf and i-f amplifiers can produce weak colors over the entire picture viewing area. Misalignment of the chroma subcarrier oscillator can produce transposed colors, ie, red for green, blue for red, etc. If the oscillator is badly misaligned, the color may be entirely absent from the televised picture.

11-1142. Improper color rendition can be attributed to mechanical adjustments such as the color purity, convergence, and black-and-white level adjustments. In a color television receiver, defects in the mechanical adjustments mentioned can be easily detected when receiving black-and-white program material. Color purity defects can cause a uniform color tint across the cathode-ray-tube screen. Alignment of the color purity magnets is required when the central viewing area produces an acceptable black-and-white picture and colors appear only at the outer rim of the picture. Flashes of color through the black-and-white picture do not indicate alignment defects; they indicate a malfunctioning color-killer circuit.

11-1143. Alignment of the convergence magnets and circuits is usually required when the black-and-white images have colored edges or fringes. You should first inspect the black-and-white image very closely in the central viewing area, and then in the outer regions toward the edges of the picture. Follow this by an inspection to see whether such a defect is more prevalent in the horizontal or vertical dimensions of the picture. Finally, inspect the picture for predominance of a single color.

11-1144. A total alignment of the convergence circuits is required when the color fringes appear in both the central and outer portions of the picture. The check at the outer edges of the picture will show you whether the dynamic convergence circuits are faulty. The check with respect to the horizontal and vertical dimensions will indicate whether the horizontal or vertical, or both, dynamic convergence circuits need alignment. As you check for the predominance of a single color, you must also consider whether alignment of the color matrix section is required. A color matrix is not employed in all color television sets. Refer to the schematic diagram of the receiver. If the luminance (Y) signal is injected into the cathodes of all three electron guns, and the demodulated color signal is introduced into the grid of its respective gun, no color matrix section is used.

11-1145. PRELIMINARY CATHODE-RAY-TUBE ADJUSTMENTS.

WARNING

A serious shock hazard exists in the high voltage section of a color television set. Do not forget to discharge the high voltage before you touch any leads.

11-1146. The high voltage section of a color television set is quite different from the high voltage section of a black-and-white television set. As you will recall, in black-and-white telvision sets the filter capacitor has a value of approximately 500 picofarads. Color television sets, however, use a filter capacitor on the order of 4 nanofarads; this means that an increase of 8 times as much energy can be released on accidental exposure. Another factor contributing to the dangerous shock hazard in a color television set is the tremendously improved high voltage regulation.

11-1147. Some color television sets employ an interlock spring as a safety device. The purpose of this spring is to discharge the high voltage power supply to ground when the cabinet cover is removed from the receiver. However, certain adjustment procedures require the measurement of these high voltages. If the spring is not disabled, the output high voltage will be grounded and the television set high voltage supply can be damaged; on the other hand, disabling this spring removes all attempts by the manufacturer to protect you from a shock hazard (that is, the cabinet cover is removed and you are exposed to operating high voltage).

11-1148. PRELIMINARY HIGH VOLTAGE POWER SUPPLY ADJUSTMENTS. You must adjust the high voltage power supply in a color television set before you attempt any convergence adjustments. Accordingly, turn the television set off and remove the back of the cabinet. Remove the high voltage cage covering and discharge the high voltage supply to ground. Disable the safety interlock spring, if any, and then apply power to the television set. Turn the brightness and contrast controls to produce a very dim raster on the cathode-ray tube. Using an electronic voltmeter and a 50-kilovolt high voltage probe, measure the high voltage at the corona shield. Adjust the high voltage regulator control until the meter indicates the correct voltage. Disconnect the high voltage probe.

11-1149. Adjust the line voltage, if possible, by means of a variac or connection of the television set into a source of regulated 115 volts ac. Tune in a picture signal, either color or black-and-white, by means of the channel selector and fine tuning controls, and adjust the horizontal and vertical hold controls for a stable television picture.

11-1150. Turn the power to the receiver off; then remove the horizontal output fuse, generally located on the rear chassis apron, under a metal cover. Connect a 10k resistor across the fuse terminals. Next, you should connect an electronic voltmeter across the 10k resistor you have just connected into the circuit. Set the voltmeter function switch to the positive dc volts position, and the range switch to the 3v or 5v position, whichever is available. After applying power to the television set, adjust the plate limiting coil in the receiver until the voltmeter indicates 2.1 volts.

11-1151. Turn the horizontal drive control until the drive line is visible near the center of the viewing area. Reduce the setting of this control until the drive line just disappears. Observe the electronic voltmeter; if the voltage has changed, reset the plate limiting coil until the meter again indicates 2.1 volts. Now you can turn the television set off, remove the resistor and voltmeter, and replace the fuse in its holder. Replace the metal fuse cover also.

11-1152. VERTICAL AND HORIZONTAL DIMENSIONAL ADJUSTMENTS. Connect a crosshatch generator to the antenna terminals of the color television receiver. Operate the channel selector control of the color television set to channel 3 or channel 4, depending upon the crystal drive unit supplied with the crosshatch generator. Turn the output selector switch of the signal generator to the rf carrier position. Set the pattern selector to the crosshatch lines position, and then set the rf output control to approximately its mid-range position.

11-1153. Inspect the crosshatch display carefully in the vertical dimension for the following normal conditions; refer to figure 11-144.

a. The crosshatch display is vertically centered on the viewing screen of the cathode-ray tube.

b. The top and bottom of the test pattern extends to, but does not exceed, 1/4 inch beyond the edges of the picture mask opening.

c. The horizontal lines of the test pattern are spaced equal distances apart from top to bottom of the crosshatch pattern.

11-1154. Television rasters that are not centered on the picture tube in the vertical dimension can be adjusted by means of a vertical centering control. The vertical centering control is usually a potentiometer located on the rear chassis apron of the television receiver. Color television equipments do not employ centering magnets placed around the neck of the cathode-ray tube.

11-1155. If the crosshatch pattern does not fill the viewing area from top to bottom, adjust the vertical size (height) control alternately with the vertical linearity control. Stabilize the display, using the vertical hold control, if the display tends to roll. Remember that the vertical size and linearity controls are interacting controls. Adjustment of the vertical height control affects the height of the entire picture area, but it affects the bottom half more than the top half. The vertical linearity control does not produce much effect in changing the bottom edge of the picture, but it does have a very pronounced effect on the top half of the picture.

11-1156. In case a test pattern is not available, adjust the vertical hold control until the picture begins to roll slowly. Inspect the vertical blanking bar for the same thickness as it moves completely across the screen. Use the vertical size and linearity controls for this adjustment.

11-1157. Now you can direct your attention to the conditions existing in the horizontal dimensions; the following conditions are normal:

a. The crosshatch display is horizontally centered on the viewing screen of the cathode-ray tube.

b. The right- and left-hand edges of the test pattern extend to, but do not exceed, 1/4 inch beyond the edges of the picture mask opening.

c. The vertical lines of the crosshatch test pattern are spaced equal distances apart from side to side of the picture.

11-1158. Off-center television rasters can be centered by adjusting the horizontal centering control. The horizontal centering control is a potentiometer, and is generally located on the rear chassis apron.

11-1159. If the vertical lines of the crosshatch pattern are crowded closer together at either side of the picture, adjust the horizontal linearity control to correct this condition. You can use the horizontal size (width) control to adjust the pattern for the proper width with respect to the picture mask edges. You should not make the mistake of expecting the horizontal linearity control to react only on one side of the picture, as you have learned about the vertical linearity control. The horizontal linearity control reacts by crowding one side of the picture

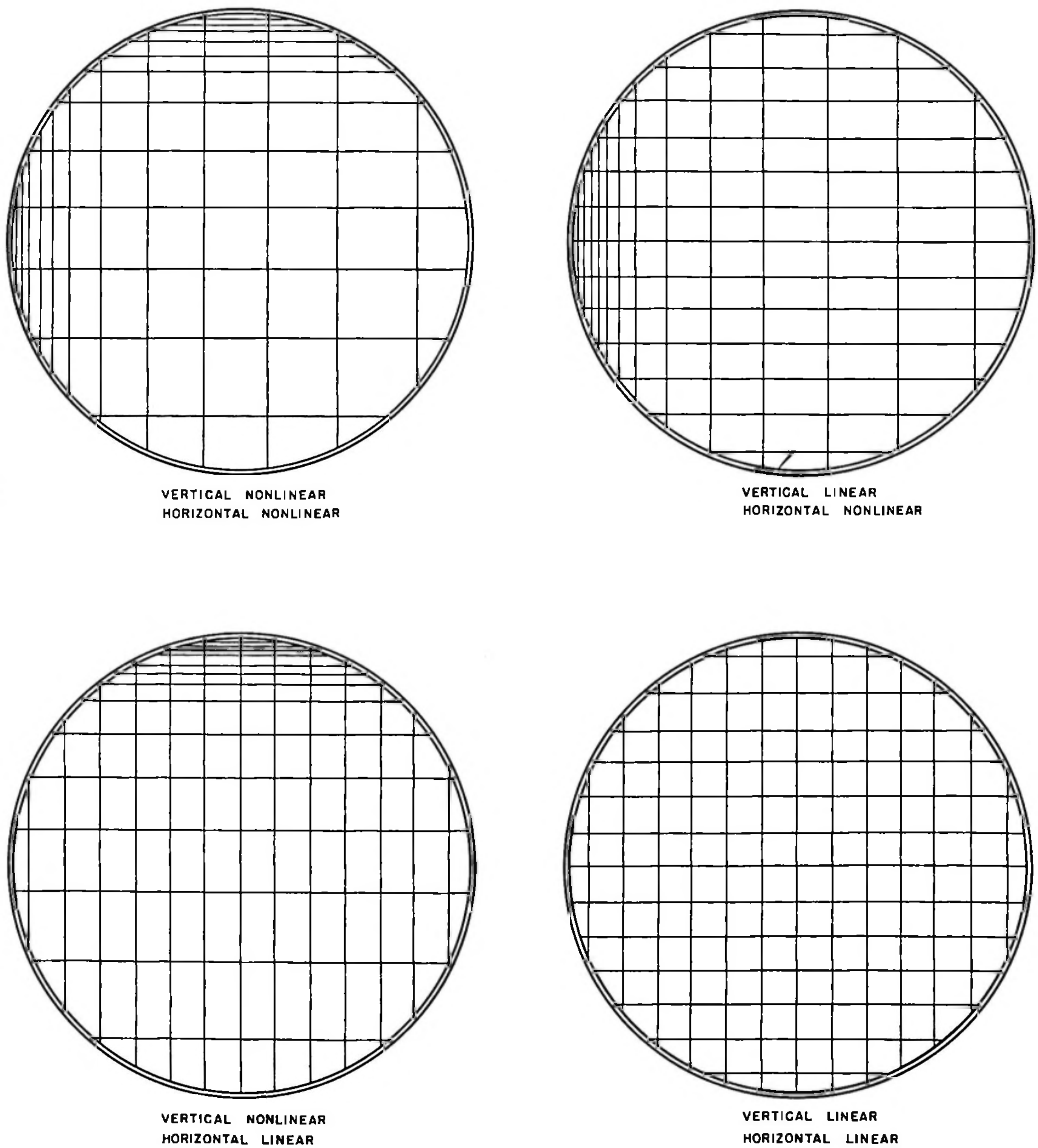


Figure 11-144. Horizontal and Vertical Dimensional Characteristics

and stretching the other side, and vice versa.

11-1160. FINAL HIGH VOLTAGE POWER SUPPLY ADJUSTMENTS. You can set the contrast and brightness controls, located on the front of the color television receiver, to approximately their mid-range points. Referring to paragraphs 11-1149 and 11-1150, measure the high voltage at the corona shield. The electronic voltmeter should indicate the correct high voltage; if it does not, adjust the high voltage regulator. Disconnect the high voltage probe from the corona shield, advance the brightness control to its maximum clockwise position, and measure the high voltage again. The voltage indication should be about 500 volts more than previously obtained, but should not exceed the previous measurement by more than 1000 volts.

11-1161. HORIZONTAL OSCILLATOR AND BURST TIMING ADJUSTMENTS. You are now prepared to accomplish the final preliminary cathode-ray-tube adjustments. Set the horizontal hold control of the receiver to approximately its mid-range position. Connect a temporary jumper across the terminals of the horizontal stabilizing coil, if one is used, and another jumper from ground to the output of the horizontal phase comparator circuit. The horizontal frequency control must now be adjusted so that the picture is moving very slowly across the face of the cathode-ray tube. You must not expect the crosshatch display to remain completely stationary as long as the horizontal stabilizing coil is short-circuited.

11-1162. Remove the jumper from the stabilizing coil and adjust the coil until the picture is as stationary as possible. Carefully inspecting the television display, you should rotate the horizontal hold control fully clockwise and count the number of blanking bars in the display. You should count four bars for this position of the control. Rotate the horizontal hold control to its fully counterclockwise position, and observe the display for three blanking bars. If you cannot obtain this condition, readjust the horizontal frequency control slightly and try the above adjustment again. While the use of three blanking bars in the first case and four in the second case is strongly suggested, it is not an ironclad rule. The requirement is to obtain one more bar in the clockwise position than in the counterclockwise position.

11-1163. Connect an oscilloscope to view the waveform at the grid of the burst gate amplifier. Remove the temporary jumper from across the horizontal phase comparator, and adjust the oscilloscope controls for a stable display of convenient size. Turn the horizontal hold control to its maximum clockwise position. As you adjust the burst time control, make the right-hand edge of the burst signal line up with the right-hand edge of the gate pulse; refer to figure 4-246.

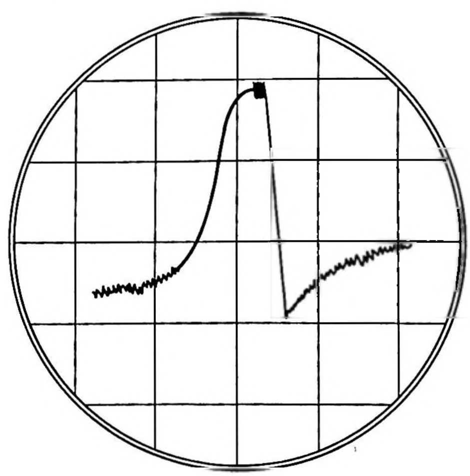


Figure 11-145. Positioning of Burst on Gating Pulse

11-1164. All preliminary cathode-ray-tube adjustments are completed at this time. You can now proceed with the preliminary convergence adjustments if they are required.

11-1165. PRELIMINARY CONVERGENCE ADJUSTMENTS.

11-1166. GENERAL. Before performing the preliminary convergence adjustments, you should first understand the terms used in the explanation. References to G_1 should be understood as meaning the control grid of an electron gun, and G_2 should be understood as meaning the screen grid of an electron gun. In addition, you should understand the use of a red G_2 control. The phosphors used in a cathode-ray tube to obtain a red trace have a very low efficiency as compared with the phosphors used for green and blue traces. Many color television sets, accordingly, provide attenuation for the green and blue beams only. In cases where the television set does not provide the red G_2 control, the master brightness control can be adjusted as a substitute.

11-1167. PRELIMINARY STATIC CONVERGENCE ADJUSTMENTS. Connect an rf cable between the rf output terminal of the generator and the antenna terminals of the color television receiver. Turn the receiver channel selector to channel 3 or channel 4, depending upon the crystal frequency of the signal generator. Turn the rotary output selector switch of the generator to the rf carrier position. Set the pattern selector to the crosshatch lines position, and then turn the rf output control to approximately its mid-range position.

11-1168. During the period required for the television receiver to warm up, you can preset the receiver controls as follows:

a. Set the red G_2, the green G_2, and blue G_2 controls for approximately equal brightness

b. Set the vertical amplitude controls for the red, green, and blue amplifiers to their maximum counterclockwise positions. Then you hould apply the same procedure to the horizontal amplitude controls for the red, green, and blue amplifiers.

c. Set the red, green, and blue vertical tilt controls to approximately their mid-range positions.

11-1169. STATIC CONVERGENCE. You should now closely inspect the display of the television cathode-ray tube. The resulting pattern of a properly adjusted receiver is a crosshatch of white horizontal and vertical lines. These lines will be white because the red, green, and blue electron beams are made to converge and provide equal excitation of the cathode-ray tube screen phosphors. You will also notice that the convergence at the center of the screen is very good while the convergence becomes poorer nearer the outer edges of the cathode-ray tube. While the convergence becomes poorer near the outer rim, it is important that the three beams be equally displaced in all corner areas for both the horizontal and vertical lines.

11-1170. If the three beams do not converge, the static convergence magnets located on the neck of the cathode-ray tube (refer to figure 11-136) must be adjusted until convergence is obtained. The coil and magnet assemblies, shown in figure 11-146, may be adjusted by turning the magnet on the top of the desired coil and rotating the magnet until convergence of the colored beams is obtained in the central portion of the picture tube; refer to figure 11-147. A dark-metal wing nut is used to index the position of the blue convergence coil. It is quite possible that you will have difficulty in converging the blue vertical lines with the red and green lines. You may find it advisable to adjust the blue lateral magnet in order to obtain blue convergence; refer to figure 11-136 for the location of the blue lateral magnet assembly.

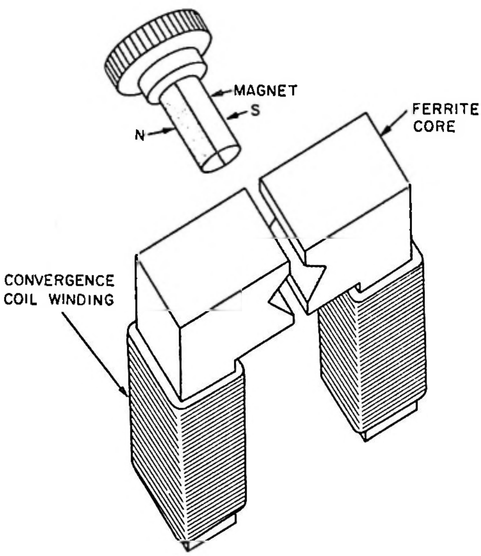


Figure 11-146. Mechanical Details of Static Convergence Magnet Assembly

11-1171. The amount of misconvergence at the corner areas can be equalized by repositioning the deflection yoke horizontally (side-to-side) or vertically (up-and-down) on the cathode-ray-tube neck.

11-1172. The preliminary static convergence adjustment has been completed. You will now be able to closely examine the crosshatch display for other defects, such as display centering, proper height, width, and linearity, etc. These controls are generally located on the rear chassis apron of the color television receiver. The focus control should now be adjusted to obtain well defined horizontal and vertical lines. Color television receivers do not employ permanent magnet focusing (or centering) mechanisms along the neck of the picture tube. Focusing coils are used, and the current through these coils are controlled by a potentiometer located on the rear chassis apron of the receiver.

11-1173. PRELIMINARY DYNAMIC CONVERGENCE ADJUSTMENTS. Reduce the blue G_2 control to its minimum (counterclockwise) position. You should now observe that the blue display lines disappear, that the crosshatch lines are yellow where they were converged in previous steps, and that the outer portions of the display where you could not get convergence are separate red and green crosshatch lines.

11-1174. Carefully observe the horizontal crosshatch lines while you turn both the red vertical amplitude control and the green vertical amplitude control. Adjust these controls so that the red and green lines are as nearly parallel as possible. Now you should make the two colored lines converge in the center of the display area, using the red and green static convergence magnets on the neck of the cathode-ray-tube assembly; refer to figure 11-148.

11-1175. Turn the blue G_2 control to approximately its previous position so that the blue lines are just as bright as the red and green lines. Observe the horizontal crosshatch lines closely while adjusting the blue vertical amplitude control to get the blue lines as nearly parallel as you can with the red and green crosshatch lines; refer to figure 11-149. Now converge the blue lines with the red and green lines, using the blue static convergence magnet. Observe that the lines, where they are converged in the central portion of the picture area, are nearly white. Refer to figure 11-150.

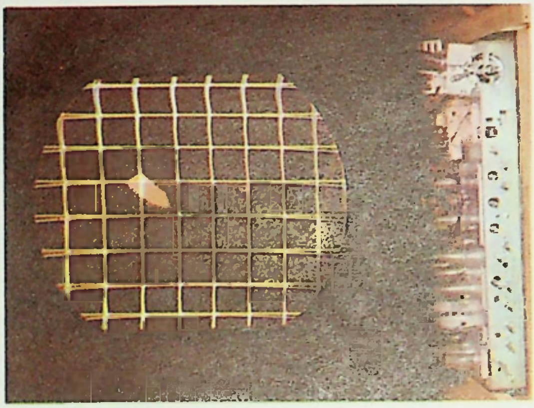


Figure 11-147. Preliminary Static Convergence

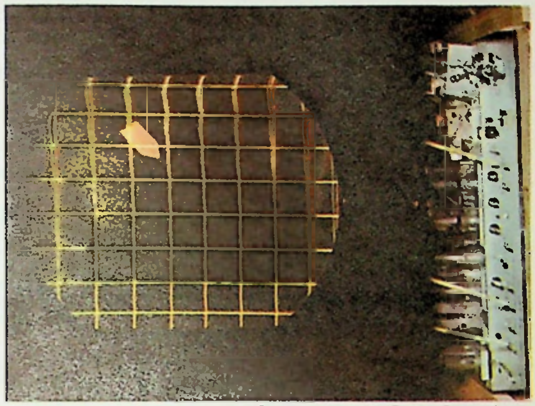


Figure 11-148. Adjustment of Red and Green Vertical Dynamic Amplitude Controls

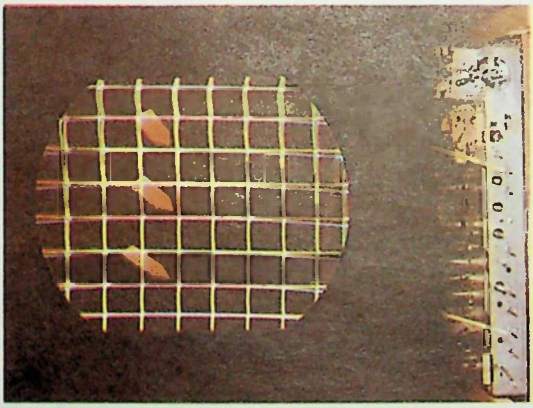


Figure 11-149. Adjustment of Blue Vertical Dynamic Amplitude Control

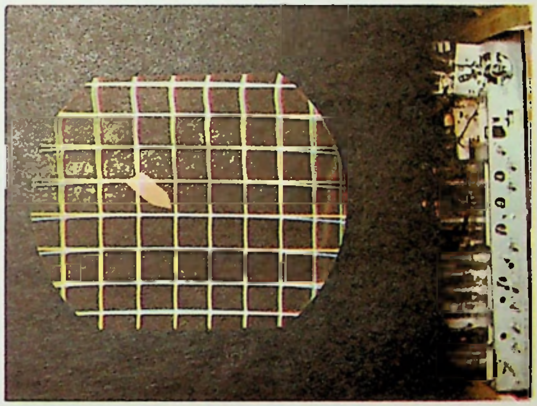


Figure 11-150. Red, Green and Blue Static Convergence

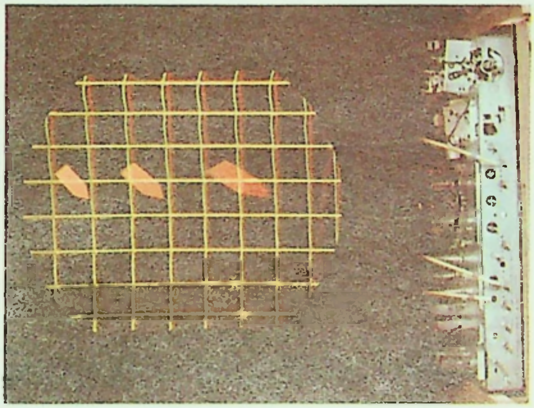


Figure 11-151. Adjustment of Red Horizontal Dynamic Phase Control

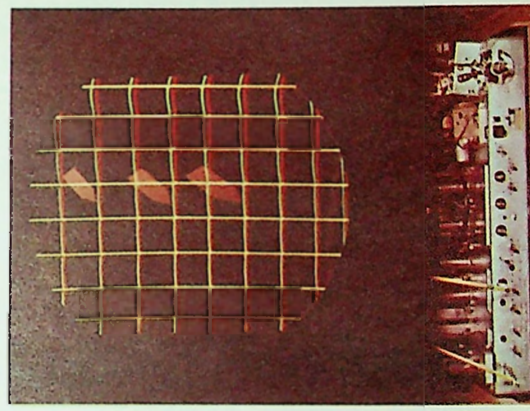


Figure 11-152. Adjustment of Green Horizontal Dynamic Phase Control

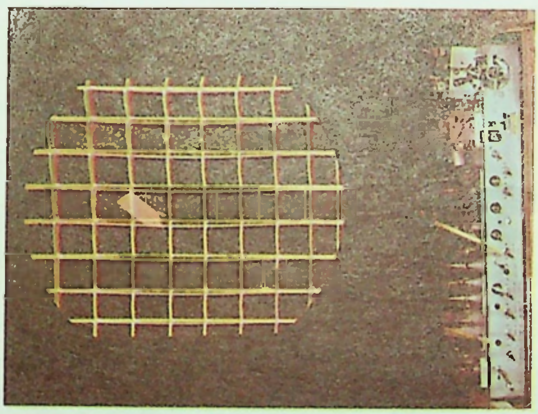


Figure 11-153. Adjustment of Blue Horizontal Dynamic Phase Control

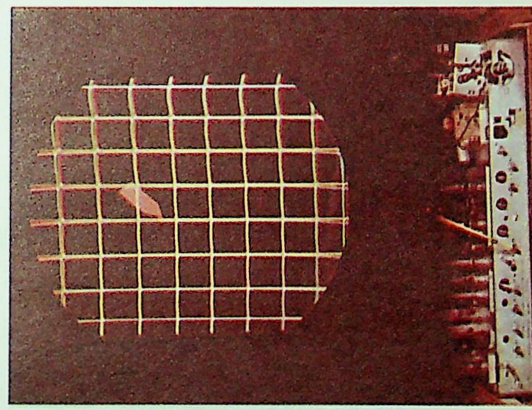


Figure 11-154. Adjustment of Blue Horizontal Amplitude Control

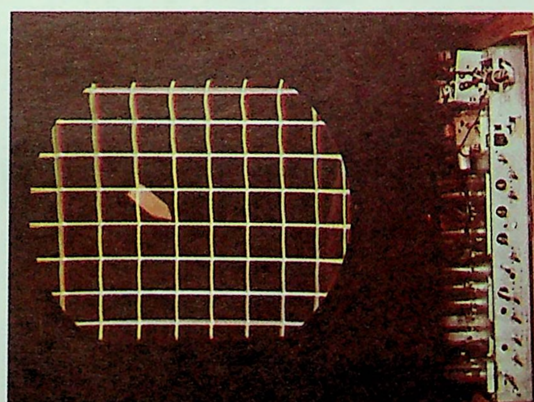


Figure 11-155. Final Red, Green, and Blue Static Convergence Resulting from Preliminary Convergence Procedure

11-1176. Turn the blue G_2 control to its minimum (counterclockwise) position, and again observe that the blue crosshatch lines disappear from the display. You are now ready to obtain convergence by observing the vertical crosshatch lines while you are adjusting the horizontal amplitude controls.

11-1177. Turn the red horizontal amplitude control to its maximum brightness position. Adjust the red horizontal phase control so that the vertical red crosshatch lines are parallel with respect to the green crosshatch lines; refer to figure 11-151. You should now reverse the process by setting the red horizontal amplitude control to the minimum brightness position, and advancing the green horizontal amplitude control for maximum brightness in the display. Adjust the green horizontal phase control so that the vertical green crosshatch lines are parallel with respect to the red crosshatch lines; refer to figure 11-152. In this step, notice that the two lines may tilt slightly with respect to each other as they are adjusted, but make sure that their displacements are equal and parallel. If these lines are not curved the same with respect to each other (parallel), the adjustment must be made again because the amplitude ratio of red to green is maladjusted. You should reconverge the red and green crosshatch lines by adjusting their respective static convergence magnets on the picture tube neck.

11-1178. Advance the blue G_2 control once more so that the blue lines are as bright as the red and green crosshatch lines. Place the blue horizontal amplitude control in its maximum clockwise position. Adjust the blue horizontal phase control for a symmetrical and parallel displacement from the red and green lines; refer to figure 11-153. Following this, you should adjust the blue horizontal amplitude control to bring the vertical blue crosshatch lines parallel with the red and green lines, refer to figure 11-154. Complete the preliminary convergence adjustments by adjusting all three static convergence magnets, if necessary, on the neck of the cathode-ray tube; refer to figure 11-155.

11-1179. COLOR PURITY ADJUSTMENTS. The color purity adjustments of a color television receiver should not be attempted unless convergence of the three electron beams is known to be correct. If there is any doubt about the condition of the convergence circuit adjustments, perform the preliminary convergence tests discussed in the preceding paragraphs. If the preliminary convergence tests were properly made, then you should temporarily disconnect the crosshatch generator from the receiver antenna terminals.

11-1180. Set the green G_2 and blue G_2 controls to their minimum (counterclockwise) position, observing that the blue and green fields disappear. The remaining field is red. Each magnet in the magnetic-field equalizer assembly must be pulled outward into its housing to produce minimum field intensity. Next, the color purity control, located on the rear chassis apron of the receiver, is set approximately to its mid-range position and the field neutralizing coil control to approximately its mid-position. However, not all color television receivers use the neutralizing coil assembly because they depend very heavily upon the equalizer assembly.

11-1181. Some television receivers employ a color purifying coil assembly around the neck of the cathode-ray tube. Other color receivers use a dual magnetic ring as the color purifying magnet assembly, mounted on the neck of the cathode-ray tube, as shown in figure 11-156. The rings can be rotated individually with respect to each other; in addition, the entire assembly can be rotated around the tube neck.

11-1182. Rotate the color purifying assembly around the neck of the cathode-ray tubes,

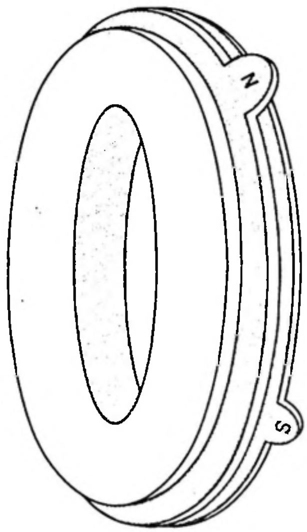


Figure 11-156. Color Purifying Magnet and Ring Assembly

When you are observing the display, disregard the size of the red area; concentrate on the purity of the red in the central portion of the display area. After the entire assembly has been rotated, try increasing the purity of the red display by rotating the rings with respect to each other. Separating the tabs increases the field strength of the magnetic assembly, while bringing them closer together decreases the field strength. To increase the field strength of a purifying coil, rotate the purity control in a clockwise direction. In any case, use the least amount of correction to produce the desired result; refer to figure 11-157.

11-1183. Loosen the retaining screws of the deflection yoke. Then slide the yoke either backward or forward until the largest uniform red field is obtained; refer to figure 11-158. There are several precautions to be taken with this part of the adjustment:

a. When moving the deflection yoke forward and backward, do not shift it vertically (up or down) or laterally (side to side); otherwise, all the convergence adjustments will be destroyed.

b. Do not strike or accidentally bump the convergence coil and magnet assembly because this is likely to destroy all the convergence adjustments made so far.

c. When moving the deflection yoke backward, remember that a shadow can be introduced in the corner of the display area. It is also possible to introduce a shadow which is apparent only on the blue or green field. The remainder of this procedure provides a check on this type of trouble.

11-1184. At this point in the procedure, color contamination should exist only in the outer rim of the picture display area. You may wish to readjust the color purifying magnets again after you securely lock the deflection yoke in place. The color contamination existing at the rim of the display area can be eliminated by moving the equalizer magnet that is closest to the contaminated area toward the soft iron ring surrounding the picture tube; refer to figure 11-159. The equalizer magnets may also be rotated in their fixed housing to increase color purity by causing the fields of the equalizing magnets to aid the field of the purifying magnets within a localized portion of the display.

11-1185. Set the red G_2 control to its minimum position and observe the disappearance of the red field. Advance the green G_2 control and examine the display for color contamination and shadows in the corner of the display area; then return the control to its former position. Advance the blue G_2 control and examine the blue field for the same type of defects. If contamination exists, the equalizing magnets should be adjusted for a compromise setting which gives the best purity in all color fields.

11-1186. At this point, you must establish a preliminary white balance. First set the

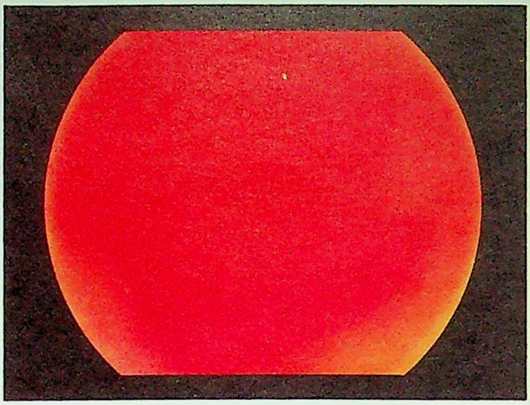


Figure 11-157. Preliminary Red Field Purity

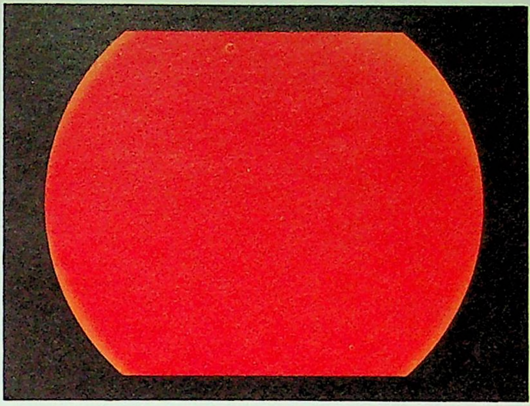


Figure 11-158. Improved Red Field Purity by Deflection Yoke Adjustment

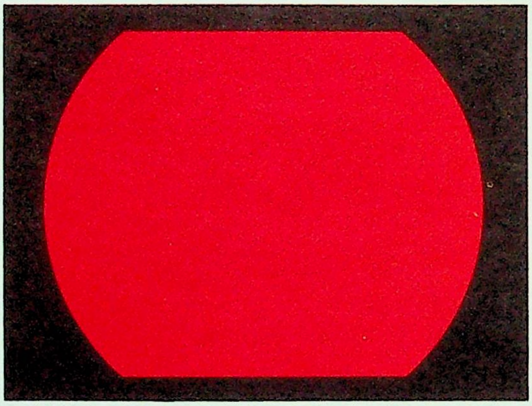


Figure 11-159. Final Red Field Purity

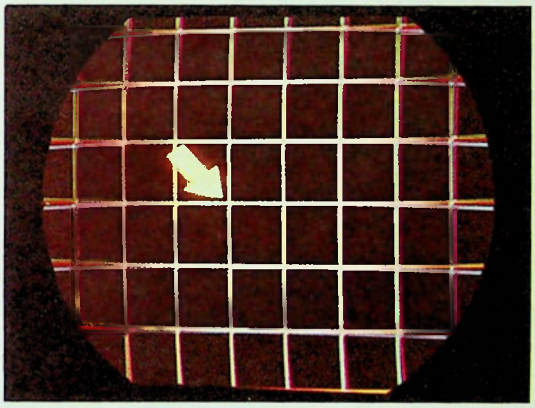


Figure 11-160. Static Convergence of Red, Green and Blue Beams

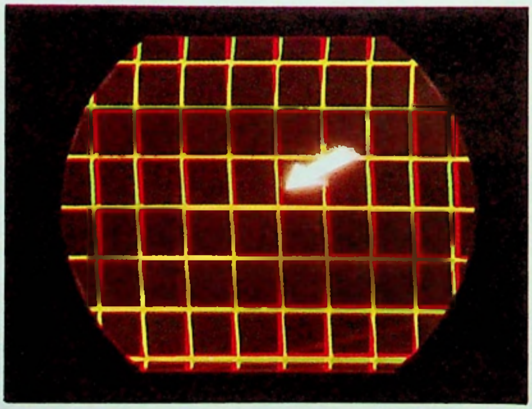


Figure 11-161. Adjustment of Red Vertical Tilt Control

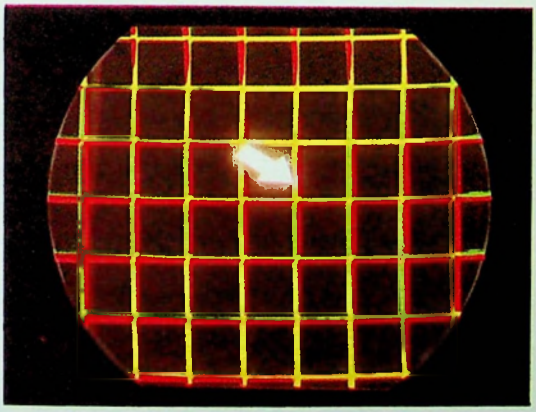


Figure 11-162. Adjustment of Green Vertical Tilt Control

red G_2 control to approximately its mid-range position, and the blue and green G_2 controls to their minimum positions. The display should now be red. Advance the green G_2 control to obtain a yellow field, and then advance the blue G_2 control to obtain a white field. Inspect the display for color impurities again, using the equalizer magnets as necessary to correct the white balance. Advance the purity control through its entire range while observing the change in color purity of the display. Set the purity control to the position which gives the most uniform white field.

11-1187. FINAL CONVERGENCE ADJUSTMENTS. Connect the crosshatch generator to the antenna terminals of the color television receiver, and adjust the generator controls as specified in paragraph 11-1167.

11-1188. Preset the receiver controls as follows:

a. Set the red, green, and blue vertical amplitude controls to their maximum counterclockwise position.

b. Set the red, green, and blue vertical tilt controls to their mid-range positions.

c. Adjust the static convergence magnets so that the crosshatch lines converge to form white lines in the center of the display area; see figure 11-160.

d. Set the blue G_2 control to its minimum position so that a yellow line appears where the beam is converged, and red and green lines appear where the beams are not converged.

11-1189. Operate the red vertical amplitude control to its maximum clockwise position. Examine the red and green vertical lines while adjusting the red vertical tilt control, observing a symmetrical bowing of the red vertical lines with respect to the green ones, as shown in figure 11-161. Reverse the procedure by turning the red vertical amplitude control to its minimum position and advancing the green vertical amplitude control to its maximum position. Observe the same bowing action, as shown in figure 11-162, for the green vertical line as you did previously for the red vertical line.

11-1190. Set the red and green vertical amplitude controls to their minimum positions; advance each control alternately a little at a time until the red and green vertical lines converge to form a single yellow line.

11-1191. Examine the horizontal lines for convergence very closely. If the lines are not converged, it is extremely important that the spacing between the red and yellow horizontal lines is the same for all horizontal lines in the crosshatch display. Ask yourself the following questions:

a. Does the separation appear greater at the top and bottom of the display than at the center?

b. Do the lines converge fairly well at the bottom of the display, but seem to get farther and farther apart near the top of the display?

c. Does the separation appear to be uniformly the same from top to bottom?

11-1192. If the answer to the first question is yes, then the ratio of the red vertical amplitude control to the green vertical amplitude control is incorrect. Adjust these two controls carefully to obtain the required equal spacing. At the same time, examine the red and green vertical lines for convergence. If the vertical lines do not remain converged, there is no serious trouble as long as they remain parallel. Converge the red and green beams by means of the static convergence magnets; refer to figure 11-163.

11-1193. If the answer to the second question is yes, then the red and green vertical tilt controls must be adjusted with respect to each other to obtain the required equal spacing. If the vertical lines tend to separate, there is no serious trouble as long as they remain parallel to each other. When the spacing adjustment has been accomplished correctly, you must reconverge the display as shown in figure 11-163 by means of the static convergence magnets.

11-1194. If the answer to the third question is yes, you can simply adjust the static convergence magnets to converge the entire crosshatch display, as illustrated in figure 11-163.

11-1195. Advance the blue G_2 control to the same brightness as the red and green lines. Inspect the display for convergence at the center of the screen; if the beams are not converged, adjust the blue lateral convergence magnet assembly on the neck of the cathode-ray tube to correct this condition.

11-1196. Set the blue vertical amplitude control to its maximum position, and adjust the blue vertical tilt control for equal spacing of the horizontal blue lines with respect to the horizontal yellow lines, as shown in figure 11-164. Adjust the blue vertical amplitude control to straighten out the blue lines, as shown in figure 11-165; then bring all three beams into convergence, using the static convergence magnet assembly, to obtain the result shown in figure 11-166.

11-1197. Reduce the blue G_2 control to minimum. Leave the blue and green horizontal amplitude controls at their minimum positions, and advance the red horizontal amplitude control to its maximum clockwise position. Adjust the red horizontal phase control for equal spacing of the vertical red lines from the green vertical lines for all the vertical crosshatch lines in the display area; see figure 11-167. Then return the phase control to its minimum position. Reverse the procedure by advancing the green horizontal phase control for equal spacing of green vertical lines from the red vertical lines, as shown in figure 11-168, and then return the control to its minimum point.

11-1198. Alternately, slightly increase the red and green horizontal phase controls. As the red and green horizontal lines approach each other, they should be parallel, as shown in figure 11-169. If they are not parallel, that is, if they are curved toward or away from each other, the amplitude ratio of red to green is not balanced. It is also possible that the red line crosses over the green line, or vice versa. In this case, the red and green horizontal phase controls must be readjusted. Reconverge the red and green lines in the central portion of the picture area by means of the static convergence magnets on the neck of the cathode-ray tube.

11-1199. Advance the blue G_2 control, and set the blue horizontal amplitude control to its maximum position. Adjust the blue horizontal phase control for a symmetrical curvature of the blue horizontal lines with respect to the yellow horizontal lines; see figure 11-170. Then you may adjust the blue horizontal amplitude control to straighten out the blue horizontal lines to make them parallel with the yellow lines. Converge the three beams by means of the static convergence magnets; see figure 11-171.

11-1200. The convergence adjustments are now completed. It is possible that there is a slight amount of misconvergence remaining in the corner of the picture viewing area, but further convergence adjustments will not correct this condition. Slight misconvergence in the corner areas are easily tolerated when the entire picture area is being viewed.

11-1201. WHITE BALANCE ADJUSTMENTS.

11-1202. GENERAL. Calibrate the vertical channel of an oscilloscope to provide a peak-to-peak deflection of 65 volts, and connect the oscilloscope between ground and the cathode of the red electron gun in the receiver cathode-ray tube. Then adjust the oscilloscope horizontal frequency and synchronization controls to observe a stationary display of the video signal.

11-1203. Turn the brightness control of the television receiver to approximately its mid-range position. Now adjust the contrast control of the receiver so that the video display of the oscilloscope is also 65 volts, peak to peak. As you observe the oscilloscope display, turn the receiver brightness control, carefully inspecting the synchronization pulse tips for compression. Then turn the brightness control in the opposite direction until the compression just disappears.

11-1204. Set the blue G_1 and G_2 controls to their minimum (counterclockwise) position, and then set the green G_1 and G_2 controls to their minimum (counterclockwise) position. Observe that the color television receiver displays only a red image of the black-and-white televised picture. Adjust the vertical hold control of the receiver until the vertical blanking bar of the televised picture is midway between the top and bottom of the picture display area; see figure 11-172. Adjust the red G_2 control until the vertical blanking bar just turns black, and then readjust the receiver vertical hold control to bring the picture back into vertical synchronization.

11-1205. BLACK-AND-WHITE BAR PATTERN METHODS. If you have a color bar generator that also produces black-and-white bar patterns, such as the Philco Model 7100, disconnect the antenna from the television receiver, and connect the generator to the receiver antenna terminals. These black-and-white bars are more convenient to use because they are more stable and because they eliminate errors due to guessing at "almost" black or "almost" white picture areas.

11-1206. Set the color bar generator to channel 3 or channel 4, depending upon the crystal frequency of the generator. If you use the Model 7100 Color Bar Generator, set the OUTPUT SELECTOR switch of the color bar generator to the rf carrier position. Turn the pattern selector control to the black-white bars position, and then turn the rf output control to its mid-range position. Tune the color television receiver fine tuning control until the 920-kilocycle beat note is reduced to a minimum. One horizontal dark bar and one horizontal light bar should appear in the television receiver picture area.

11-1207. Adjust the green G_1 and G_2 controls for a yellow receiver display; refer to figure 11-173. It is important that you notice the areas of the picture affected by the G_1 control and the G_2 control. The G_1 control affects the lighter portions of the resulting display to a greater extent than it does the darker areas. The G_2 control works in just the opposite way. It is possible, therefore, for the color display to present different colors, depending upon the ratio of the G_1 and G_2 control settings.

11-1208. Advance the green G_1 control until the lighter portion of the display is nearly yellow, and the darker portion of the display is still red-tinged. Then as you advance the green G_2 control, the darker portion of the display will approach yellow, and the lighter portion of the display will become completely yellow.

11-1209. Advance the blue G_1 control until the lighter portion of the display is nearly white, then adjust the G_2 control for a black bar in the darker portion of the display area. The interaction of the G_1 and G_2 controls should bring the nearly white portion of the display to full white.

11-1210. Refine the green G_1 and G_2 and the blue G_1 and G_2 controls, if necessary, to obtain the white balance shown in figure 11-174. Do not disturb the red G_2 control. This completes the adjustment of the color television receiver for white balance.

11-1211. BLACK-AND-WHITE PICTURE METHOD. In case a generator that produces black-and-white bar patterns is not available, you can use a black-and-white program as an alternate method of adjusting the G_1 and G_2 controls. This method is outlined in the following procedure:

a. Turn the contrast control and blue and green G_2 controls to their minimum (fully counterclockwise) positions; then turn the brightness control fully clockwise to its maximum position.

b. Adjust the red G_2 control for an image that is just barely visible. Advance the green G_2 control until the image becomes a green-yellow color, and then advance the blue G_2 control until the image becomes a neutral gray.

c. Turn the contrast control to receive a normal picture; then readjust the blue G_1 and the green G_1 controls until the picture is black-and-white.

d. Reduce the screen brightness until the picture is barely visible; then readjust the blue and green G_1 controls for a gray picture. Turn both the brightness control and the contrast control throughout their ranges. If the picture exhibits any tinting, refine the adjustments, especially the G_1 controls.

11-1212. RF TUNER ADJUSTMENTS.

11-1213. TUNER OSCILLATOR. Set the channel selector of the television receiver to channel 13, and set the fine tuning control to its mid-range position. Remove the knobs to the channel selector and fine tuning control. Remove the antenna lead-in from the receiver antenna terminals.

11-1214. Disconnect the automatic gain control bus line and substitute a fixed source of bias to the tuner. A suitable bias substitution supply is shown in figure 11-175. Connect the bias supply negative lead to the tuner lead, and the bias supply positive lead to chassis ground.

11-1215. You should also disconnect the B plus lead to the receiver tuner, and connect a 3300-ohm, 1-watt carbon resistor in series between the broken B plus leads. Be sure that the resistor is a carbon unit, because wire-wound resistors are subject to inductive effects that would nullify your alignment. Connect the vertical input lead of an oscilloscope to the tuner side of this added resistor, and then connect the oscilloscope ground lead to the television receiver chassis.

11-1216. Set up a radio frequency signal generator for an unmodulated output signal that is 44 megacycles above the center frequency of the television channel. For example, for channel 13, the calculation is as follows:

a. Frequency limits of channel 13 are 210 to 216 megacycles.

b. Frequency bandwidth of channel 13 is 216-210 = 6 megacycles.

c. Center of frequency range is 6/2 = 3 megacycles.

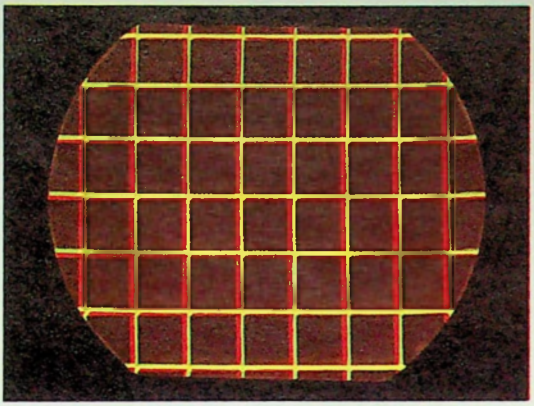


Figure 11-169. Adjustment of Red and Green Horizontal Amplitude Controls

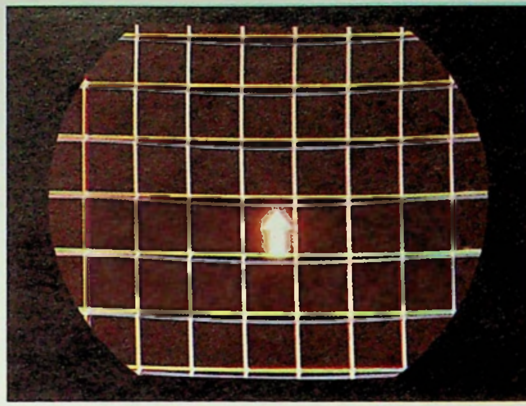


Figure 11-170. Adjustment of Blue Horizontal Phase Control

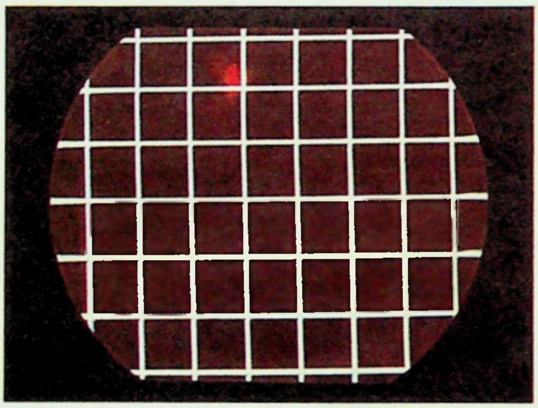


Figure 11-171. Over-all Convergence of Red, Green and Blue Beams

Figure 11-172. Adjustment of Red G_2 Control

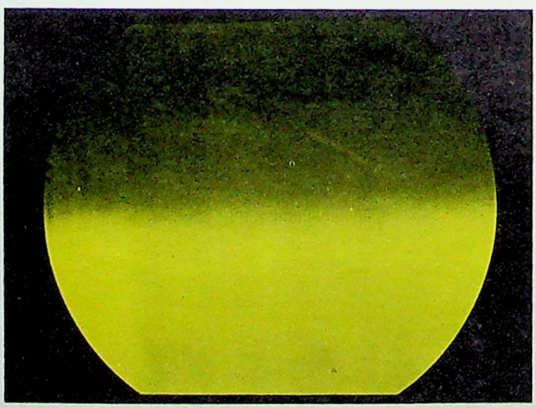


Figure 11-173. Adjustment of Green G_1 and G_2 Controls for Yellow Field

Figure 11-174. Adjustment of Blue G_1 and G_2 Controls for Final White Balance

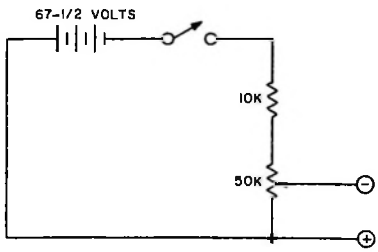


Figure 11-175. Bias Supply Used as a Substitute for Automatic Gain Control Voltage

d. Channel center frequency is 210 + 3 = 213 megacycles.

e. Signal generator frequency is 213 + 44 = 257 megacycles.

11-1217. Connect the signal generator directly to the antenna terminals of the television receiver. Connect the horizontal amplifier of the oscilloscope to the receiver antenna terminals, and adjust the oscilloscope controls to observe a Lissajous display. Advance the voltage from the fixed bias supply until the tuner ceases to overload.

11-1218. The oscillator is aligned by adjusting the oscillator tuning slugs to produce a zero beat as indicated on the oscilloscope. You should inspect the television set to determine whether a turret tuner or an incremental tuner is employed. Incremental tuners must be tuned in strict sequence, beginning with channel 13 and ending with channel 2. Turret tuners can be aligned in any sequence without affecting the remaining channels.

11-1219. Where a turret tuner is employed, a small tuning access hole may be provided in the receiver chassis to the right of the channel selector shaft. Where an incremental tuner is employed, an access hole in the receiver chassis may be provided for each oscillator tuning slug surrounding the channel selector shaft. However, some television tuners do not provide access holes; they require "bunching" or "stretching" of the coil segments until the proper frequency is reached. This process, referred to as "pruning", should be attempted only with the aid of detailed instructions.

11-1220. After the oscillator is aligned, disconnect the radio frequency generator, the oscilloscope, and the 3300-ohm resistor; then reconnect the B plus bus. If you intend to follow the alignment procedure further, do not disturb your other connections, but if you do not intend to follow the procedure any further, restore the television receiver connections from their present temporary condition to normal.

11-1221. TUNER BANDPASS. Before conducting a tuner bandpass alignment for a color television receiver, you must disconnect the tuner output cable from the video intermediate frequency input amplifier, and terminate the open end of the cable with a 40- to 70-ohm carbon resistor. If you are aligning a black-and-white receiver you may ignore this instruction and proceed according to the remaining instructions for bandpass adjustments.

11-1222. Connect the output of the antenna impedance matching network to the antenna terminals of the television receiver, as shown in figure 11-176; then connect the sweep generator output terminals to the input of the impedance matching network. Connect the vertical amplifier of an oscilloscope between ground and the tuner test point provided by the receiver manufacturer. Turn the oscilloscope sweep control to the horizontal amplifier position, and connect the horizontal amplifier of the oscilloscope between ground and the phase terminal of the sweep generator. Loosely couple the marker generator to the input of the receiver, using a 1/2-pf capacitor, as shown in

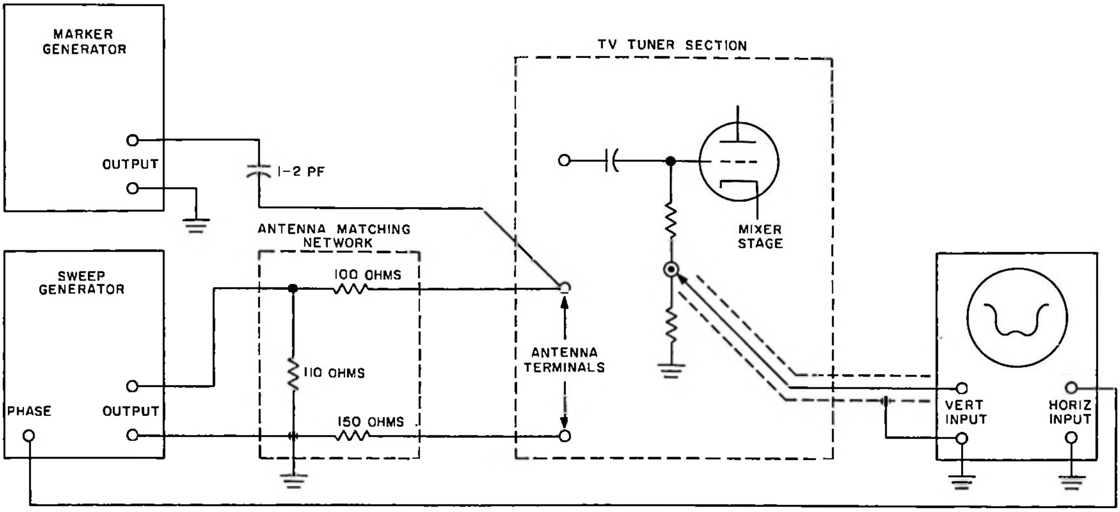


Figure 11-176. Tuner Bandpass Adjustment Test Equipment Setup

the figure, to inject a marker signal. The complete test equipment setup is shown in figure 11-176.

11-1223. Tune the sweep generator to 195 megacycles, and set the sweep control for 10 megacycles. Tune the marker generator to 197.75 megacycles, which is the sound carrier frequency. Operate the television receiver channel selector to channel 10 and adjust the oscilloscope controls to view the rf amplifier/mixer response curve. Use the minimum amount of signal drive from the sweep and marker generators consistent with a convenient oscilloscope display.

11-1224. Tune the trimmer capacitor in the secondary of the antenna transformer first. Tune the trimmer capacitor in the primary circuit of the rf amplifier output transformer. Then tune the trimmer capacitor in the input grid circuit of the mixer stage. Adjustment of the rf amplifier output trimmer capacitor may produce no noticeable effects at this time.

11-1225. You should exercise care to avoid "stagger-tuning" procedures. As you tune the trimmer capacitors involved, tune for maximum amplitude at the center of the response curve, and ignore the amplitude of the peaks of the curve.

11-1226. After these adjustments are made, set the marker generator to 193.25 megacycles, which is the picture carrier frequency. Repeat the trimmer adjustments until the final curve appears as shown in figure 11-177. The dip in the center of the response curve should never exceed 10 percent of the total display amplitude for a color television set. This requirement for black-and-white television sets can be increased to 30 percent. The peaks of the curve must also be the same amplitude, plus or minus 5 percent. Change the sweep generator frequency controls to obtain 85 megacycles, set the marker generator to the sound carrier frequency of 87.75 megacycles, and change the television receiver channel selector to channel 6. You should observe the same shape of curve you obtained before. Set the marker generator to the picture carrier (83.25 mc), observing that the humps of the curve appear at the marker positions. Adjust the trimmer capacitor across the primary winding of the rf amplifier output transformer for maximum amplitude of the oscilloscope display.

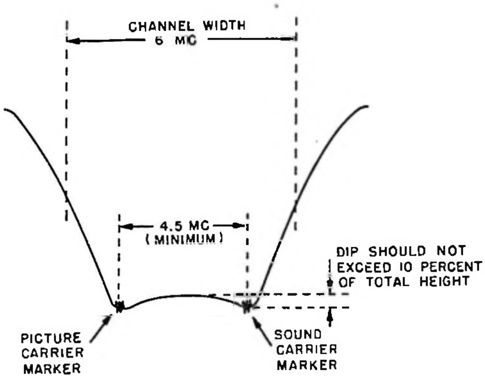


Figure 11-177. Tuner Bandpass Characteristics

11-1227. The response curves of all the remaining channels should appear the same, and should be within the specified tolerance. Check them and make any slight changes in the adjustments that may be necessary. Finally, disconnect all test equipment and restore all television receiver connections from their present temporary condition to normal.

11-1228. VIDEO I-F ALIGNMENT.

11-1229. PREPARATION. Connect an amplitude modulated rf signal generator into the output circuit of the mixer. In some receivers a special input jack is provided to couple the test video signals into the video i-f amplifiers. In other receivers you will have to locate the grid of the first video i-f amplifier and inject the signal at this point. Refer to figure 11-178.

11-1230. Disconnect the automatic gain control bus at its source, that is, the picture detector and agc tube. Connect the negative lead of a substitution bias supply to the agc bus line, and the positive lead to the receiver chassis. Consult the manufacturer's literature for the specific value of fixed bias voltage to be injected into the circuit at this time. Use an electronic voltmeter to measure this voltage during adjustment.

11-1231. Connect one end of a 10k isolation resistor to a probe attached to the vertical input terminals of an oscilloscope, and connect the other end of the resistor to the video detector output circuit. Connect another lead between the receiver chassis and oscilloscope ground. Set the oscilloscope sweep controls for a convenient display, using a linear time base.

11-1232. TRAPS. Adjust the rf signal generator to 47.25 megacycles, and set the modulation control for about 30 percent modulation of the output signal. Tune the 47.25-mc traps for minimum signal voltage as indicated by the vertical height of the oscilloscope trace. Then transfer the oscilloscope probe to the chroma detector output circuit and adjust any remaining 47.25-mc traps for minimum signal output.

11-1233. You should now adjust the rf signal generator to 41.25 megacycles. Observe the schematic diagram of the television receiver, locating those plate or grid circuits containing two wave traps that are very close in frequency. You will reduce interaction between these traps if you connect a carbon swamping resistor of a low value across the 39.5-mc trap coil. Adjust the 41.25-mc traps for minimum signal output as indicated by the oscilloscope. Now you should transfer the oscilloscope probe from the chroma detector output to the video detector output and adjust any remaining 41.25-mc traps.

11-1234. Transfer the swamping resistor from the 39.5-mc traps to the 41.25-mc trap coils. Adjust your signal generator to 39.5 megacycles. Tune the 39.5-mc traps

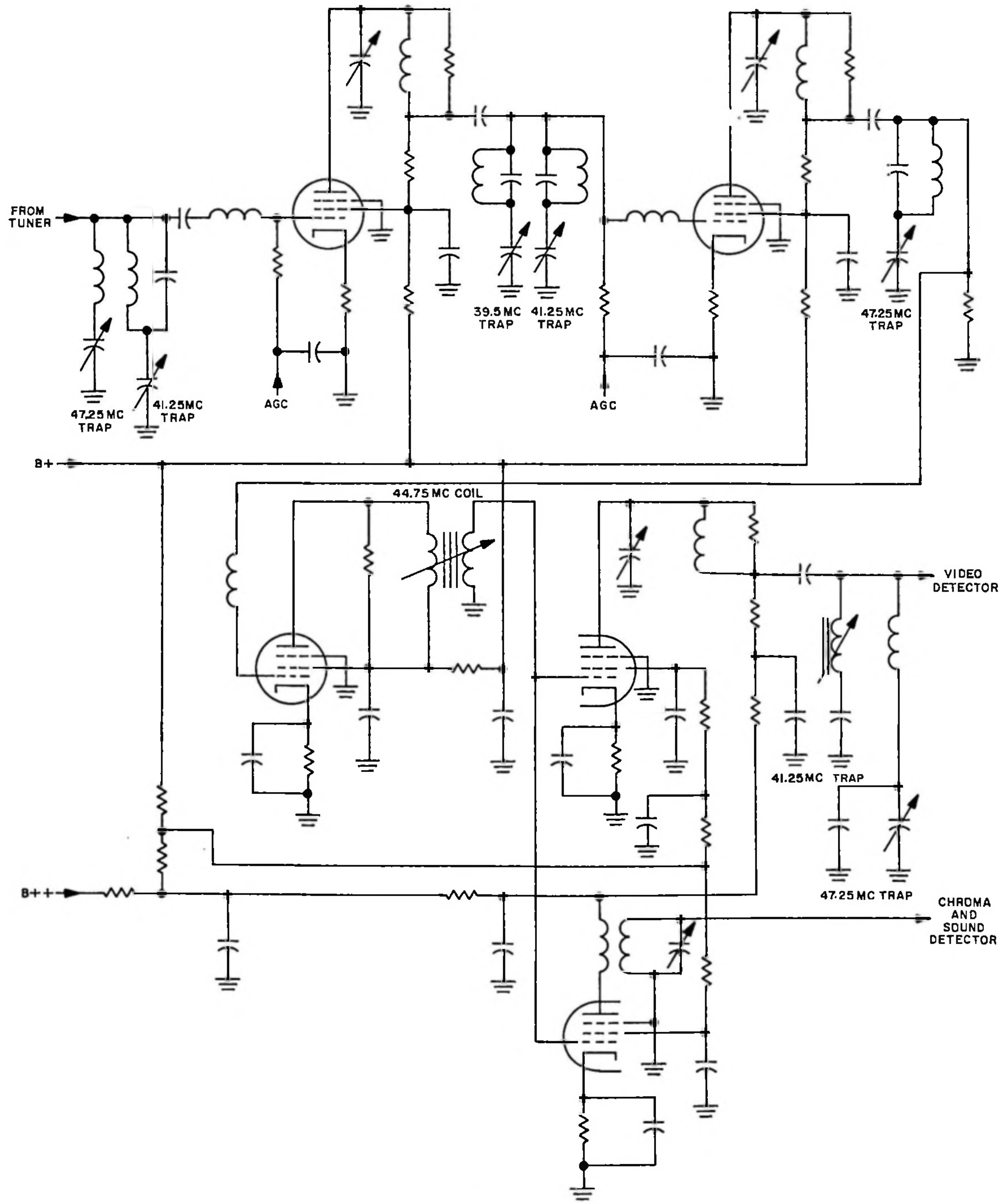


Figure 11-178. Color Television Video I-F Stages

for minimum signal voltage as indicated on the oscilloscope. Transfer the isolation resistor from the video detector output circuit to the chroma detector output circuit and tune any remaining 39.5-mc traps to minimum output.

11-1235. Adjust the radio frequency signal generator to 44.75 megacycles, and tune the coil or pole settings for maximum output voltage as indicated by the oscilloscope display.

11-1236. FINE TUNING CONTROL. Disconnect the rf signal generator from the television set. Connect the output of the antenna impedance matching network to the receiver antenna terminals, and connect the rf signal generator to the input of the antenna matching network. Adjust the rf signal generator to 65.75 megacycles; then turn the television channel selector to channel 4. Connect the oscilloscope and the 10k isolation resistor between the video detector output circuit and ground, as shown in figure 11-179. You should adjust the receiver fine tuning control, inspecting the oscilloscope display for minimum lower-adjacent-channel sound signal. This step ensures that the marker signals in the following steps will be converted to the correct intermediate frequencies. Remove the rf signal generator from the antenna matching network, and connect the generator, through a 1- or 2-picofarad capacitor, between one antenna terminal and ground, as shown in figure 11-176. Turn the modulation control off so that an unmodulated rf signal is obtained.

11-1237. OVER-ALL RESPONSE CURVES. Connect a sweep generator to the input of the antenna matching network. Next, connect the low-impedance detector, shown in figure 11-180, between the last video i-f amplifier and the vertical input terminals of an oscilloscope. Turn the oscilloscope sweep control to the horizontal amplifier position, and connect the oscilloscope horizontal input terminals between ground and the phase terminal of the sweep generator. The test equipment setup will be similar to that illustrated in figure 11-176, except that the signal applied to the oscilloscope is obtained from the last video amplifier instead of the mixer test point.

11-1238. Set the sweep generator frequency control to 195 megacycles, and the sweep control to 10 megacycles. Adjust the rf signal generator to 194.75 megacycles. Adjust all the trimmer capacitors, except those located in the trap circuits, for an over-all response curve as shown in figure 11-181. Note the position of the video carrier (45.75 mc) on the response curve. Set the marker generator to 196.5 megacycles and observe that the marker moves near the middle of the response curve at 44 megacycles; the amplitude of the response curve should be greatest near this point. Set the marker generator to 198.5 megacycles and observe that the marker moves over the response curve to a region very close to the edge of the curve at 42 megacycles.

11-1239. Set the marker generator to 240.5 megacycles and transfer the oscilloscope probe to the video detector output circuit. Locate the video amplifier stage preceding the video detector, and tune its resonant circuits to obtain the over-all response curve shown in figure 11-182. Set the marker generator to 195.5 megacycles and notice that the marker has moved up the response curve slope almost to the flat-topped portion at 45 megacycles. Set the marker generator to 197.5 megacycles and observe the marker travel across the response curve to a point which is slightly below the flat-topped portion of the curve at 43 megacycles.

11-1240. Transfer the oscilloscope probe to the chroma detector output circuit; then return the marker generator to 240.5 megacycles. Locate the video amplifier

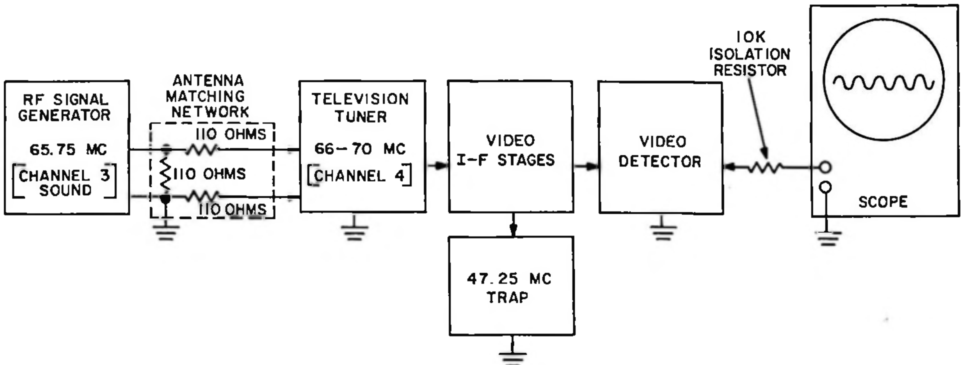


Figure 11-179. Centering the Fine Tuning Control, Block Diagram

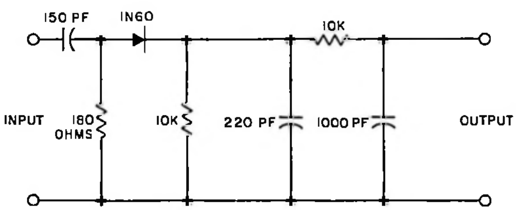


Figure 11-180. Low Impedance Detector

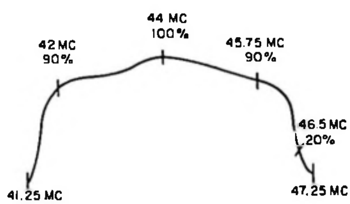


Figure 11-181. Over-all Response, Last Video I-F Output

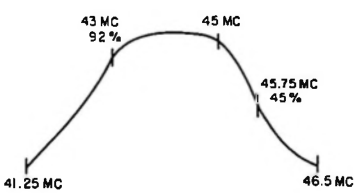


Figure 11-182. Over-all Response, Video Detector Output

preceding the chroma detector and tune its resonant circuits to produce the response curve shown in figure 11-183. Set the marker generator to 195.5 megacycles, observing the marker travel up the slope of the response curve to the point of maximum amplitude at 45 megacycles. Set the marker generator to 198.5 megacycles and observe that the marker moves to the other edge of the response curve at 42 megacycles. Disconnect all test equipment because the alignment of the video i-f stages is completed.

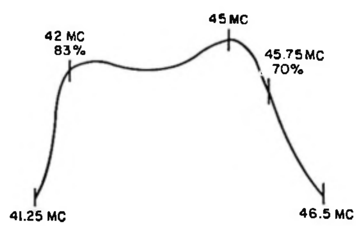


Figure 11-183. Over-all Response, Chroma and Sound Detector Output

11-1241. SOUND CHANNEL ALIGNMENT.

11-1242. The procedures for the alignment of the sound channel of a television receiver are the same as those given in paragraphs 11-202 (for the Foster-Seeley circuit) and 11-207 (for the ratio detector circuit).

11-1243. CHROMA CHANNEL ALIGNMENT.

11-1244. CHROMA AMPLIFIER. Remove the 3.58-mc oscillator tube and the horizontal output tube from the receiver. Then connect the high impedance detector, shown in figure 11-184, between the vertical oscilloscope terminals and the R-Y demodulator. Connect a 10k series resistor between the a-m rf signal generator and the input terminal of the chroma amplifier channel.

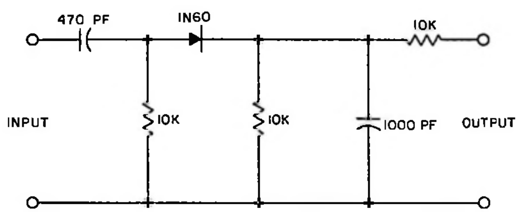


Figure 11-184. High Impedance Detector

11-1245. Disconnect the agc bus line at its source, and substitute the bias supply shown in figure 11-175. Advance the bias supply control until the video intermediate frequency amplifiers are cut off. Adjust the frequency of the signal generator to 4.5 megacycles. Adjust the 4.5-megacycles sound trap in the output of the chroma amplifier for a minimum oscilloscope indication. Change the signal generator frequency to 3.58 megacycles and tune all the chroma coils for maximum oscilloscope indication. Disconnect the a-m signal generator from the input of the chroma amplifier.

11-1246. You should now connect a sweep generator to the input terminals of the chroma amplifier. Set the center frequency to 3.58 megacycles, and then set the sweep control of the generator to give a sweep that is approximately 3 megacycles wide. Loosely couple an rf signal generator to the sweep generator, to provide marker pips at the frequencies shown in figure 11-185. Now adjust the chroma output inductors for a response curve similar to that shown in the figure. When adjusting for the required response curve, do not adjust the output transformer of the chroma amplifier any more than necessary. Adjustments in excess of a quarter-turn can produce serious phase shift. The dotted lines in the figure show the variations in the response curve as the chroma control is varied. The test equipment is now removed because the chroma channel alignment is completed.

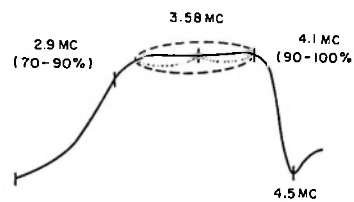


Figure 11-185. Chroma Output Signal, Over-all Response

11-1247. BURST AMPLIFIER. Connect a color bar generator such as the Philco Model 7100, to the antenna terminals of the television receiver. Set the channel selector to channel 3 or 4, depending on the generator crystal frequency. Set the OUTPUT SELECTOR to the R.F. CARRIER position, and then rotate the PATTERN SELECTOR to the COLOR BARS position. Adjust the R.F. OUTPUT CONTROL to its mid-range position. Tune the fine tuning control of the television receiver to eliminate the 900-kc beat note. Set the hue control on the receiver to approximately its mid-range position.

11-1248. Set the function control of a vacuum tube voltmeter to the -dc position and the range switch to the 50-volt range. Connect the voltmeter between ground and the diode plate on the output side of the phase detector, at point A; see figure 11-186. Adjust the burst transformer for maximum indication on the voltmeter.

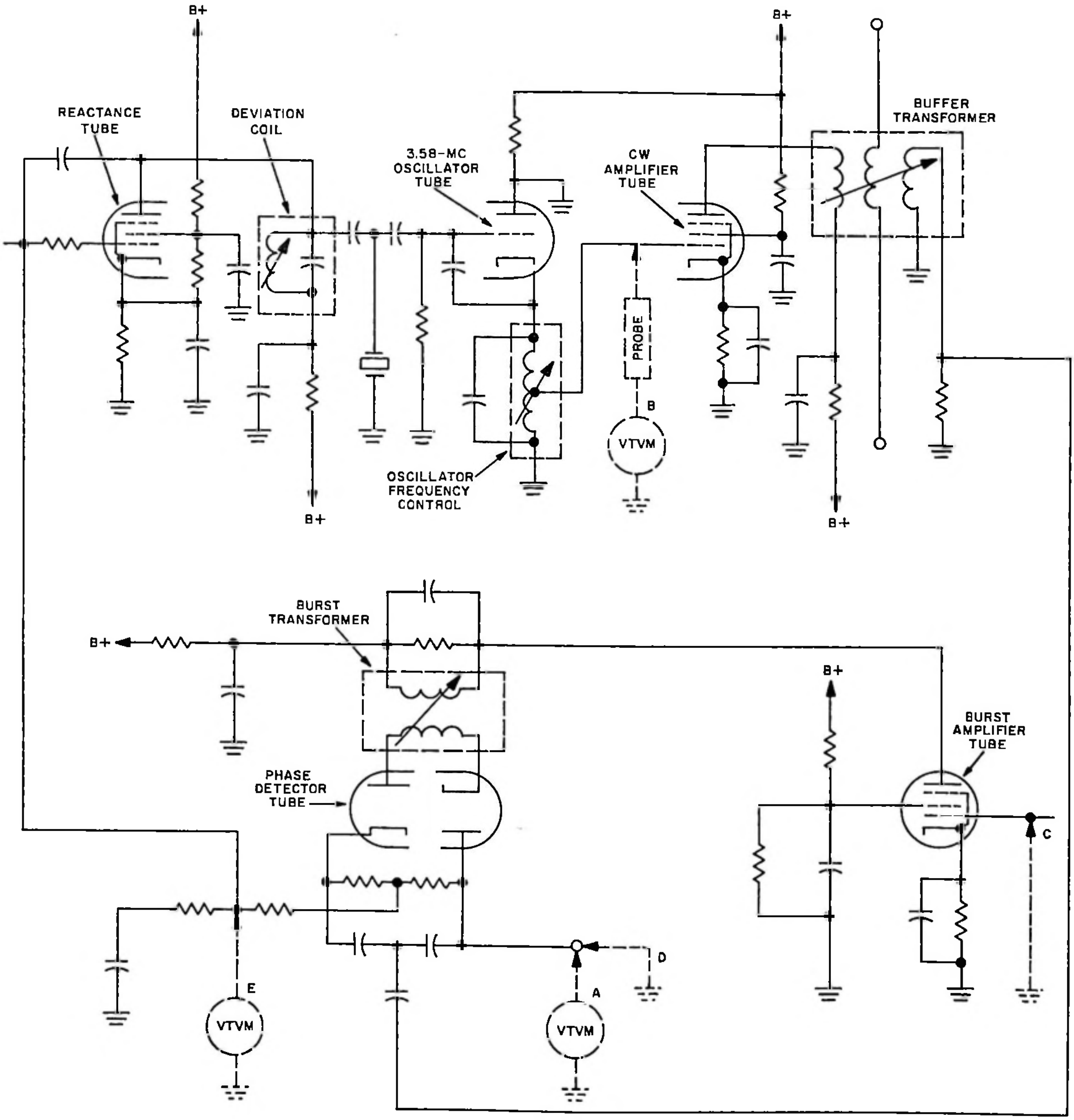


Figure 11-186. Color Television Video Stages, Schematic Diagram, Showing Temporary Connections and Test Points

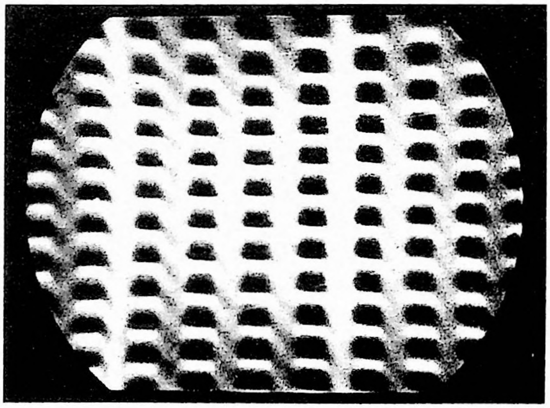


Figure 11-187. Receiver Color Oscillator Out of Synchronization

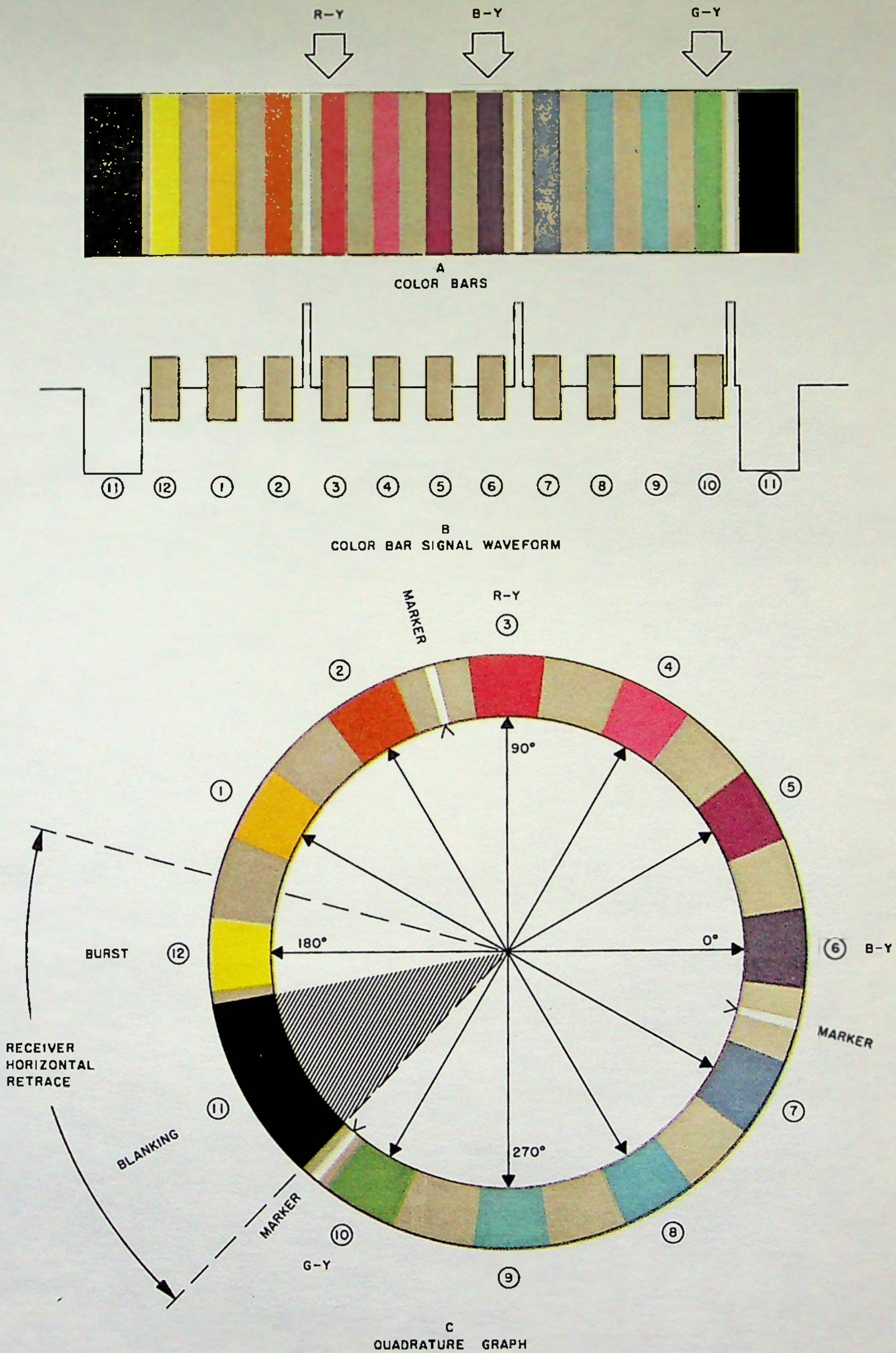


Figure 11-188. Color Bar Signal Relationships

11-1249. 3.5-MC OSCILLATOR. Connect an rf probe between the electronic voltmeter and the control grid of the cw amplifier, at point B in figure 11-186. Turn the tuning slug of the deviation coil, located in the plate circuit of the reactance modulator, all the way to its top position. Notice that the color bars lose synchronization and appear as shown in figure 11-187. Reverse the direction of rotation, screwing the slug into the coil about one-quarter of the way.

11-1250. Turn the tuning slug of the oscillator frequency control all the way to the top, noting that the oscillator ceases to function. Reverse the direction of rotation, screwing the slug into the coil until the voltmeter indicates maximum deflection. This shows that the oscillator is functioning. Continue adjusting the tuning slug one or two more turns beyond this oscillation point.

11-1251. If the grid of the burst gate or its clamp is jumpered to ground at point C of figure 11-186, observe that the color bars are replaced by black-and-white bars. Jumper the color-killer voltage to ground at point D of the figure, where it was injected into the grid circuit of the chroma amplifier. The color should now be restored to the color bars. As you adjust the deviation coil in the plate circuit of the reactance tube, inspect the color pattern. Stop the adjustment when the color bars stop running; however, they will not remain stationary very long. Remove the jumper from the color-killer bus line.

11-1252. CW AMPLIFIER ADJUSTMENT. Set the function control of an electronic voltmeter to −dc position and the range switch to 50-volt position. Connect the voltmeter between the output plate of the phase detector and ground, as shown at point A of figure 11-186. Adjust the voltmeter range control for a mid-scale indication, and adjust the buffer transformer of the cw amplifier for maximum meter indication. Remove the temporary jumper from the grid of the burst gate clamp. Vary the R.F. OUTPUT CONTROL of the color bar generator to simulate strong and weak signal conditions. The color bars should remain in synchronization.

11-1253. DEMODULATOR ALIGNMENT.

11-1254. GENERAL. Demodulator adjustments are the final and most difficult electrical adjustment in the color television set. If you do not thoroughly understand oscilloscope waveforms, you will be seriously handicapped while performing these adjustments. The Philco Model 7100 Color Bar Generator output signal and its signal relationships are shown in figure 11-188.

11-1255. MASTER PHASE ADJUSTMENT. Connect the vertical amplifier channel of an oscilloscope to the red gun of the cathode-ray tube. If the red gun connections are not available at the tube base, then make your connection at the terminal strip tie points. The chroma signal will then be displayed as shown in figure 11-189. The position and amplitude of each one of these bars is important. You should also notice and identify the blanking signal. The blanking signal is always coincident with the zero reference line in the oscilloscope display, and is recognizable by a pronounced gap in the display. The next bar in the display is the burst signal; from this point, each bar is numbered as shown in the figure.

11-1256. At the grid of the red electron gun the waveform should be inspected for the following information:

a. The maximum amplitude signal should be color bar number 3, the red bar.

b. Color bars number 6 (blue) and number 12 (burst) should be located at zero reference level of the display. If not, adjust the burst transformer.

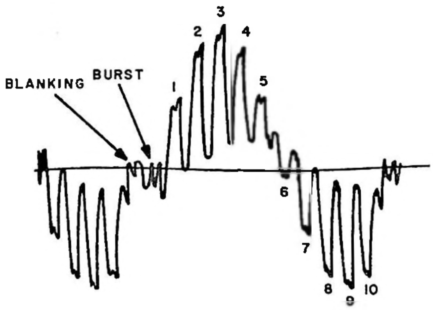


Figure 11-189. Demodulated Chroma Signal on Grid of Red Electron Gun

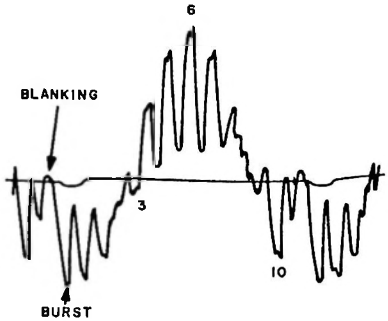


Figure 11-190. Demodulated Chroma Signal on Grid of Blue Electron Gun

c. Color bar number 9 (bluish-green) should be at the most negative point of the display.

d. The curve should be symmetrical with respect to the zero reference line.

11-1257. QUADRATURE ADJUSTMENT. Transfer the oscilloscope probe to the blue gun; the display will appear as shown in figure 11-190. Inspect the waveform for the following information:

a. Color bars number 3 (red), and number 9 (bluish-green) should be located at the zero reference line. If so, the proper quadrature is established; if not, adjust the phase-shift coil.

b. Color bar number 6 (blue) should have the maximum amplitude.

c. Color bar number 10 (green) is the first color bar in the negative direction.

d. Color bar number 12 (burst) should be the most negative bar in the display.

e. The curve should be symmetrical with respect to the zero reference line.

11-1258. PHASE DETECTOR BALANCE. Connect an electronic voltmeter at point E of figure 11-186. The voltage measured at this point should not exceed plus or minus 1/2 volt. Disconnect the color bar generator from the antenna terminals of the receiver. The noise balance, as read on the voltmeter, should not exceed plus or minus 0.8 volt. If the voltages you obtain are not within these ranges, a circuit failure is indicated.

11-1259. COLOR MATRIX ALIGNMENT. Those receivers which inject the luminance (Y) signal at the cathodes of all three electron guns, and inject the chrominance signals into the grids of their respective electron guns, do not employ matrix circuits. Therefore, this procedure is to be used only if a matrix section is included. In any event, before making this adjustment, align the demodulator stages.

11-1260. Following the demodulator stage alignment, perform the color matrix alignment. Connect the oscilloscope to the red gun. Adjust the master chroma and contrast controls to approximately their normal levels. Calibrate the oscilloscope for a 1-inch deflection from the top of the synchronization pulse to the base line. Readjust the master chroma gain control to give a peak-to-peak demodulated wave occupying a vertical amplitude of 1 inch, as shown in part A of figure 11-191.

11-1261. Transfer the oscilloscope connections to the blue electron gun. Recalibrate the oscilloscope for a 1-inch peak-to-peak deflection. Adjust the B-Y gain of the receiver for a 1.78-inch peak-to-peak deflection of the oscilloscope display, as shown in part B of figure 11-191.

11-1262. Transfer the oscilloscope connections to the green electron gun. Calibrate the oscilloscope for a 1-inch peak-to-peak deflection again. Adjust the G-Y gain of the receiver for a 0.63-inch peak-to-peak deflection of the oscilloscope display, as shown in part C of figure 11-191.

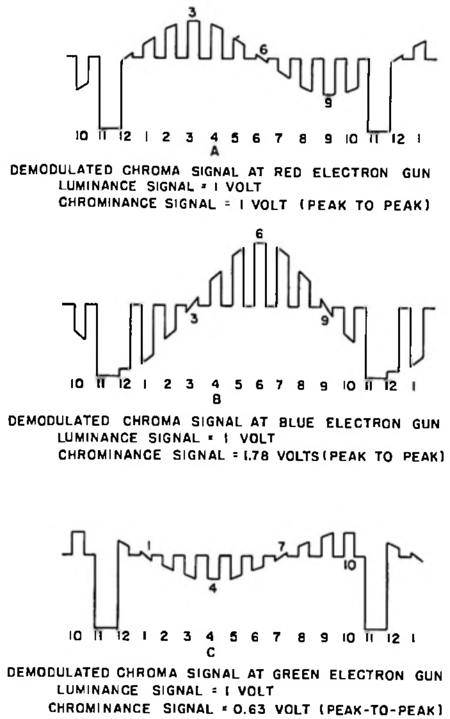


Figure 11-191. Typical Waveforms Observed at Respective Electron Guns

SECTION VI

RADIAC EQUIPMENT TESTING

11-1263. GENERAL.

11-1264. Radiac equipment is designed for the detection and measurement of radioactivity. The short designation RADIAC is derived from RAdioactivity Detection Identification And Computation. In general, the testing of radiac equipment can be accomplished with the aid of test equipment available to the electronics technician. Some components, such as subminiature tubes, Geiger-Mueller tubes, and resistors of very high values cannot be thoroughly checked with ordinary testing equipment, and therefore require special consideration. A brief outline of the basic types of radiac equipment is presented first. The test equipment necessary to check the special components encountered in radiac equipment and to calibrate some radiacmeters is then discussed.

11-1265. RADIAC FUNDAMENTALS.

11-1266. GENERAL.

11-1267. Radioactivity is the disintegration or breaking up of the atoms of an unstable element. Many chemical elements such as uranium, radium, radon, etc., have natural radioactive properties. These elements emit (radiate) specific kinds of particles and waves in various quantities and intensities, depending on the nature of the element from which the emission originates. These particles and waves are emitted without the addition of any external energy to the element.

11-1268. Small amounts of different elements are continually created by the disintegration of these radioactive elements. In turn, these different elements disintegrate into other elements. This decay process is not affected by such physical factors as pressure and temperature. The rate at which this decay occurs varies with each element. Thus, the half life of radium (the length of time elapsed before radium loses one-half its original activity) is approximately 1600 years; the half life of radon (a new element created by the decay of radium) is 3.82 days.

11-1269. Natural radioactive properties are shown usually by elements whose atomic weight is greater than that of lead (lead is the final stable element in the uranium series of disintegration). Under certain conditions normally stable elements can be made artificially radioactive. These elements then have their own characteristic half life of decay.

11-1270. EFFECTS OF RADIATION.

11-1271. The minute particles or waves emitted from a radioactive source are invisible. However, these particles or rays have certain observable effects as follows:

a. The radiations produce ionization in gases.

b. The radiation can penetrate materials that are opaque to ordinary light. The amount of penetration varies with each type of particle or wave.

c. Radiations cause certain types of material to fluoresce.

d. Materials that absorb radiation develop heat.

e. Damage of living tissues by radiations vary with the time and area of exposure.

Radiations can readily penetrate the human body, and if of sufficient amount and duration, can cause bodily injury or death. Various means have been devised to detect effects of radioactivity and to measure the amount or strength of the radiations over a given period of time.

11-1272. TYPES OF RADIATION.

11-1273. The principal emanations from radioactive materials are three types of rays (particles or waves): alpha particles (α), beta particles (β), and gamma waves (γ). Their characteristics and properties are important aids in their detection and measurement.

11-1274. ALPHA PARTICLES.

11-1275. Alpha particles are helium nuclei (helium ions with a double positive charge). They have velocities up to about 7 percent of the speed of light (3×10^{10} cm/sec). These particles have short range, poor penetrating power, and very strong ionizing power. The two protons which make up the helium nuclei give the particle a positive charge equal to twice the negative charge of an electron.

11-1276. BETA PARTICLES.

11-1277. Beta particles are simply high-speed electrons. The charge of a beta particle is therefore negative. These particles can move with a speed almost equal to the speed of light (about 95 percent). Beta particles have strong ionizing power (about 1/100 that of the alpha particle), and are able to produce measurable effects after passing through shields 100 times the thickness required to stop alpha particles.

11-1278. GAMMA WAVES.

11-1279. Gamma waves, or rays, constitute a type of electromagnetic radiation, similar to X rays, but in general have a much higher frequency of vibration and are far more penetrating. However, for the same wavelength of radiation the properties of the two types of rays are the same. Gamma rays are not particles and carry no charge. They are considered photons or bundles of electromagnetic energy similar to light waves. The wavelength of gamma rays is much shorter than that of light waves. In fact, this difference in wavelength is the main distinction between different types of electromagnetic radiation, including radio waves, radiant heat, infrared, visible light, ultraviolet, X rays, gamma rays, and cosmic rays. The sequence given is in order of decreasing wavelength and increasing penetrating power. Gamma rays have mild ionization power (about 1/10,000 that of the alpha particle) and are intensely penetrating (about 10,000 times that of the alpha particle). These gamma rays can be detected after passing through 12 inches of steel.

11-1280. UNIT OF RADIATION MEASUREMENT.

11-1281. The unit of measurement of radiation is called the "roentgen," or "r," and is defined as the amount of gamma radiation that will produce one electrostatic unit of charge in one cubic centimeter of air that is surrounded by an infinite mass of air at standard conditions. Radiation values are usually expressed in milliroentgens per hour (mr/hr). Human tolerance to radiation is usually defined in terms of these units. Radiation intensity decreases rapidly with an increase in distance from the radioactive object (approximately inversely as the square of the distance).

11-1282. RADIATION DETECTORS.

11-1283. IONIZATION CHAMBER METHOD. The ability of alpha, beta, and gamma radiation to ionize gases is the best characteristic most frequently used to detect the presence of radiation. The ionization chamber is used for collecting the ions formed by the ionization of air by radiation. A cross-sectional view of the chamber and the associated schematic are shown in figure 11-192. The chamber consists of two electrodes. The outer electrode (anode) is formed by an aquadag coating on the inner surface of the polystyrene walls. The inner electrode (cathode) is a rectangular loop of wire placed so that it is approximately 1-1/4 inches from the aquadag coating. The wire loop is insulated from the wall coating. The wall coating has an atomic number equal to about 7 or 8 (approximately the atomic number of air).

11-1284. The chamber is filled with air to a pressure of 760 millimeters of mercury (equal to the standard atmospheric pressure at sea level). The ionization chamber is sensitive to the effects of radiation. (Specially constructed chambers are used for alpha and beta or gamma radiation measurement.) When ionizing rays enter the chamber, they collide with the gas atoms in the chamber. These collisions release electrons from the gas atoms and the gas becomes ionized. Under the influence of the electric field maintained between the two electrodes, the positive and negative ions move to the cathode and anode, respectively. The movement of the ions in the chamber results in a minute current flow. (With no radiation present, the chamber acts as an open circuit, and no current can flow.)

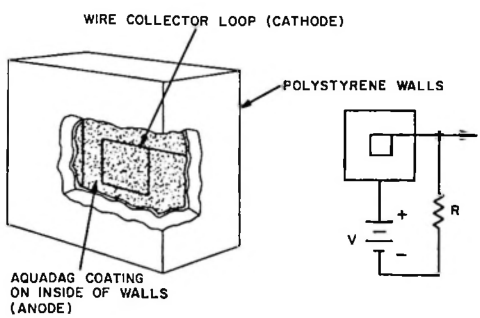


Figure 11-192. Ionization Chamber and Associated Circuit

11-1285. Potential V is applied across the chamber through resistor R (figure 11-192). Resistor R is a specially treated, high megohm resistance. The current flow through the ionization chamber is extremely small (a fraction of a microampere). This current flows through the extremely high resistance R and produces a potential difference that is fed to an amplifier.

11-1286. In an ionization chamber, the atoms of air are normally in a neutral state, and are not affected by any potential difference between the electrodes of the chamber. When ionized, they are influenced by this potential. Ionization consists of removing one or more electrons from an atom. This electron is now termed a negative ion. The electron is emitted with enough energy to force additional electrons from other atoms by collision. However, each time the electron collides with an atom it loses energy and is thus slowed down and left with a reduced ion creating capacity. If no potential is applied to the ionization chamber, electrons released by the original ionizing event would eventually be slowed down to a point where they would be captured by the positive ion. The charge of the positive ion would be neutralized and it would become a neutral atom.

11-1287. If a small voltage (for example 100 volts) is applied to the chamber, an electric field exists in the space between the positive aquadag coating on the walls and the negative collector loop. Positive ions formed by the ionization of the gas in the chamber tend to drift toward the negatively charged collector loop; negative ions (electrons) tend to drift toward the positively charged aquadag coating. Not all of the electrons reach the aquadag coating since there is a tendency for some of the ions to recombine and form neutral atoms. The electrons lost by this recombination do not contribute to the final current flow in the chamber. The longer the time that the electrons require to reach the anode, the greater is the possibility of recombination.

11-1288. If the applied voltage is increased to 160 volts, the electric field is sufficient to prevent recombination of the positive and negative ions. Under the influence of the field, the positive ions move toward the negatively charged collector loop of the ionization chamber, while all the negatively charged electrons move toward the positively charged aquadag coating. Thus, ionization within the chamber results in a current flow through the chamber.

11-1289. When the current in the chamber reaches the saturation value, further increases in applied voltage (within the limits of the ionization chamber region) do not increase the current flow (the gas amplification factor equals one). Thus, the chamber is operated over fairly wide limits of applied voltage without change in the current flow within the chamber. Variations in the ionization current result solely from changes in the intensity of the radiation.

11-1290. GEIGER-MUELLER TUBE METHOD. A simple device for such detection is a Geiger-Mueller (G-M) tube (figure 11-193). The tube is filled with a gas mixture at low pressure. A thin wire, the anode of the tube, is oriented axially to a cylinder and insulated from it. A voltage is impressed across the tube so that the wire is positive with respect to the cylinder. The magnitude of the impressed voltage is just below that necessary to ionize the gas molecules and cause conduction. In this dormant condition, no current flows. When radiation is present in the vicinity of the tube, an incoming radiation usually ionizes some molecules of the gas within the tube. The ionized gas particles are attracted toward either the cylinder or the wire, depending on their charge. On their way through the gas, these ionized gas particles collide with non-ionized gas molecules and ionize them. As a result, a large portion of the gas becomes ionized, thus producing a large current flow for only one initially created ion pair. Therefore the output from the tube is much greater per ionizing pulse than in the lower voltage ionization chamber. This current flow is quenched quickly, either by a small amount of organic vapor which is included in the gas mixture or by the use of external circuits which reduce the potential between the tube elements after conduction. As soon as tube conduction stops, the voltage across the tube is returned to the original pre-ignition value, and the tube awaits the next ionizing event. The duration of tube conduction is short compared to the average time between ionizing events and, therefore,the tube output is in the form of a series of pulses. Because of the fluctuating intensity of the ionizing radiations, the random time interval between ionizing events, and the chance arrangement of the gas molecules in the G-M tube, the pulses produced by the tube vary in amplitude (one half volt to 50 volts) and duration (50 to 100 microseconds), and occur at random time intervals. These pulses are generally used to activate various indicating devices.

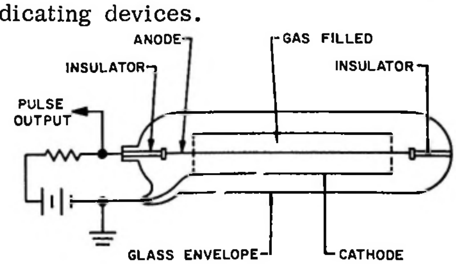


Figure 11-193. Typical Geiger-Mueller Tube

11-1291. TYPES OF RADIAC EQUIPMENT.

11-1292. GENERAL. The approved definitions of radiological terms are:

a. CHARGER, RADIAC DETECTOR. A device for providing an electrostatic charge to a radiac detector. May include means for measuring the amount of charge.

b. COMPUTER-INDICATOR, RADIAC. A device which performs the combined functions of computing and indicating radiac data.

c. COMPUTER, RADIAC. A device which receives information from a radiac detector and does one or more of the following: scales, integrates, or counts. Does not indicate.

d. DENSITOMETER. A device specifically designed to measure the optical density or opacity of material.

e. DETECTOR, RADIAC. A device that is sensitive to radioactivity or free nuclear particles and provides a reaction which can be interpreted or measured by various means.

f. INDICATOR, RADIAC. A device which displays radioactivity detection, identification, or computation information.

g. RADIACMETER. A device specifically designed to detect and indicate radioactivity. May, or may not, include radiac computer.

h. RADIAC SET. All the components and items required for a complete radioactivity detecting and measuring system.

i. TRANSMITTING SET, RADIAC DATA. All the components and items required to detect radioactivity and transmit radioactivity data as modulation on a carrier.

11-1293. DOSIMETER. A typical radiacmeter (dosimeter) of the Radiac-Dosimeter series IM-9/PD and Radiacmeter-Dosimeter IM-9C/PD series is shown in figure 11-194. Its function is to measure and indicate the accumulated dose of gamma radiation to which the wearer has been exposed. At one end of the radiacmeter is an optical eyepiece, and at the other end is the charging contact. The radiacmeter contains an ionization chamber into which is mounted a small electrometer. A scale calibrated from zero to 200 milliroengens is mounted in such a manner that the amount of radiation to which the wearer was exposed since the charging of the electrometer can be read directly by holding the radiacmeter up to a source of light and looking into the eyepiece. A radiacdetector charger is required to charge the electrometer. The dosimeter is four inches long and of tubular construction. It is provided with a clip similar to those used on pencils and pens, and may be worn by personnel in a similar manner.

Figure 11-194. Radiacmeter-Dosimeter IM-9C/PD

11-1294. RADIAC SETS. Two typical equipment series are Radiac Sets AN/PDR-18 and AN/PDR-27. These radiacmeters are portable, hand-carried equipments used to detect and measure the amount of radiation at a particular location. Both types are equipped with headphones, a calibrated meter, and a push-button-controlled meter illuminating light, and are supplied with power from internal batteries. The functional difference between these radiacmeters is in the degree of radiation they are designed to measure. The AN/PDR-27 series is referred to as a "low-intensity" meter, and is calibrated in milliroentgens per hour; the AN/PDR-18 series is referred to as a "high-intensity" meter, and is calibrated in roentgens per hour.

11-1295. RADIATION COUNTERS. For measurement of gamma radiation in counts per second, an equipment such as Computer-Indicator CP-79/UD may be employed. When used in conjunction with a radiac detector, the computer-indicator forms a counting system, such as the AN/UDR-9. For the measurement of gamma radiation in a sample of sea water, the detector is placed in the water and the information from the detector is fed to the computer-indicator, where the amount of radiation from the water sample is indicated in counts per second. The Computer-Indicator CP-79/UD employs a system of indicator lights mounted in six vertical columns, ten lights per column, to indicate visually the counting information. Another similar type of counting equipment is Radiac Set AN/UDR-3, which is employed for personnel radiation surveying. This equipment consists of several subsidiary units mounted in a panel. Both the hands and feet may be simultaneously monitored for gamma radiation by placing them in the openings provided. Six thin-walled, high-voltage Geiger-Mueller tubes provide the detection, while computation is furnished by five scale-of-eight counting circuits, augmented by electric timers and registers. The maximum counting rate is approximately 120,000 counts per minute.

11-1296. RADIAC TESTING.

11-1297. GENERAL

11-1298. The care and maintenance of radiac equipment entails very little special consideration other than the techniques normally used when testing electronic equipments. There are a few measurements and tests that require specialized test equipment, but the majority of testing can be accomplished with conventional test equipment. The calibration of radiac equipments should be checked at periodic intervals, and this requires a known source of radiation. This source is available only at radiac equipment repair shops.

11-1299. CALIBRATION.

11-1300. The calibration of radiation detecting equipments is beyond the capabilities of the common electronic test equipment available to the technician. Therefore, when equipment calibration is required, the services of the nearest radiac repair shop should be requested. The repair facility has equipment such as Radiac Calibrator Set AN/UDM-1, figure 11-195, by which accurate calibration of detecting equipment can be made. The AN/UDM-1 is fundamentally a source of radioactivity of known intensity. A lead chamber, mounted on a stationary table, contains a capsule of radioactive cobalt. The capsule is raised or lowered inside the chamber by an external control. When the capsule is lowered it is in the "safe" position; when raised, it is positioned in front of an opening in the wall of the chamber, and is in the "exposed" position. Mounted on rails directly in front of the source chamber is a movable table. The detecting device to be calibrated is mounted on this movable table, and the table is rolled along the rails until the proper distance between the radiation source capsule and the detecting device is obtained. Two attenuation plugs can be placed in front of the capsule, thereby permitting three degrees of radiation for one particular distance setting. A positioning chart and correction-factor table supplied with the equipment are used to determine the proper distance and attenuator control setting, depending upon the degree of radiation required to calibrate the particular detecting device.

11-1301. RADIOACTIVE TEST SAMPLE.

11-1302. Provided as a component of many low-intensity radiac test sets is a sample source of radiation. It is meant to be used to test the operation of the radiacmeter, and not as a source for calibration. An example of such a test source is Radioactive Test Sample MX-1083B/PDR-27, which is supplied as a component of Radiac Set AN/PDR-27F. The test source is employed as follows:

a. Energize the radiacmeter and hold the test source (by the clear, uncoated plastic end) perpendicular to, and half way between the ends, of the external Geiger-Mueller probe.

b. On the 0.5-milliroentgen scale, the meter should indicate saturation and return to zero as soon as the test source is moved a foot or two away from the probe. If the meter remains saturated when the test probe is removed, incorrect operation of the radiacmeter is indicated.

c. With the test source as in step a, switch to the 5-milliroentgen scale. The meter should indicate between 3 and 4 milliroentgens.

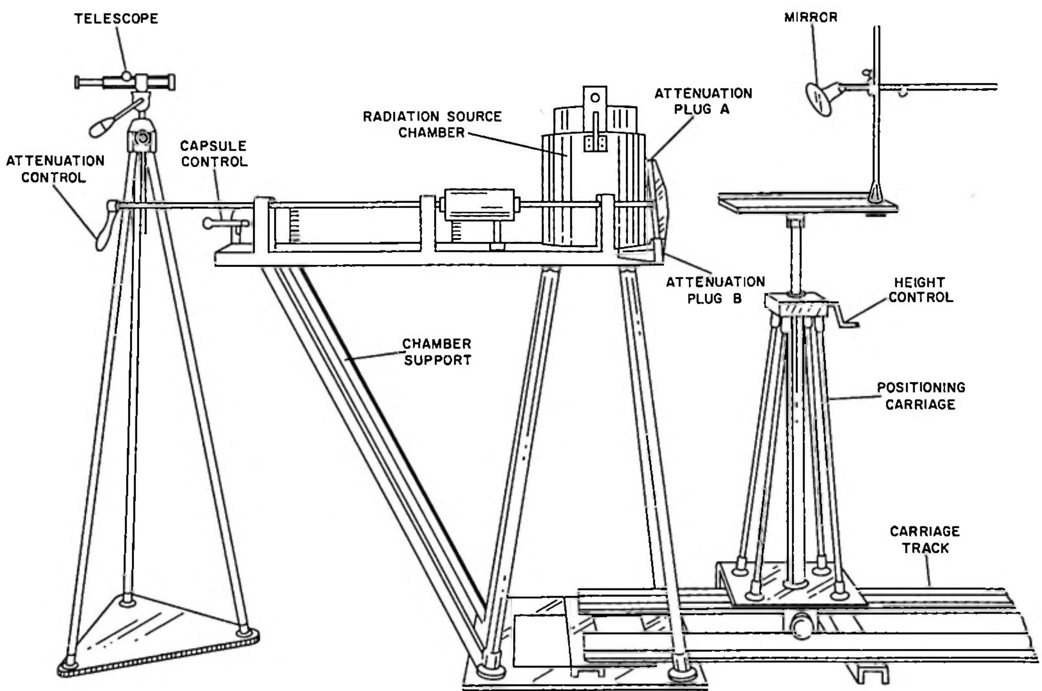


Figure 11-195. Radiac Calibration Set AN/UDM-1

d. Position the test source perpendicular to the G-M tube which is mounted inside the radiacmeter. On the 50-milliroentgen scale, the meter should indicate between 10 and 15 milliroentgens.

e. Switch to the 500-milliroentgen scale; the meter should indicate approximately 10 milliroentgens.

11-1303. HIGH RESISTANCE MEASUREMENTS.

11-1304. Some radiac equipments employ resistors with values up to a terohm (R x 10^{12} ohms). To measure these extremely high resistances requires specialized test equipment, such as Electron Tube Test Set TV-6/U. The TV-6/U is used to measure these resistors as follows:

a. Withdraw the drawer-type compartment on the front upper left-hand corner of the test set. Fold drawer front panel down.

CAUTION

To help prevent leakage paths due to dirt or grease, you should not handle the body of the resistor, hold only the leads at either end. If conditions demand that the body of the resistor be held, a tool such as tweezers should be used.

b. Place the resistor, figure 11-196, into the tube-resistor holder block so that one lead can be placed under the thumb nut of the Teflon insulated terminal, in the center of the drawer, marked high Z.

c. Connect the ohm clip (white) to the other end of the resistor. Return drawer securely in place. Set BAT-TEST switch to BAT position and check that batteries are operating properly. Set BAT-TEST switch to TEST position.

d. Set main selector switch to the approximate range scale required. The main selector switch has eight positions, R x 10^7 through and including R x 10^{14}. Place READ-SET switch to SET position. Rotate SET knob until needle on meter falls accurately on the meter scale line marked SET.

e. Set READ-SET switch to READ position. Adjust the PRECISION INDICATOR DIAL until the meter again falls on the SET mark. If the meter indication is too low with the maximum setting of the PRECISION INDICATOR DIAL, the main selector switch should be advanced to the next highest setting and the procedure repeated. Should it be impossible to make the meter fall back to the SET mark, the range being used is too high. The next lower scale should be used and the procedure repeated.

f. When the meter indicates properly, read the PRECISION INDICATOR DIAL and multiply the reading by the main selector switch setting. The inner dial of the precision indicator is calibrated in units of 1 to 10; the outer dial is calibrated in tenths and subdivided in hundredths. For example, a reading of 3.78 is typical.

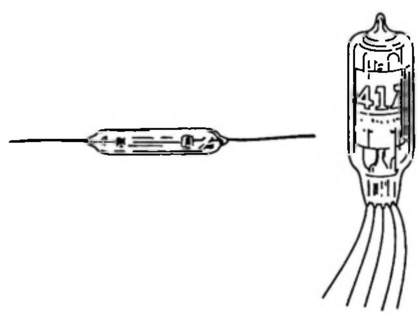


Figure 11-196. Subminiature Electrometer Tube and High Value Resistor

11-1305. TUBE TESTING.

11-1306. Because subminiature electrometer tubes, figure 11-196, and Geiger-Mueller tubes are used in many radiac equipments, tube testing presents a problem to the technician. To test these tubes, specialized tube testers such as Electron Tube Test Sets TV-6/U and AN/USM-23 (TV-9) have been developed. The TV-6/U is designed to test subminiature tubes. The tubes are placed in a compartment in the upper left-hand corner of the front panel. The connections to the tubes are made with a series of leads and clips, each clip having a different color, according to the element to which it is connected. Caution should be used when handling these subminiature tubes. Never hold the tubes near the base of the glass envelope; hold them by either the wire leads or the top of the glass envelope. This prevents grease or dirt from forming leakage paths. After the tube is properly placed in the tube holder and the leads connected, the correct operating voltage to each individual element of the tube is adjusted by means of the front panel controls. The operating voltages for a particular tube are found in the technical manual for the equipment from which the tube was removed, or in which it is to be used. The tube tester is capable of measuring both mutual conductance and plate current. The AN/USM-23 (TV-9) equipment is designed to test three specific types of Geiger-Mueller tubes, types BS-1, BS-2, and BS-101. The tubes are inserted in a socket designed to hold the type being measured; a chart attached to the tube tester furnishes detailed instructions for the testing procedure.

11-1306A. RADIATION MONITORING.

11-1306B. GENERAL.

11-1306C. The technique of radiation monitoring discussed herein is that of monitoring gamma radiation which can be caused by nuclear detonation. Radiation monitoring can be used following nuclear detonation in time of war to determine safe or uncontaminated areas, or for peaceful purposes such as determining radiation levels around construction areas where nuclear blasting has been used for excavation purposes. Figure 11-196A shows a typical arrangement used for monitoring a given area.

11-1306D. The radiation monitoring equipment is composed of a control console, located at a central command station, and ten or more remote sensor stations strategically placed in the area which is to be monitored for gamma radiation. In addition to the major equipments, ancillary equipments such as coaxial cable, fittings, batteries, and an external control circuit for use in remote sensor station testing are also provided.

11-1306E. CONTROL CONSOLE.

11-1306F. The control console is the heart of the radiation monitoring equipment in that it comprises the central piece of equipment which is used to turn on, to interrogate, and to turn off each remote sensor station employed in this facility. These functions enable the operator to determine directly the radiation levels at each point where a remote sensor station is located; figure 11-196B shows the operating controls and indicators on the control console.

11-1306G. The circuits which generate coded interrogation pulses sequentially transmit these pulses to the remote sensor stations to initiate the various operational cycles. Response pulses are transmitted back from the remote sensor stations and are then converted into useful data at the control console. Tone coding is incorporated so that each remote sensor station may be addressed and interrogated on an individual basis by means of a communications link (such as a single-pair wire-line loop or a

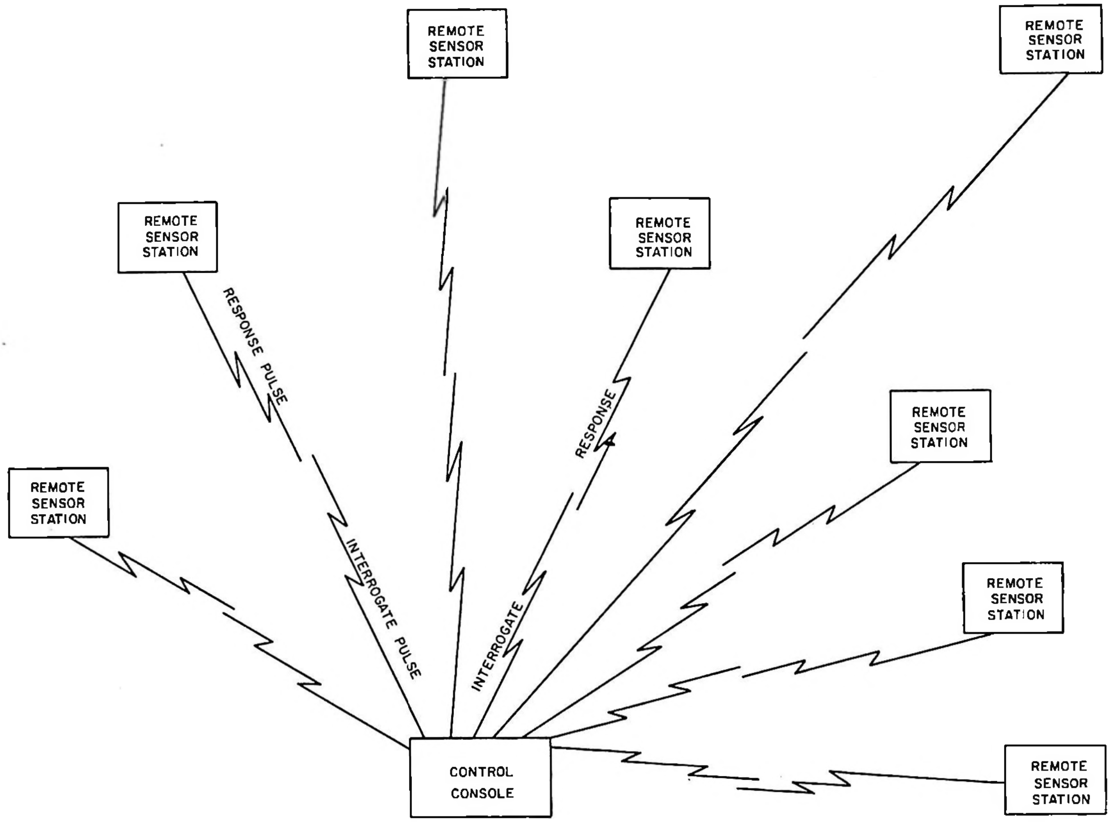


Figure 11-196A. Typical Arrangement for Monitoring a Given Area

single radio frequency) which is common to all stations. The interrogation pulses are composed of 200-millisecond bursts of 2 audio tones in coded combination. The circuits in the control console, which are used to generate the interrogation pulses, are capable of producing 50 unique combinations, each of which can be used to control a separate remote sensor station. In addition to the tone-signalling circuits, the control console also contains the circuits which provide a lighted indication revealing which remote sensor station is being interrogated, the step in the interrogation cycle being performed, the successful accomplishment of each step, and finally, a readout in approximate dose rate corresponding to the radiation level present at each remote sensor station.

11-1306H. REMOTE SENSOR STATION.

11-1306I. Each remote sensor station is composed of an ion chamber, an electrometer, a sensor logic unit, a radiation simulator, an interface circuit card, a transceiver, a power supply, telephone in-plant-type storage batteries, and separate receiver and transmitter antennas. Figure 11-196C illustrates the equipment installed in a typical remote sensor station.

11-1306J. The gas-filled ion chamber is encased in a protective metal housing which is provided with a mounting flange so that it can be attached to the threaded end of a three-inch galvanized-steel pipe for use in field installation. The electrometer, the sensor logic unit, the radiation simulator, the in-

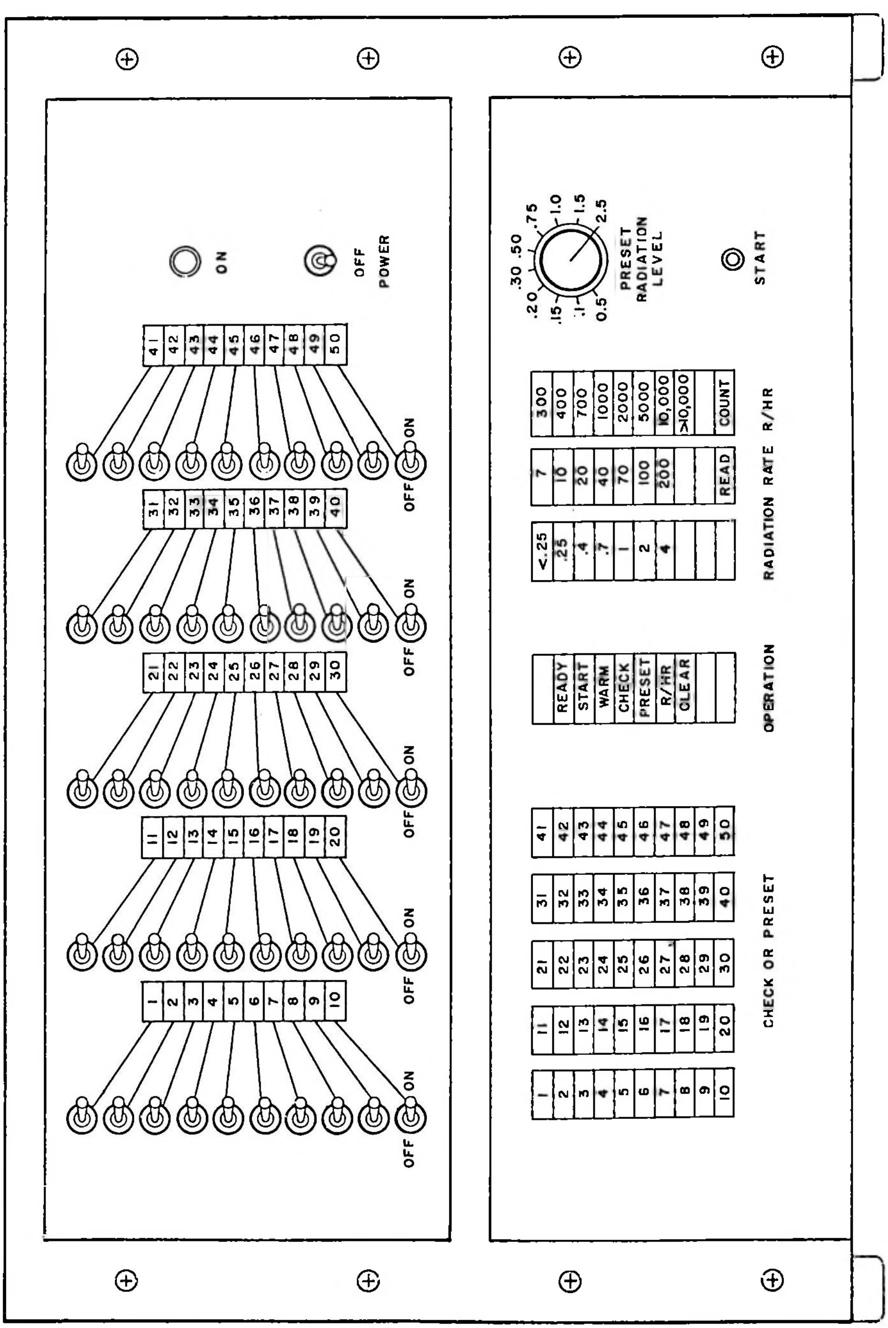


Figure 11-196B. Control Console, Front View terface circuit card, the transceiver, the power supply, and the storage batteries are located inside a weatherproof housing which is mounted on a utility pole (telephone, pow er, etc.). The separate receiver and transmitter antennas are located on top of this utility pole.

11-1306K. The coaxial cable which connects the ion chamber to the equipment in the weatherproof housing is installed in conduit and buried underground between the former and latter points. If necessary, the gas-filled ion chamber can be located as far as 300 feet from the equipment in the weatherproof housing without any appreciable loss of effectiveness.

11-1306L. ELECTROMETER. The battery-powered electrometer is a high-impedance amplifier which imposes a 600-volt dc potential on the electrodes of the ion chamber. A typical electrometer is shown in figure 11-196C.

11-1306M. When gamma radiation is present, the gas sealed in the ion chamber is ionized in proportion to the radiation level; consequently, a current proportional to the radiation level flows between the electrodes in the ion chamber. This current flow is used to charge a capacitive network in the electrometer. When a preset firing level is reached, the capacitive network discharges, thus generating the pulse which marks the time between the charge and the discharge of the capacitive network.

11-1306N. SENSOR LOGIC UNIT. The battery-powered sensor logic unit, shown in figure 11-196C, contains digital circuitry which receives and translates the 200-millisecond coded audio-tone pulses from the control console, controls the electrometer operation, and generates, at appropriate times, 2-millisecond pulses of 2-kHz tone in response to interrogation from the control console.

11-1306O. RADIATION SIMULATOR. The radiation simulator is designed to simulate the very-low current output of an ion chamber in a radiation field, and is intended for use in testing the radiation monitoring equipment and in conducting radiological defense exercises.

11-1306P. Realism is provided by several features incorporated in this device. For instance, fallout arrival time, after initial turn-on of all instruments in the equipment, is continuously variable from 10 to 60 minutes. Simulated peak-radiation values can be adjusted in 10-percent increments through a range from 25 milliroentgens per hour (mr/hr) to 15,000 roentgens per hr (r/hr). Multiple peaks may be programmed to simulate fallout from separate nuclear detonations, or to indicate redistribution of the contaminant. Decay characteristics of the radioactive material being simulated can be selected in accordance with the requirements of the exercise being programmed.

11-1306Q. Variation in the behavior of the fallout, such as fallout arrival, buildup, decay, and apparent additional fallout arrival, is accomplished by the use of eccentric cams. These cams are driven at the rate of one revolution per 8 hours. Eccentric cams can be prepared for virtually any fallout behavior pattern which is desired. The radiation curves shown in figure 11-196D illustrate the fallout behavior pattern effected by two different eccentric cams. Figure 11-196E shows an eccentric cam installed in the radiation simulator.

11-1306R. Prior to actual radiation-simulated operation, disconnect the ion chamber and connect the radiation simulator to the electrometer. The appropriate cam must be installed in the radiation simulator and the peak-radiation level must be selected in accordance with the fallout model being simulated. In addition, the cam must be rotated

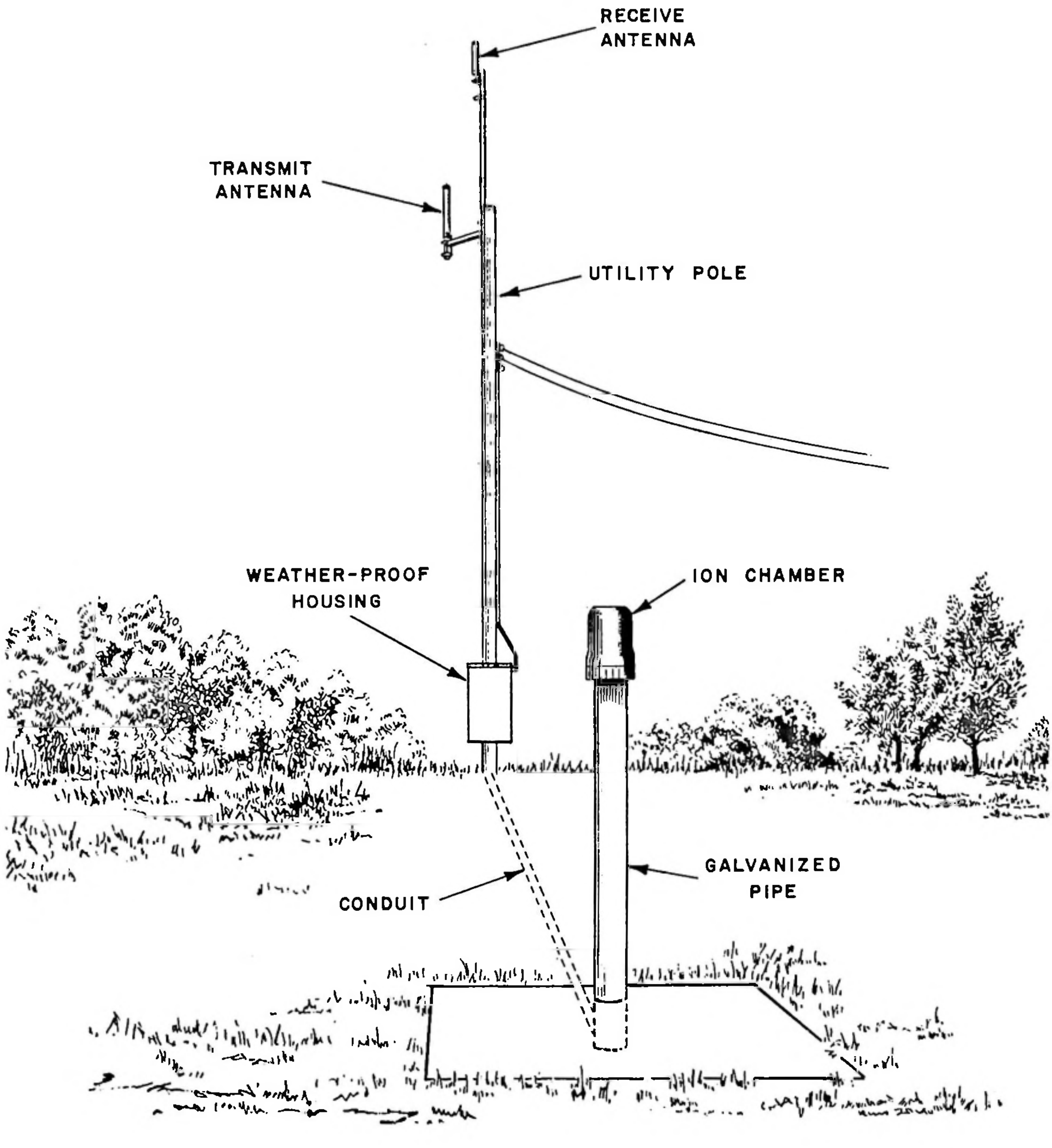


Figure 11-196C. Remote Sensor Station (Sheet 1 of 2)

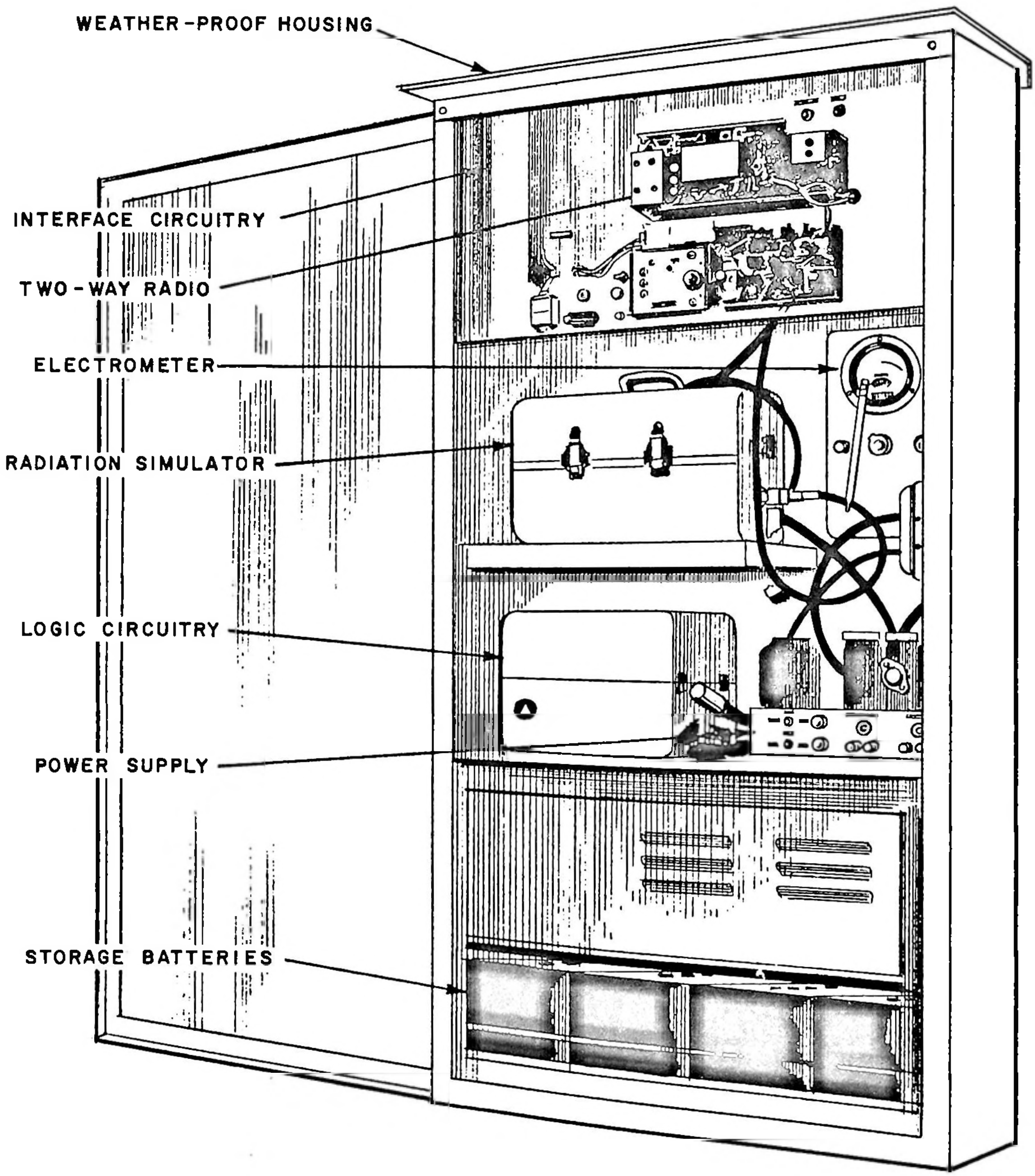


Figure 11-196C. Remote Sensor Station (Sheet 2 of 2)

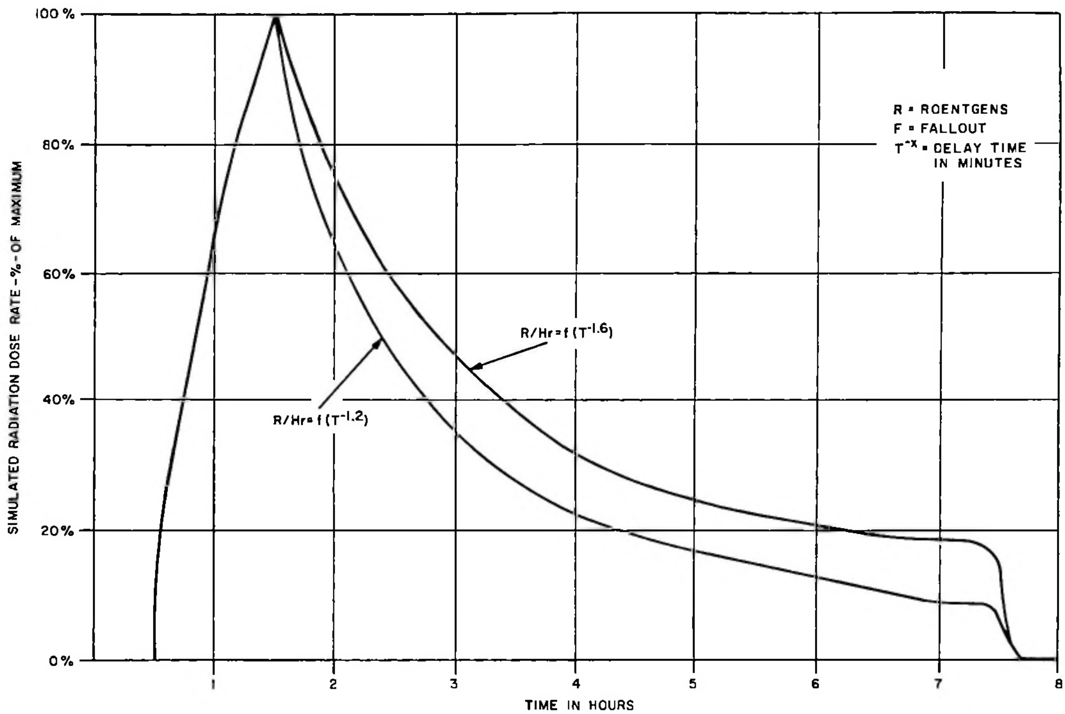


Figure 11-196D. Typical Radiation Curves Effected By a Radiation Simulator

on the shaft of the drive motor to set the desired fallout-delay time. This will govern the apparent time of fallout arrival at the remote sensor station after the start of the exercise. When the ac switch is turned on, the radiation simulator is not started as yet, but the remote sensor station has been readied for operation. All of these steps are manual and they must be done at the remote sensor station. Subsequent operation will be entirely automatic and the remote sensor station needs no further attention unless the exercise program requires changing or is completed. Referring to the block diagram (figure 11-196F), radiation simulator operation is as follows:

11-1306S. When the control console is activated at the beginning of the exercise, and when the remote sensor station is turned on for the first interrogation, the radiation simulator is energized by a signal generated in the sensor logic circuit. This signal energizes the ac power control which applies power to both the drive motor and the dc power supply. The drive motor turns the rotating cam assembly which in turn operates the servo potentiometer in accordance with the particular cam contour used. This contour is electro-mechanically mixed with the desired peak-radiation level and the output becomes a current variable with time over an 8-hour period. This current effectively simulates the desired roentgens-per-hour indication versus time curve at the electrometer input. At the end of the 8-hour cycle, the rotating cam assembly trips a switch in the ac power control which turns off the simulator at time zero. The entire 8-hour cycle may be repeated as desired by the remote turn-on function. If the equipment is to be returned to operational condition with the ion chambers connected, you must visit each remote sensor station and manually disconnect the simulators and reconnect the ion chambers.

11-1306T. The radiation-simulator schematic diagram, figure 11-196G, provides a reference for use in conjunction with the following detailed circuit analysis.

11-1306U. When the ac line switch S3 is closed, the dc power supply is energized, thereby supplying dc voltage to the fallout simulator. However, at this stage, voltage is not yet applied to the drive motor and the oven. The -9.1 volt regulated output of the dc power supply is divided by resistor R11 and VOLTAGE ADJUST potentiometer P1, thus resulting in an output voltage of -7.6 volts ±1%. This voltage is then applied to the peak r/hr increment selector, which, along with the contour generator potentiometer P_2, forms another voltage divider, whose output is a relatively-low voltage. The peak-r/hr decade selector is a set of precision resistors kept at a constant temperature in order to maintain stability. These resistors are connected in series with the radiation simulator output jack by the r/hr selector switch and they maintain a constant current source at the output jack.

11-1306V. When a set of relay contacts in the sensor closes, the ac voltage is applied

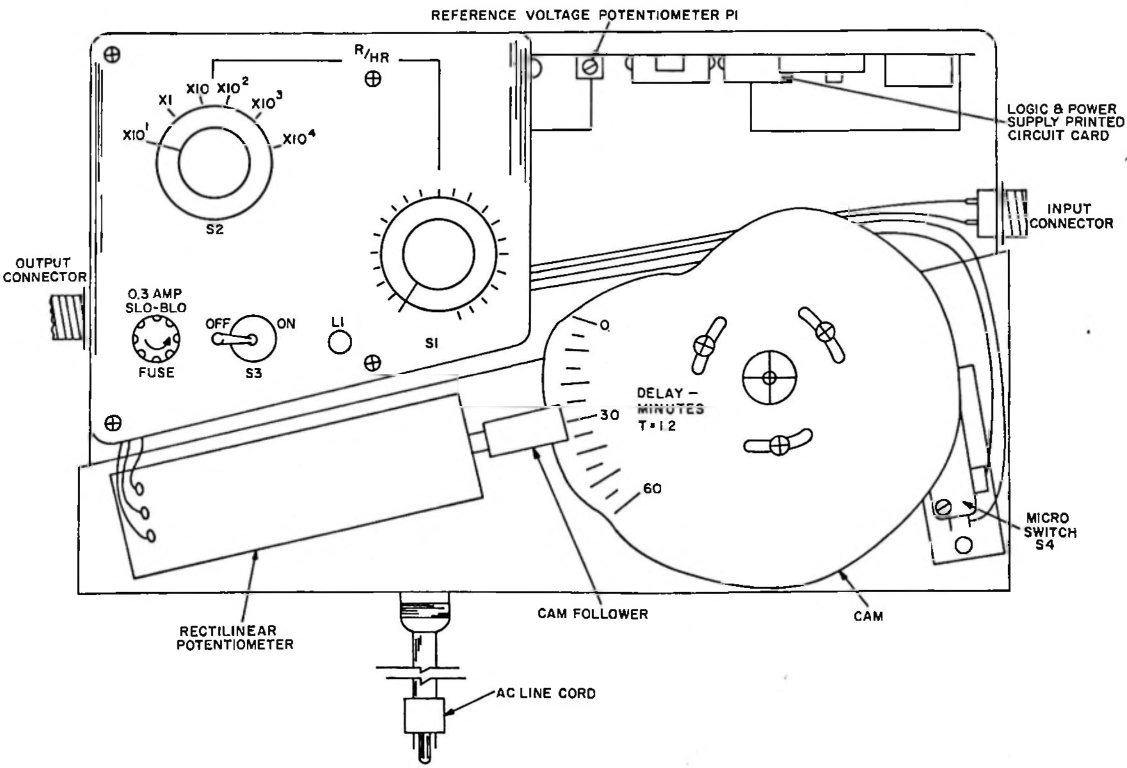


Figure 11-196E. Radiation Simulator, Top View

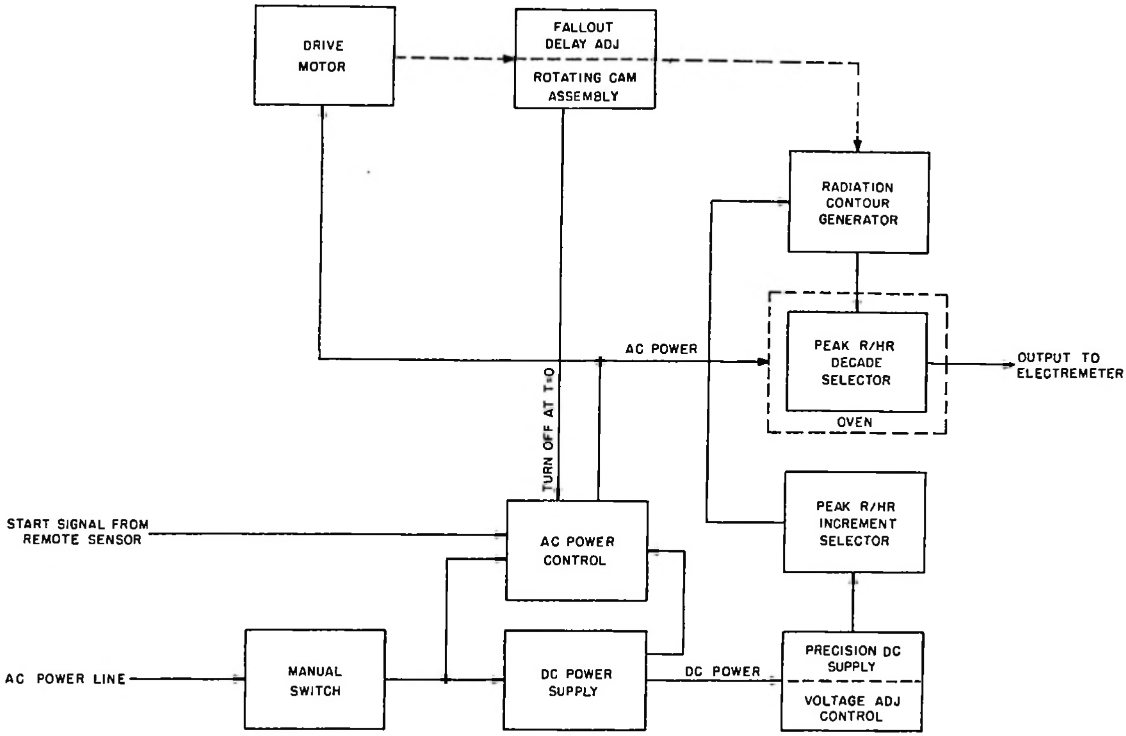


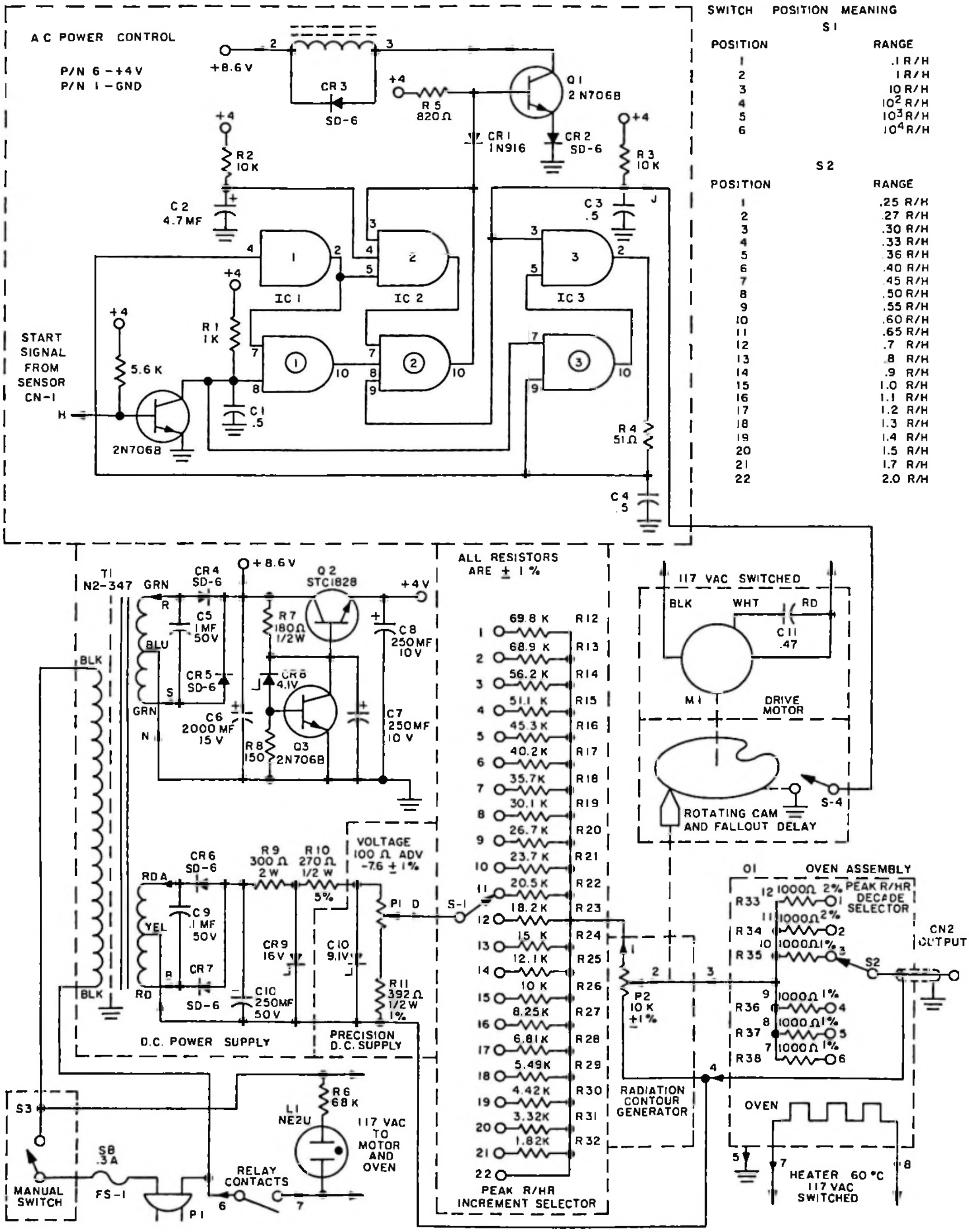
Figure 11-196F. Radiation Simulator, Block Diagram

to the motor and the oven, thereby activating them and starting cam rotation. The radiation contour potentiometer P_2 adjusts the voltage applied to the peak r/hr decade selector in accordance with the shape of the cam. Subsequently, the radiation-simulator output current is proportionately varied. At the end of 8 hours (one cam revolution) the cam-actuated switch S4 resets the ac power control, shutting off the motor and the oven, and enabling the remote start signal input.

11-1306W. The ac power control is composed of 3 dual 3-input micro-electronic integrated-circuit nand gates, 2 discrete component buffer stages, and one relay. The integrated circuits are arranged asynchronously to perform the required control. R2 and C2 form a delay circuit which guarantees the turn-on state. R4 and C4 form a feedback-delay circuit which eliminates a critical race hazard. Capacitors C1 and C3, together with resistors R1 and R3, form a network which feeds back enough voltage to the relay coil so that it will be locked up, thereby preventing the relay contacts from bouncing.

11-1306X. POWER SUPPLY. To provide the relatively-high stand-by current and high-peak current required by the remote sensor station, power to these remote sites is supplied from a 117-volt, regulated power supply to telephone-plant-type storage batteries. The power supply maintains the batteries at full charge and in the event of primary power failure, the secondary supply, represented by the batteries, can be used to operate the equipment for a period of one month.

11-1306Y. The schematic diagram of the



SWITCH POSITION MEANING

S 1

POSITION	RANGE
1	.1 R/H
2	1 R/H
3	10 R/H
4	10^2 R/H
5	10^3 R/H
6	10^4 R/H

S 2

POSITION	RANGE
1	.25 R/H
2	.27 R/H
3	.30 R/H
4	.33 R/H
5	.36 R/H
6	.40 R/H
7	.45 R/H
8	.50 R/H
9	.55 R/H
10	.60 R/H
11	.65 R/H
12	.7 R/H
13	.8 R/H
14	.9 R/H
15	1.0 R/H
16	1.1 R/H
17	1.2 R/H
18	1.3 R/H
19	1.4 R/H
20	1.5 R/H
21	1.7 R/H
22	2.0 R/H

Figure 11-196G. Radiation Simulator, Schematic Diagram remote sensor station power supply is shown in figure 11-196H. Two power supplies are involved; one at a nominal 12 volts, and one at a nominal 6 volts. The outputs of these power supplies are connected directly across three 6-volt storage batteries. These batteries are special long-life units and require exact charging voltages. If supplied with a floating charge from a source that is 6.45 volts per six-volt battery, the life expectancy of each battery is rated at 20 years. Twice-a-year maintenance is required to replenish the water level of the electrolyte solution. These regulated power supplies are standard series-regulated units with a current-limiting feature employed to limit the maximum charging current to 750 milliamperes. This is accomplished by diode clamping the collector of the constant-current supply transistor (Q1 or Q5) to the collector of the series regulator (Q4 or Q8) whenever the supply drain reaches 750 milliamperes.

11-1306Z. This power supply provides the sensor logic package with the necessary +6 volts and -12 volts (as referenced to sensor internal ground), and provides the transceiver with a 12-volt supply for all receiver circuits and a switched 12-volt supply to the oscillator and the audio stages of the transmitter. However, the remaining transmitter stages require 24 volts at 2 amperes when transmitting. A separate 24-volt power supply is provided for the transmitter. This power supply is located on the remote telemetry transceiver interface circuits card and is shown schematically in figure 11-196I.

11-1306AA. The 24-volt transmitter power supply is a small capacity dc-to-dc converter (Q10 and Q11) working into a very large (36,000 microfarad) capacitor. Because transmission times are limited to 10 milliseconds, this capacitor can supply the necessary 2 amperes to the transmitter for 10 milliseconds with a voltage loss of less than one volt. A Schmitt-trigger circuit (Q6 and Q7) followed by a switching network (Q8 and Q9) turns the converter off when the capacitor voltage reaches 29 volts, and back on again when the voltage drops below 20 volts. Initially, approximately 45 seconds are required to charge the capacitor from no-charge to the turn-off point. A diode clamps this supply to the switched 12-volt line so that during voice communication operations, when the transmitter is keyed for longer periods of time, the transmitter supply will not drop below 12 volts. Under these circumstances, operation is still possible at a somewhat-reduced power.

11-1306AB. SENSOR UNIT TEST SET. The sensor unit test set connects directly to the sensor unit and permits the sensor unit and its associated electrometer and radiation simulator to be tested independently of the control console. The test set is composed of a set of 11 twin-T oscillators whose frequency can be adjusted by a tone control potentiometer. The output signal of each oscillator is coupled to a driver which is nothing more than a single-stage amplifier with an adjustable output. One oscillator is used for each frequency in the equipment tone code. Each oscillator and driver can be selected by a tone selector switch to feed a common amplifier and output stage. The output is normally a 200-millisecond audio pulse at one or two selected frequencies to simulate the tone-coded interrogate pulses from the control console. An astable multivibrator, controlled by a pushbutton switch, provides the output pulse time. A test mode of operation, in which the selected tones appear continuously at the output, is provided for equipment calibration. Refer to figure 11-196J for a block diagram of the sensor unit test set.

11-1306AC. The output signals of the selected oscillators are coupled through driver circuits to the common amplifier. A shunt switch at the input of the common amplifier

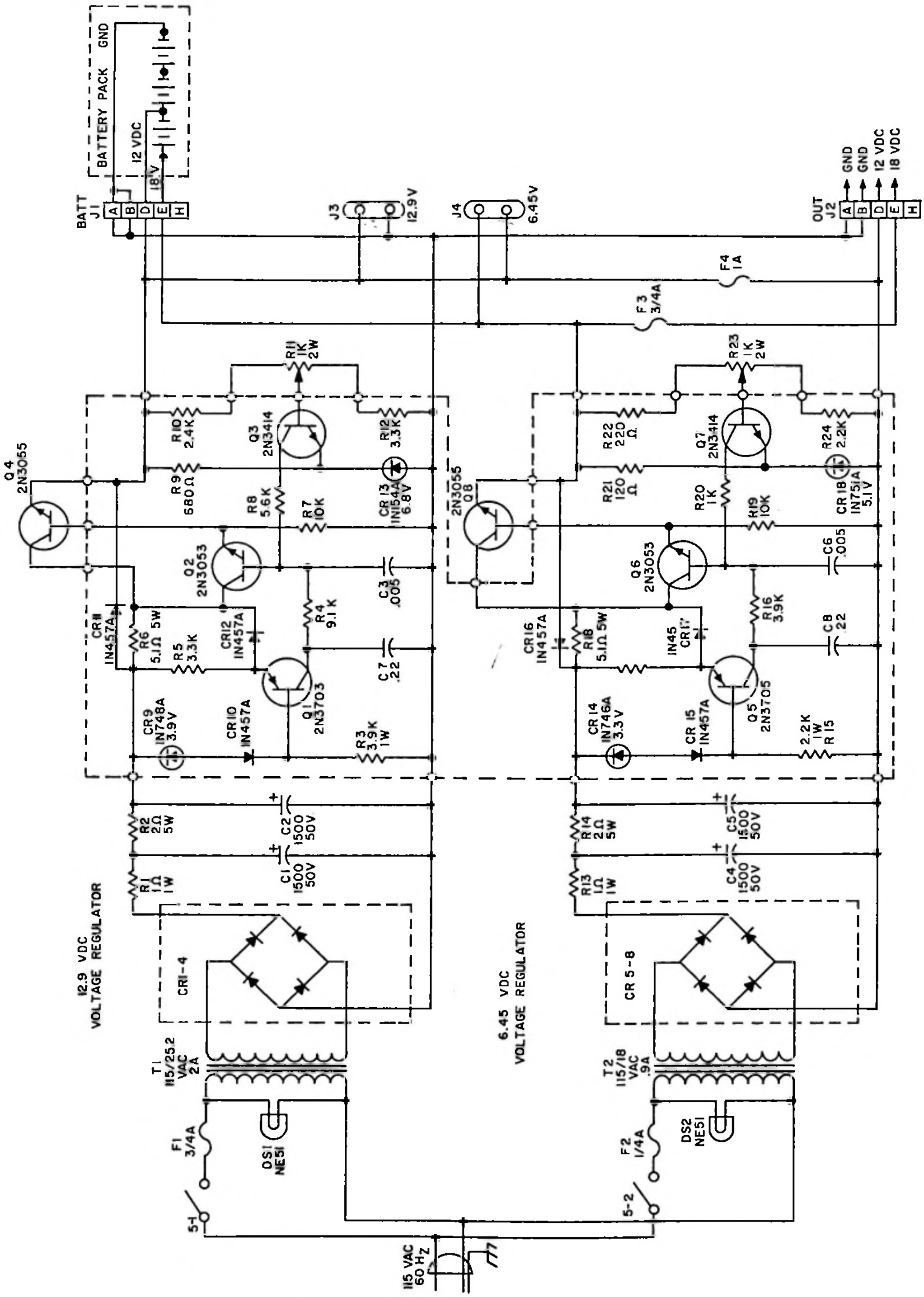


Figure 11-196H. Sensor Station Power Supply, Schematic Diagram

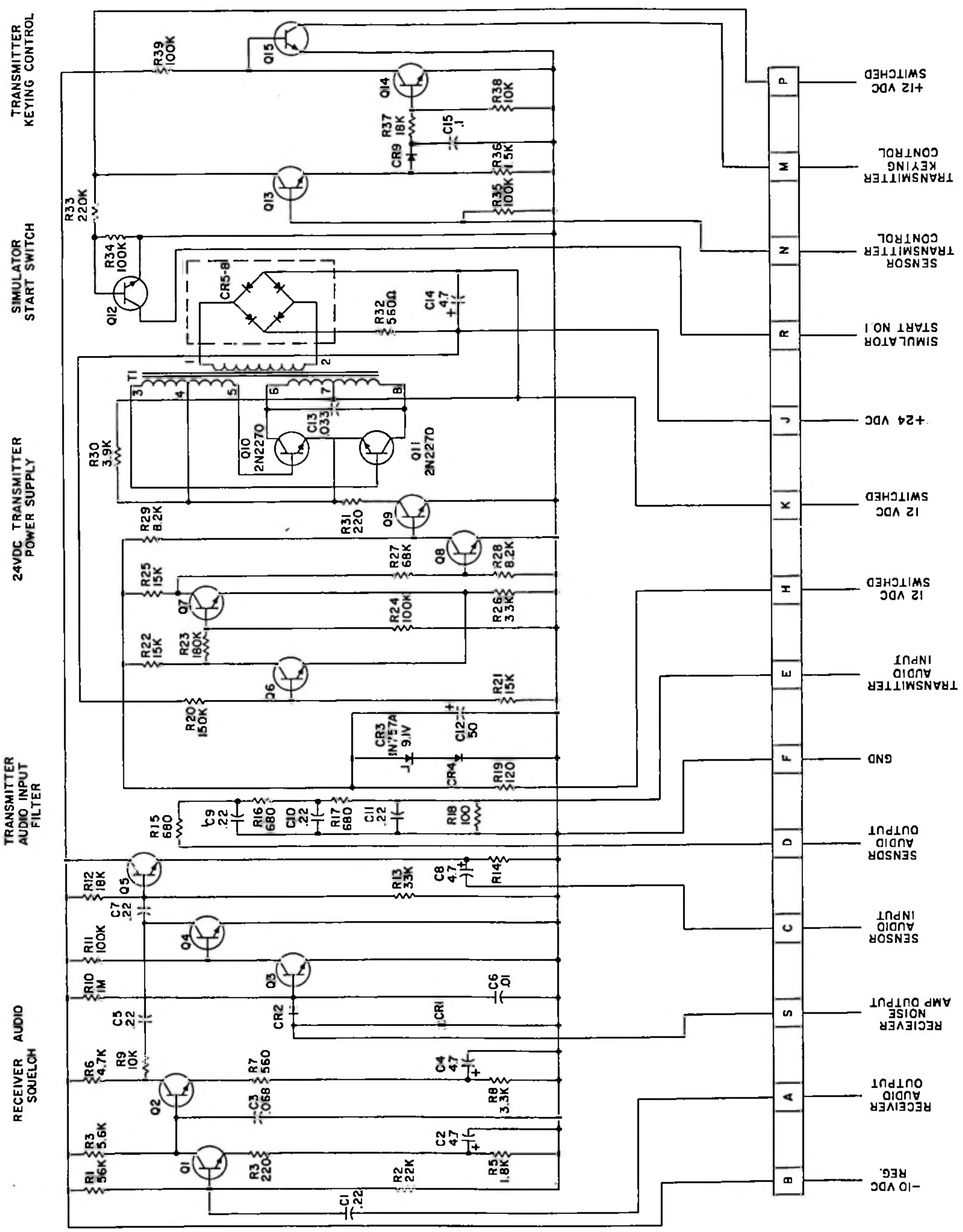


Figure 11-196I. Remote Transceiving Sensor Interface Circuit, Schematic Diagram

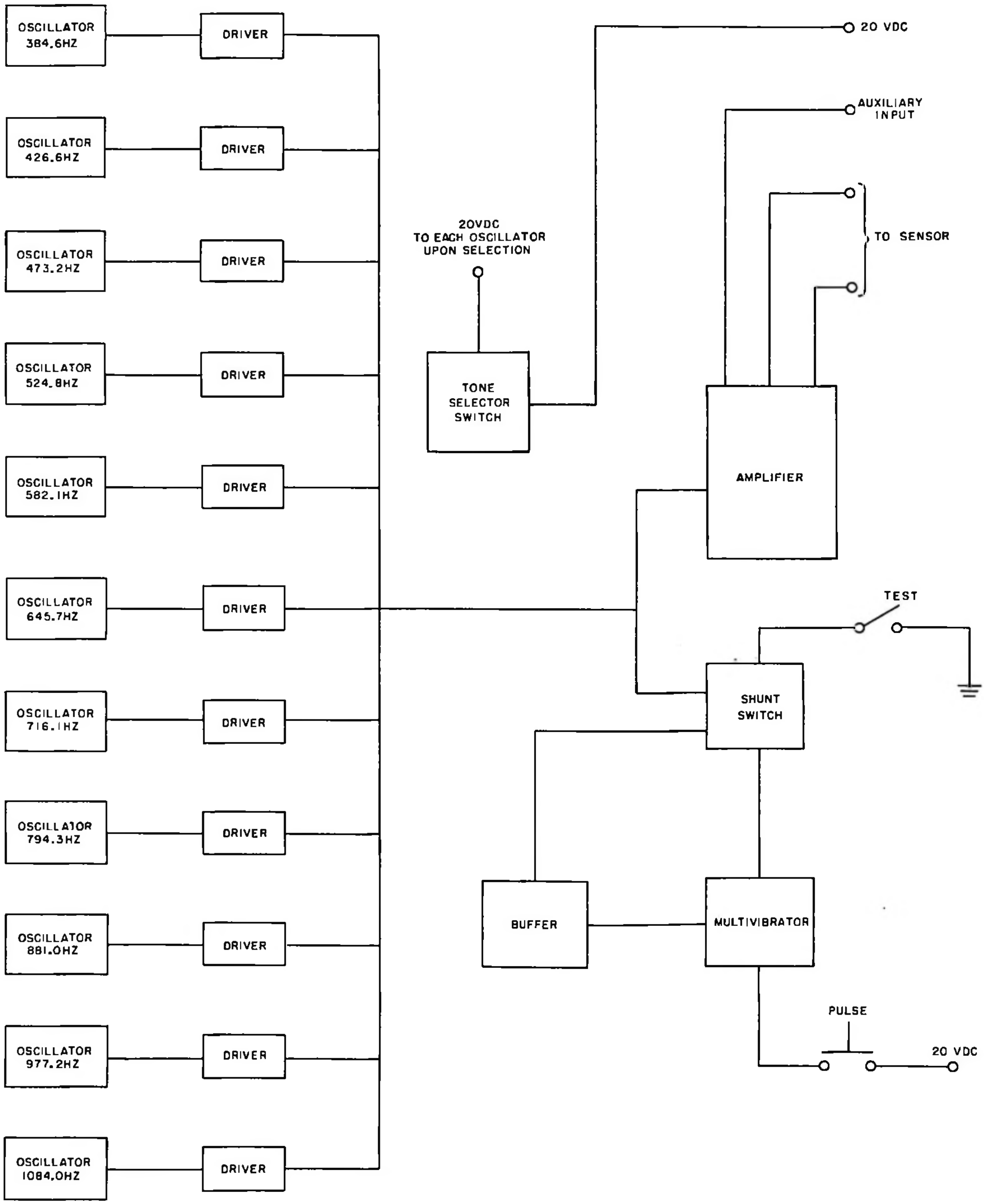


Figure 11-196J. Sensor Unit Test, Block Diagram provides output control. The shunt switch normally clamps the audio line to ground. When the 200-millisecond astable multivibrator is triggered by means of the pulse switch, the multivibrator output is coupled through a buffer stage to remove the base drive from, and to turn off, the shunt switch, thus allowing audio tones to pass for 200 milliseconds. The test switch grounds the base of the shunt switch and turns it off, thus permitting a continuous audio tone. Monitor points are provided for both the test set tone and the reply tone pulse from the sensor unit. All oscillator output signals are adjusted to a common level (approximately 0.5 volt) and applied to the common amplifier, the output of which is then adjusted to the desired input level for the sensor unit (nominally 0.5 volt). A standard series-regulated power supply is used to provide the 20 volts dc required for the test set circuits.

11-1306AD. RADIO TELEMETRY LINK.

11-1306AE. The development of a suitable radio telemetry link for this radiation monitoring method came from the consideration of radio-frequency spectrum space available to local governments. The critical demand for frequencies in the desirable vhf and uhf ranges by public safety users, broadcasters, businesses, land and air transportation, and others, plus the reservation of large blocks of frequencies by the Federal Government, means that the typical local government is able to obtain a few channels only for its use. These channels must serve all needs in the police, fire, public works, administration, and other radio-communication areas. Sharing a frequency already in use by a local government thus becomes a requirement for a practical radio telemetry link.

11-1306AF. The operation of this radio telemetry link is simplex; that is, transmitters and receivers are on the same channel and periods of transmission and reception for any unit must follow sequentially, rather than occurring simultaneously. In practice, the frequencies used by local governments occur in the 30-50 mHz, 150-174 mHz, and 450-470 mHz portions of the radio frequency spectrum. Frequency modulation is used and the channels are employed in regular daily service for voice communication and occasionally for tone signalling. Therefore, the radio equipment must be capable of stand-by operation as a single regularly-used channel without unduly interfering with the regular users, and without the regular traffic adversely affecting the reliability and accuracy of the radiation monitoring activity. In an emergency, it is assumed that the channel can be cleared of voice traffic for the time required to complete each cycle of data gathering.

11-1306AG. The operational requirements for a radio communications method are defined by the basic operating parameters of the radiation monitoring equipment. The frequencies involved are consistent with radio telemetry on voice-type channels. The interrogation tones fall between 286 Hz and 1184 Hz, while the fundamental frequency of the reply pulse is 2000 Hz; these frequencies are all well within the passband of commercial two-way fm equipment. Figure 11-196K is a block diagram of the radio telemetry link.

11-1306AH. TRANSMITTER AND INTERFACE — RADIO TELEMETRY BASE STATION. The interrogate tone pulse emitted by the control console is a 200-millisecond burst of 2 simultaneous tones in the 286-1084 Hz range. Amplitude is approximately 2 volts peak-to-peak across a 600-ohm load. The transmitter interface circuitry (reference figure 11-196L) at the radio telemetry base station passes this signal to the audio input terminals of the transmitter at the appropriate level for normal modulation. This circuitry also keys the transmitter and mutes

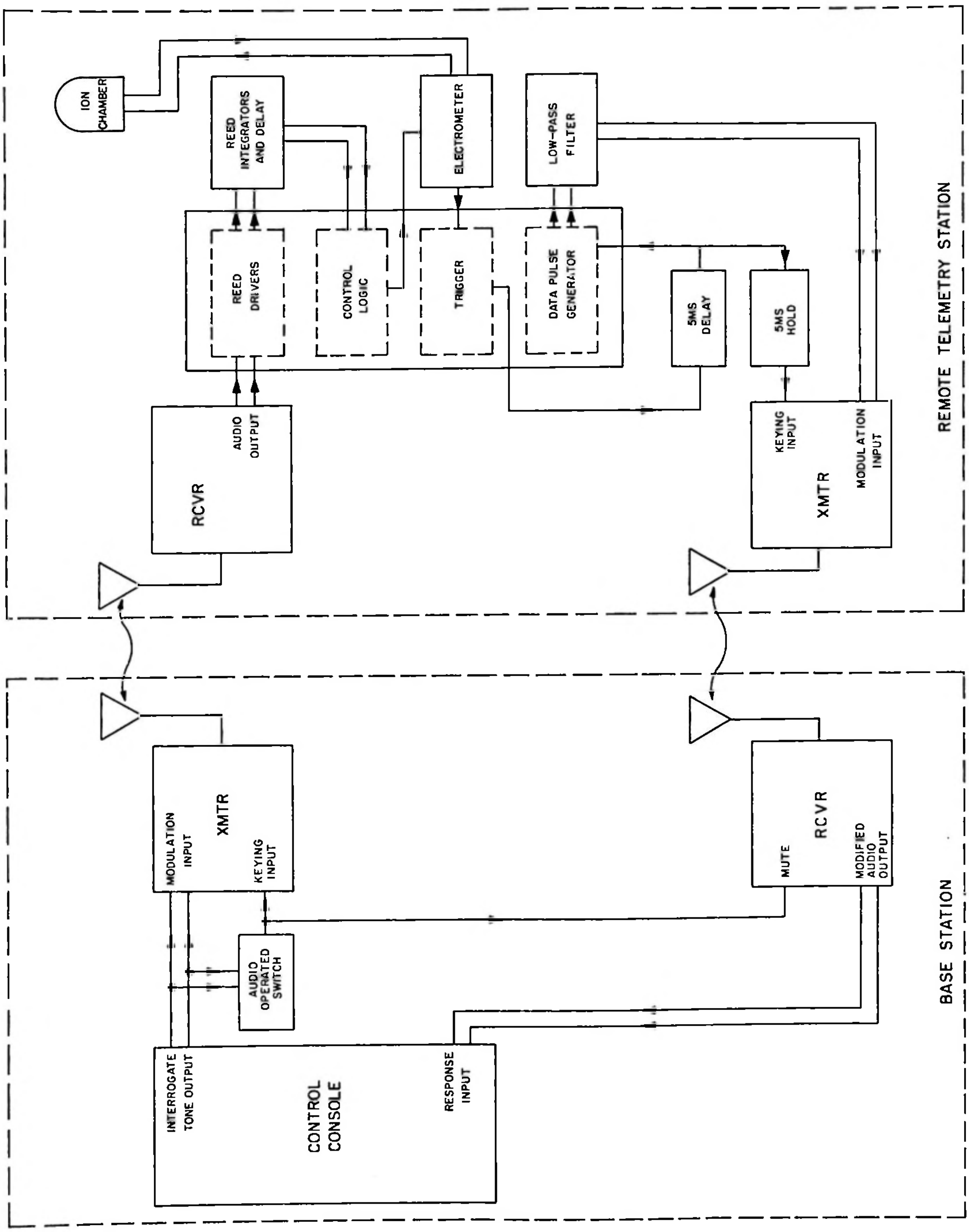


Figure 11-196K. Radio Telemetry Link, Block Diagram

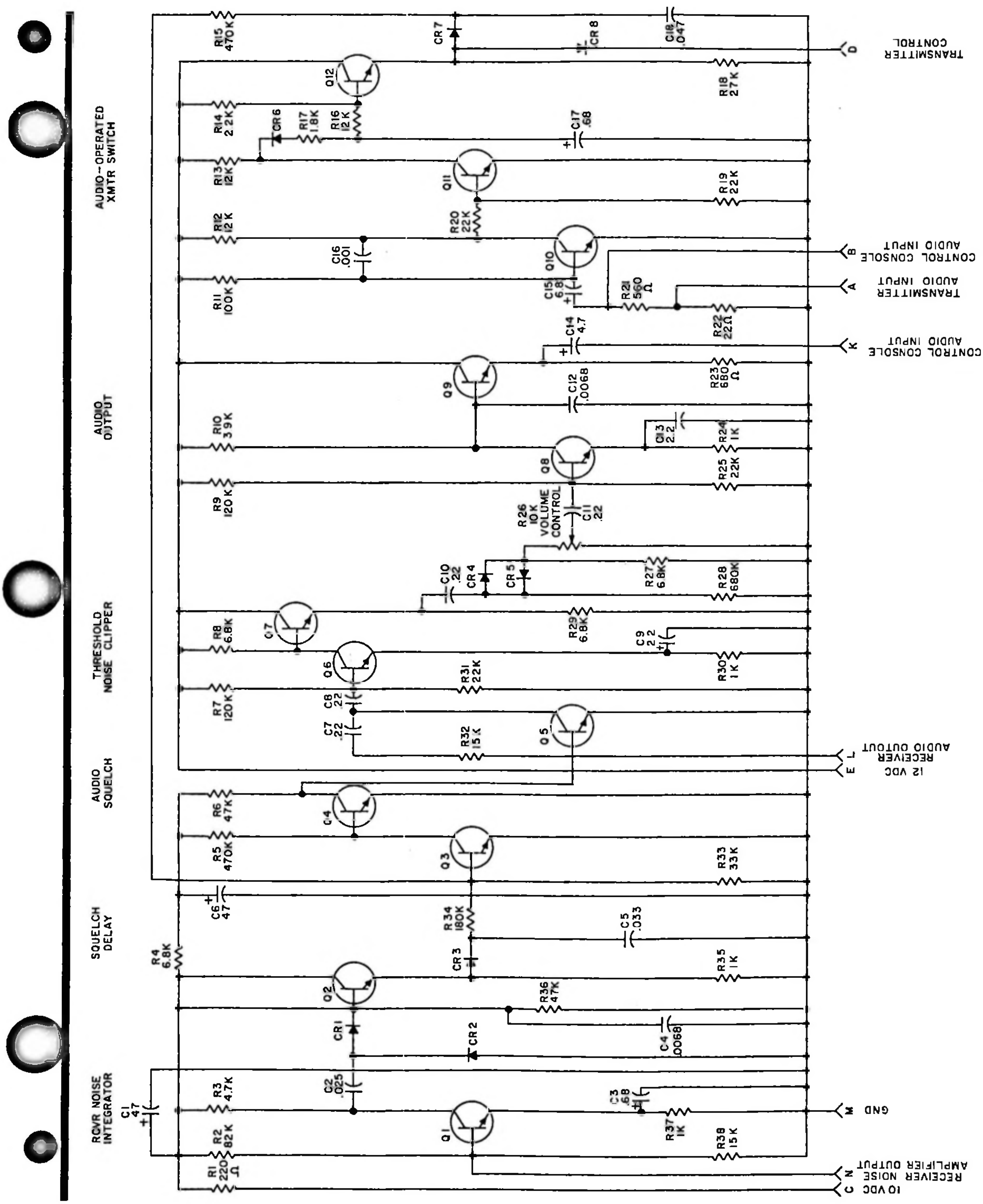


Figure 11-196L. Radio Control Console Sensor Interface Card, Schematic Diagram the receiver for the period of the pulse. The transmitter must be turned on rapidly at the start of the pulse to maintain pulse width, and it must be turned off rapidly at the end of the pulse to avoid time conflict with the reply pulse. A fast audio-operated switch supplies transmitter keying and receiver muting voltages. This switch is controlled by the tone pulses from the control console; the audio tones are integrated and the resultant dc control pulse keys the transmitter on and off within 10 milliseconds of the tone pulse edges. Voltage is continuously supplied to all stages except the oscillator. The transmitter is keyed by applying voltage to the oscillator through the audio switch in the interface circuitry. When this occurs, the oscillator operates and applies a drive signal to the class C multiplier-amplifier chain. Thus, an rf output signal is applied to the transmitting antenna.

11-1306AI. RECEIVER AND INTERFACE-RADIO TELEMETRY BASE STATION. The interface circuits in the control console discriminate between incoming data pulses and noise almost entirely on a time duration basis. That is, the signal is recognized as a desired pulse if the total duration of the signal is between 1.6 and 3.6 milliseconds. The minimum levels necessary to energize this discrimination circuitry are peak-to-peak signals which range from 90 millivolts at 2 kHz to 170 millivolts at 50 kHz. Any output signal from the receiver at any audio frequency will appear as a data pulse if the signal duration falls within the proper time limits. The carrier transmission period for each reply pulse is 10 milliseconds; the 2 milliseconds of 2-kHz modulation begins 5 milliseconds after the start of carrier transmission.

11-1306AJ. The audio and squelch circuits are on the radio-control console interface card. The schematic diagram of this interface card is contained in figure 11-196L. The audio switch used in the fast-squelch circuit is the collector-emitter path of a transistor, Q5, connected in shunt across the audio output from the receiver. Applying a dc voltage to this transistor base presents a very-low impedance path to the audio signal. Removal of the bas bias voltage causes a high collector-to-emitter impedance, subsequently, little audio attenuation is produced. Because there is no dc voltage applied to the collector of the shunt transistor, no appreciable transient voltage is generated by switching between the two operating conditions.

11-1306AK. The signal applied to the base of Q5 is provided either from the muting voltage from the transmitter keying circuit, Q10-Q12, or from a voltage derived from high-noise level coming from the discriminator. The first of these inputs ensures that no receiver transient voltage is present at transmitter switching times, and the second squelches the receiver at all times except when a sufficient carrier signal is present to quiet the receiver. In the latter case, the noise which has been separated from the audio signal by a high-pass filter is routed from the noise amplifier in the receiver module to a conditioning amplifier (Q1) and then integrated to a dc level. The diode network, CR1 and CR2, provides quick reaction to noise while maintaining the integrator time constants. The output signal of the integrating circuit is buffered by Q2, and again applied to a capacitive circuit. Application to the capacitive circuit provides a delay-squelch circuit reaction for a controlled 4 milliseconds. The capacitive circuit ensures that the receiver will not be squelched when the reply pulse modulation is received 5 milliseconds after the start of the carrier signal, but provides the maximum protection against a false pulse within the signal parameter requirements. Q3 and Q4 provide buffer and inversion functions for both this delayed noise indication and the transmitter switch-muting voltage.

11-1306AL. Following the squelching action, 2 transistors, Q6 and Q7, recondition the audio signal for the threshold clipping network. The latter, comprising a pair of complementary diodes, CR4 and CR5, will pass only those audio signals which exceed the forward bias rating of 0.7 volts. The threshold control (a signal level control at the receiver) is set so that signals which are more than a few db below the data pulse will not pass. The volume control immediately following these diodes sets the output level from the succeeding 2-stage amplifier, Q8 and Q9, at only slightly more than the threshold level required to operate the reply pulse discrimination circuits in the control console.

11-1306AM. TRANSMITTER AND INTERFACE – REMOTE TELEMETRY TRANSCEIVER. The reply pulse which is generated in, and transmitted by, the sensor unit to initiate readout at the control console is a 2-millisecond 2-kHz tone pulse (four complete cycles). Peak-to-peak signal amplitude is approximately 1.4 volts across 600 ohms. In a normal full cycle, the sensor generates this pulse four times: immediately following the reception of the 200-millisecond interrogate pulses in the check cycle, during the preset cycle, at the beginning of the r/hr cycle; and during the electrometer readout time in the r/hr cycle. The 2 pulses in the r/hr cycle can be as close together as 5 milliseconds (leading edge to leading edge) for a 10,000-r/hr reading or as far apart as 200 seconds for a 0.25-r/hr reading.

11-1306AN. Since the length of the reply pulse is of the minimum duration needed to give it recognition as a valid signal by the control console, the reply pulse must be transmitted in its entirety. Without additional signal delay, transmitter attack time and base station receiver-squelch opening time would have to be essentially zero. This is accomplished by the 5-millisecond delay one-shot multivibrator part of the interface circuitry which is located on the auxiliary circuits card in the sensor unit. The schematic diagram of the auxiliary circuits card is shown in figure 11-196M.

11-1306AO. To obtain the necessary delay, the signal which would normally key the pulse in the sensor unit is used to trigger the 5-millisecond one-shot multivibrator, Q7 and Q8. The output of this multivibrator turns on the transmitter immediately, but it also keys the 2-millisecond reply pulse with the trailing edge of the trigger signal. The transmitter is maintained in the transmit mode for another 4 or 5 milliseconds by the decay of a resistor-capacitor combination. The result is that the transmitter is keyed 5 milliseconds before the data pulse is applied as modulation. This delay adequately provides for both the finite attack time of the transmitter and the 4-millisecond squelch opening time of the base station receiver. The single disadvantage of this technique is a slight reduction in the measurable upper limit of the radiation rate. Since the 5-millisecond delay multivibrator is not capable of being retriggered until it returns to a quiescent condition, the minimum interval between pulses is approximately 6 milliseconds; this corresponds to a radiation rate of 8300-r/hr.

11-1306AP. The transmitter is normally keyed by the application of power supply output voltages to all stages, including the oscillator. The oscillator frequency is controlled by a quartz crystal. Its start time is therefore delayed several milliseconds due to the high Q of the crystal. To achieve a rapid start in the remote station transmitter, the crystal oscillator is permitted to operate continuously. To key the multiplier and the power stages, the bias on the buffer-amplifier stage following the modulator is controlled by a 3-stage switching network, Q13, Q14 and Q15, on the transceiver interface circuits card. The time delay between the application of the transmitter-keying

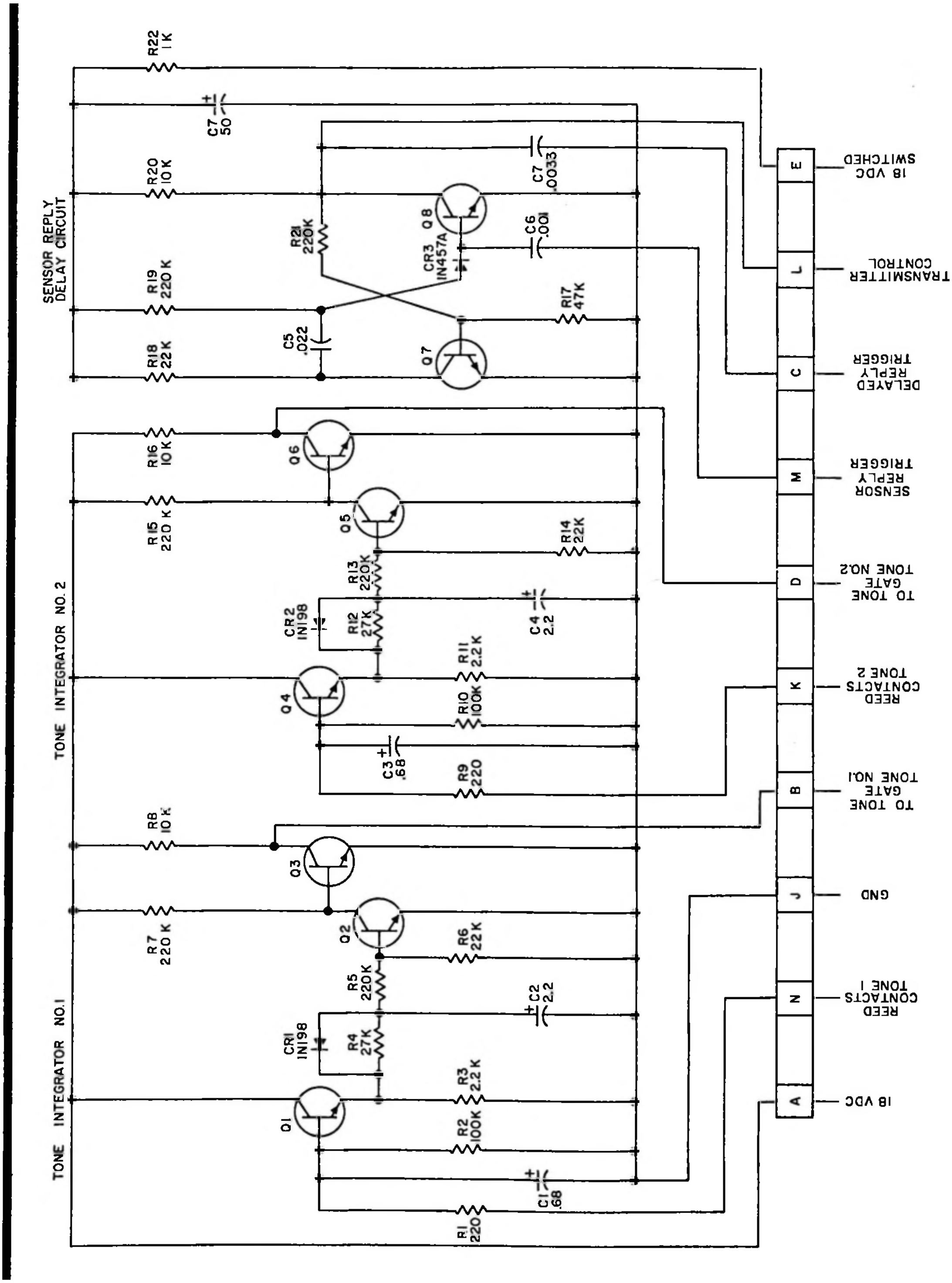


Figure 11-196M. Sensor Auxiliary Circuits Card, Schematic Diagram pulse and the appearance of a carrier signal at the antenna is approximately 10 microseconds.

11-1306AQ. The 2-millisecond, 2-kHz data pulse is obtained from a gated astable multivibrator in the sensor unit. Therefore, the waveform is a square wave rather than a sine wave. The high frequency components of this pulse fall in the same part of the audio frequency spectrum that the noise amplifier in the base station receiver uses to control the station receiver squelch circuit. As a result, there is a tendency for these distortion products to be sensed as noise which causes the base station receiver to be squelched during the modulation process. To partially eliminate this situation, the audio output from the sensor is passed through a 3-section, rc low-pass filter so that the signal applied to the transmitter modulation input more closely approximates a 2-kHz sine wave. This filter is located on the remote telemetry transceiver interface circuits card, illustrated schematically in figure 11-196N. The oscillator and the audio stages receive their power from a 12-volt switched source, thereby producing a reduced stand-by battery drain.

11-1306AR. A microphone jack is provided on the transceiver panel and it is connected in parallel with the transmitter input. Refer to figure 11-196N for a wiring diagram of the remote telemetry transceiver panel. A speaker jack connected to the receiver output circuit is also provided. This arrangement makes it possible to use the transceiver for voice communications. Normally, the speaker jack connected to the receiver output would cause reduced power with only a 12-volt supply on the transmitter output stages. However, an additional jack is provided so that an auxiliary 12-volt power supply can be used for full-power voice communications. This feature provides a two-way conversation mode.

11-1306AS. RECEIVER AND INTERFACE - REMOTE TELEMETRY TRANSCEIVER. The squelch circuit used to suppress audio output during the periods when there is no carrier signal is similar to the squelch circuit used in the base station receiver. However, since the incoming audio pulses are 200 milliseconds long, the requirements for the timing circuits are much less stringent. As shown in both figures 11-196K and 11-196D, the same transient-free shunt switch (Q4) technique is used with both the input and the output coupled to the switch. A two-stage amplifier (Q1 and Q2) is employed between the audio output of the receiver strip and the squelch switch, while only a single stage (Q3) is required between the noise-integrator network and the switch control. The signal level at the collector of the output stage (Q5) is controlled by the volume control at the output of the receiver strip.

11-1306AT. Protection of the sensor resonant reed circuits against falsing by voice communications on the shared communications frequency is essential at the remote monitoring stations. Since the equipment at these positions would be battery-operated in an emergency situation, a false turn-on of the sensor magnetic-latching relay would increase battery drain and seriously affect battery life. The sensor units were designed for voice-free private landline interconnections. Thirty milliseconds of input signal at the two resonant reed relay frequencies will activate the turn-on circuitry. To prevent this, delay circuits were added to both resonant reed relay-integrator circuits as shown in the schematic diagram of the sensor auxiliary circuits card, figure 11-196M. In these circuits, the output of the relay contacts is first connected to a dc pulse (Q1 or Q4), and then delayed in an additional capacitive network before delivery to the AND gate which drives the magnetic latching relay. An on-frequency tone of 100 milliseconds would then be required to acti-

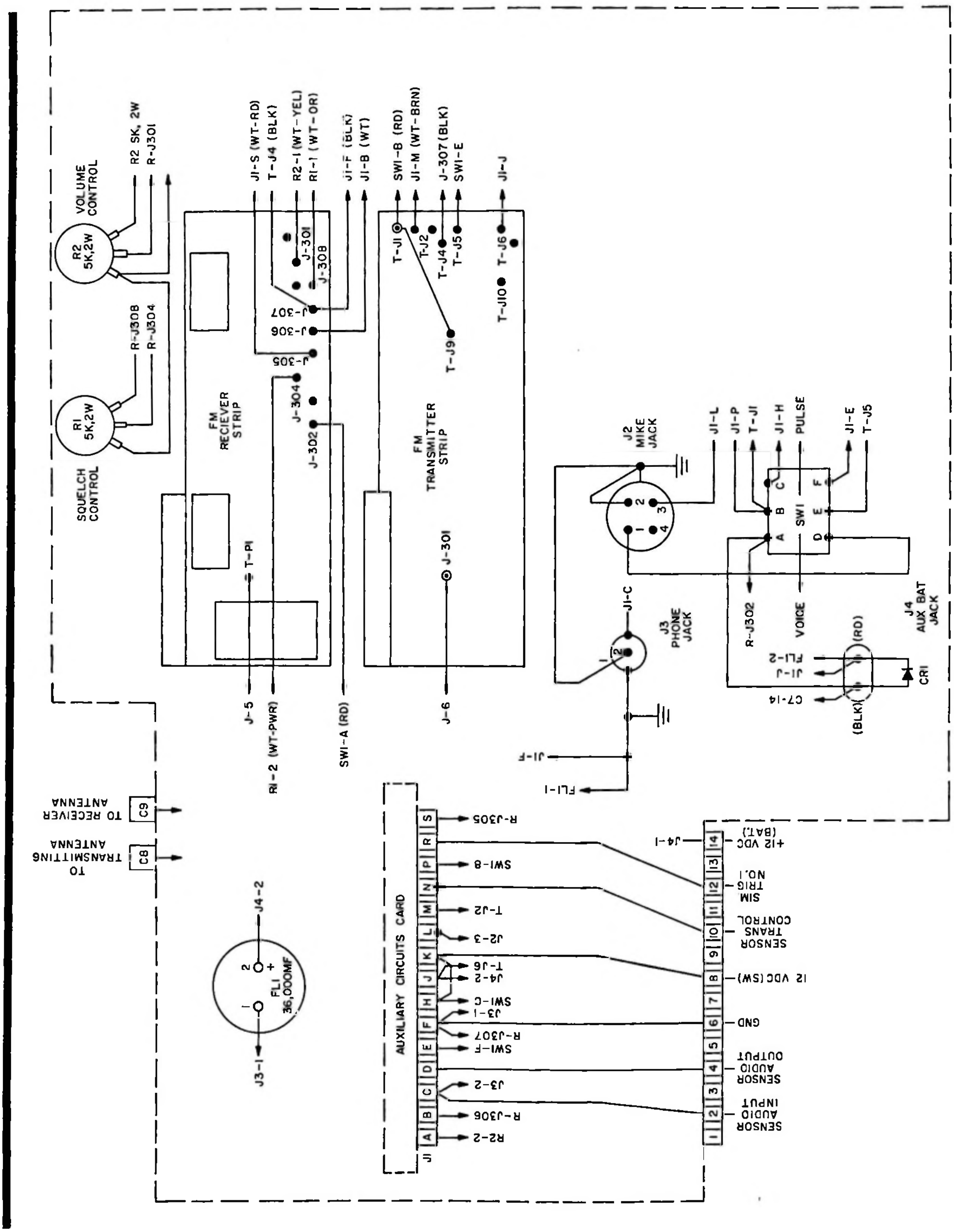


Figure 11-196N. Remote Telemetry Transceiver Panel, Wiring Diagram vate either of the resonant reed relay output circuits.

11-1306AU. INTERROGATION.

11-1306AV. Operation is initiated and continues in automatic sequence to completion after you press the start button on the control console. The operating sequence is composed of the ready position and five individual cycles during each of which a rotary stepping switch in the control console steps to each of 50 positions corresponding to the 50 possible remote sensor stations.

11-1306AW. The ready condition and the 5 cycles involved in interrogation of the equipment are displayed on the front panel of the console as READY, START, WARM, CHECK, PRESET, and R/HR. The function being performed is indicated by illumination of the corresponding window on the console. The sequence through each function proceeds as follows:

a. READY. The ready indication is displayed on the control console to inform the operator that the equipment is in a condition for interrogation. No functions are performed while the console is in the ready mode.

b. START. The start cycle ensures correct synchronization between the audio-tone generators and the output terminals to the sensor stations. This function is related to the control console only. No tone signals are transmitted to the remote sensor stations.

c. WARM. During the warm-up cycle, as the step switch advances to each position, a pulse composed of a pair of audio tones is transmitted to each remote sensor station. At each station the pulse is received by resonant reed relays, tuned to a specific audio-tone pair, translated to voltage levels, and combined to operate the latch coil of a magnetic latching relay. When the contacts of the relay latch are in the closed position, battery voltage is applied to both the electrometer and the logic circuitry, and the remote station is transferred from stand-by to an operational condition. After turn-on, the equipment pauses for 60 seconds to allow warm-up of all components.

d. CHECK. After warm-up, the console automatically proceeds to the check function, and the pair of audio tones is again transmitted to each remote sensor station. The tones are received and combined as before, but during this function, they are directed through a digital delay circuit to switch the sensor logic to the check condition. While the sensor logic is in this condition, two functions are performed:

1. The electrometer-control circuit initiates an electrometer condition in which current originating in the ion chamber (or radiation simulator) begins to charge the input network of the electrometer.

2. A 2-millisecond, 2-kHz pulse is generated and transmitted to the control console. Arrival of this pulse at the control console causes the window corresponding to that sensor station to light; thus indicating to you that the radio telemetry link is functioning and that the remote sensor station is activated.

e. PRESET. To perform the preset function, the control console again cycles through the 50 remote sensor station positions on the step switch and transmits the audio-tone pairs to each, coded as before, so that a unique combination is sent to each station.

11-1306AX. The time interval between completion of the check function and the preset function is adjustable at the control console by means of the PRESET RADIATION LEVEL switch and can be varied in steps from 20 seconds, which corresponds to 2.5 r/hr, to 1000 seconds, which corresponds to 0.05 r/hr. As each remote sensor station switches to the preset condition, the electrometer-control circuit is monitored. If the current from the ion chamber has charged the input circuit of the electrometer to the firing level during the preset dwell time, a pulse will have been emitted and the electrometer-control circuit in the sensor logic will have been reset. Under these conditions, a 2-millisecond 2-kHz pulse is generated at the remote sensor station and transmitted to the control console. If the electrometer has not reached the firing point, no pulse is generated. At the control console, a reply pulse during the preset function again lights the window for the position being interrogated and also enables the position for a readout during the r/hr cycle. If no reply pulse is received, the window in the control console will not light and the station will be skipped during the r/hr cycle.

11-1306AY. The first part of the r/hr function is similar to the check function in that the 2-tone audio pulses are again transmitted to the sensor stations but, this time only to those which are enabled during the preset function. The pulse arriving at each station switches the sensor circuitry to the r/hr condition. In this condition, the electrometer-control circuit causes the electrometer to function and to transmit the first r/hr-cycle reply pulse to the control console. A second reply pulse during this function is generated and transmitted when the electrometer reaches the firing point and the control circuit is reset.

11-1306AZ. Arrival of the first pulse at the control console starts a 1-kHz oscillator. The second pulse stops the oscillator. The output of the oscillator is directed to a counter in the control console. This counter determines the interval between the pulses by counting the number of milliseconds that the oscillator is permitted to run. Subsequently, this counter activates the readout indicators on the front panel of the control console to provide an approximate reading, in r/hr, of the radiation level at the remote sensor station.

11-1306BA. The second part of the r/hr function, which is performed for each remote sensor station before the console steps to the next sensor station, is that of turn-off. In this sequence, the 2-tone, 200-millisecond audio pulse is transmitted 3 times followed by a 200-millisecond pulse composed of only the high tone in the pair assigned to the particular remote sensor station which is being turned off. If for any reason (malfunction, temporary loss of telemetry link, etc.) the sensor unit has not advanced to the r/hr condition, the first 3 pulses of the turn-off sequence will cause the sensor unit to advance to that condition. The final single-tone pulse will then turn off the station. Reception of the single-tone pulse activates only the high frequency tuned reed relay in the sensor unit. When the reed relay is activated, the dc voltage level from the output of the high-tone relay operates the unlatch coil of the power supply relay. The contacts of this relay open and remove battery voltage from all circuits in the electrometer and in the sensor unit with the exception of the pre-amplifier and the tuned reed relay drivers. When the total operation has been performed for the last remote sensor station, the full interrogation cycles is completed and the equipment returns to the ready condition. In this condition, all stations are turned off.

SECTION VII

SYNCHRO AND SERVO EQUIPMENT TESTING

11-1307. GENERAL.

11-1308. Synchro equipment is used for remote indication or control by means of self-synchronizing motors. It consists of a series of synchro units which are used to electrically govern or follow the position of a mechanical indicator or device. The advantages of using an electrical synchro method rather than a mechanical arrangement are greater accuracy and simpler routing requirements for long-distance applications. There are five general types of synchro units, which are classified according to function. They are transmitters, receivers, differential transmitters, differential receivers, and control transformers. However, if tne power required to operate a device is large as compared with the power available from the controlling instrument (usually a synchro), power-amplifying means is provided. The term servomechanism refers to a large variety of power-amplifying devices. Servomechanisms are incorporated in synchro systems for such purposes as positioning guns, controlling radar antennas, and other automatic control applications where accuracy of reproduction is of primary importance.

11-1309. SYNCHRO AND SERVO EQUIPMENT.

11-1310. TRANSMITTER (GENERATOR) SYNCHRO.

11-1311. This unit, sometimes referred to as a synchro generator, consists of a rotor which carries a single winding, and a stator which is made up of three windings displaced 120 degrees from one another. Voltages induced in the stator windings by the rotor windings represent the instantaneous angular position of the controlling shaft of the rotor. These voltages are used to control the position of a receiving synchro connected to it.

11-1312. RECEIVER (MOTOR) SYNCHRO.

11-1313. This unit, also known as a synchro motor, follower, or repeater, is similar electrically to the synchro transmitter, and is used in conjunction with a transmitter synchro. The receiver synchro takes the electrical signal voltage generated by the transmitter synchro; the receiver synchro rotor follows in response to this voltage, so that its angular position corresponds to the position of the transmitter synchro rotor. Mechanically, the receiver synchro differs from the transmitter synchro in that a damping device is incorporated to prevent overshooting and hunting. For this reason, a transmitting synchro may not be used for receiving, although a receiver synchro may be used for transmitting.

11-1314. DIFFERENTIAL SYNCHROS.

11-1315. Differential synchros are used in conjunction with transmitter and receiver synchros. A transmitting differential synchro is used to insert a correction voltage from the transmitter synchro to compensate for errors existing in various parts of the system. In effect, the angular position of the transmitter synchro and the angular position of the differential synchro are compared, and the sum or the difference of these two positions is transmitted to a receiver synchro. Whether the sum or the difference voltage of the differential synchro is employed depends upon the method used to connect the transmitter, differential, and receiver synchros. A receiving differential synchro indicates the angular sum or difference (depending upon the connections) between two transmitter positions.

11-1316. CONTROL TRANSFORMER SYNCHROS.

11-1317. A control transformer synchro, known as a ct synchro, is used where it is desired to obtain only a voltage indication of angular position. The ct synchro is similar to an ordinary synchro except that its rotor windings are used only for generating a voltage, known as an error voltage. Because this voltage is fed to the control grid of an electron-tube amplifier, the rotor windings are wound with many turns of fine wire, to produce a high impedance. Since the rotor is not fed an exciting voltage, the current drawn by the stator windings of the ct synchro would be fairly high if the windings were of the same type as those in an ordinary synchro. To prevent this current from being excessive, the stator windings are also wound with many turns of fine wire, to present a current-limiting impedance. In addition, a capacitor is connected across each of the three windings to reduce the current flow (see paragraph heading 11-1318). In normal operation, the output from a ct synchro is nearly zero (nulled) when its angular position is the same as the generator synchro.

11-1318. SYNCHRO CAPACITORS.

11-1319. The differential synchro and the control transformer synchro both draw current from the synchro transmitter, even when the circuit reaches a position of electrical and mechanical balance. The differential synchro draws current because of the step-up turns ratio between the stator and the rotor. The control transformer synchro draws current from the transmitter because the control transformer synchro rotor is not energized, and, as a result, induces no voltage in the stator windings. To reduce current flow when either or both of these units are used, capacitors are connected into the circuit.

11-1320. Because the windings of the synchro are not a pure inductance, two currents exist in the windings; the resistive (loss) current, which represents the actual power loss in the circuit, and the inductive (out-of-phase) current, called the magnetizing (or exciting) current. When the loss current and the magnetizing current are added vectorially, the actual current lags by something less than 90 degrees. When the values of inductive reactance and resistance are known, a specific capacitance can be added to shunt the coils, since current leads in a capacitive circuit, so that the inductance-capacitance circuit will resonate at 60 cps, and the magnetizing current will be effectively cancelled by the capacitive current. In this state, the transmitter synchro will supply only the losses of the circuit.

11-1321. Synchro capacitors are manufactured with three units contained in a single package, and connected internally, as shown in figure 11-197. It must be stressed that synchro capacitors are used only when it is desired to cancel or partially cancel an

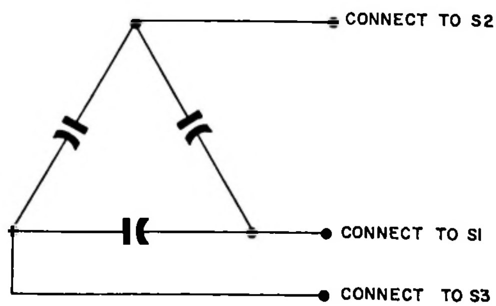


Figure 11-197. Synchro Capacitors and Connections

exciting current. Synchro capacitors are not used in a simple synchro transmitter-receiver circuit because in such a circuit, when balanced, the stator current is assumed to be zero; the use of a synchro capacitor would only increase the current.

11-1322. SERVO CIRCUITS.

11-1323. A servo circuit is a more complex method of synchro control, in which a controlled quantity is compared with an ordered quantity, and the difference between the two quantities, known as the error, is used to govern the operation of a mechanical system. There are a great variety of servo methods in use, and it is not within the scope of this book to describe the circuits in detail. There are electronic types, hydraulic types, amplidyne types, and many variations and combinations of these. All of these types of servo methods are designed for a specific task; the applicable technical manual should be referred to for specific testing and servicing instructions. All of the servo methods have an anti-hunt feature embodied in their design, in order to prevent oscillation. Electronic amplifiers used in servo circuits are classified according to the following four basic types:

a. Type 1: dc input, dc output

b. Type 2: dc input, ac output

c. Type 3: ac input, ac output

d. Type 4: ac input, dc output

Block diagrams illustrating these four basic types are shown in figure 11-198. For specialized uses, even these four basic types are varied to suit the particular application for which they may be needed.

11-1324. SYNCHRO EQUIPMENT TESTING.

11-1325. Since synchros are employed to transfer angular shaft position to another synchro, usually some distance away, long

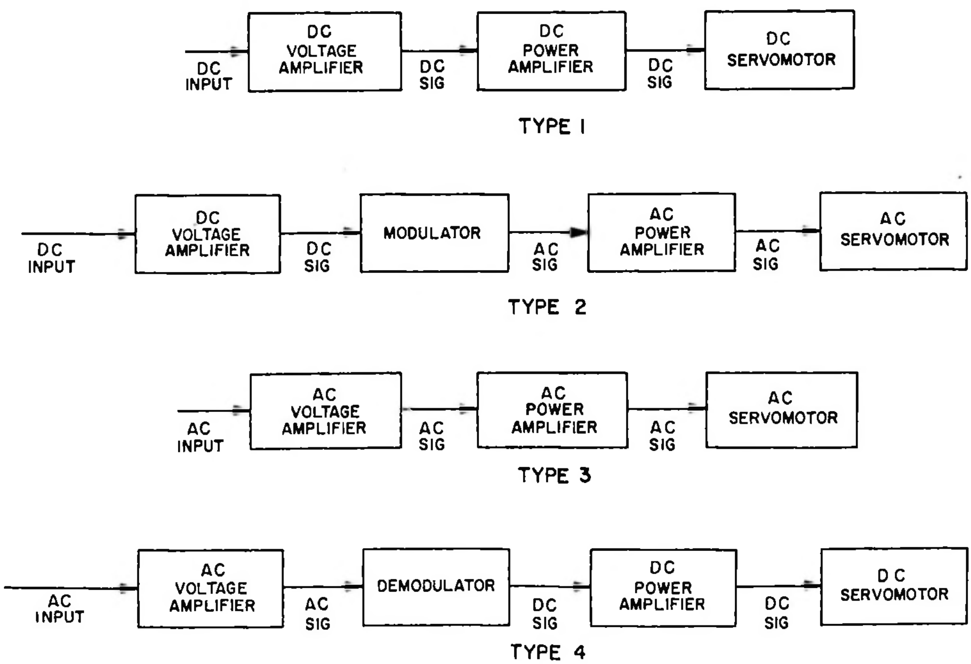


Figure 11-198. Servo Amplifiers (Electronic), Basic Types

lengths of connecting bus and/or cable are used. Although the wiring may be clearly marked or color-coded, it is advisable to check these designations if a synchro device gives evidence of improper operation. This is important if a new installation is being checked, or if an installation has been repaired or overhauled.

11-1326. OVERLOAD INDICATORS.

11-1327. An overload in a synchro circuit is usually caused by worn bearings or defective gears at the receiver synchro. This condition causes the receiver rotor to lag the transmitter rotor, allowing excess current flow in the stator windings. To detect this condition, it is necessary to measure the current in at least two of the stator leads. This is because synchro design makes it possible for one stator lead to indicate zero current while the other two leads are drawing excessive current. The usual procedures and precautions should be followed when making these measurements. Some synchro circuits may have an overload indicator included in the installation. This method uses two current transformers, the primary of each being connected in series with a stator lead. The secondary windings of these two transformers are connected in series-aiding, and the two remaining secondary leads are connected to a neon bulb. The secondary windings of these two transformers are so designed that the neon bulb fires when a predetermined unbalance in current in the two stator leads is present. The neon bulbs are usually mounted on the control switchboard of the equipment.

11-1328. BLOWN-FUSE INDICATORS.

11-1329. Some synchro circuits may have a blown-fuse indicator included in the installation. This usually consists of a transformer with two primary windings and one secondary winding. The primary power is connected to one primary winding, and the synchro excitation voltage is taken from the other primary winding. The leads of one primary are jumpered by fuses to the leads of the other primary, the phasing being such that the voltages in the two windings oppose each other, so that normally no voltage is induced in the secondary. If one of the fuses blows, the primary winding connected to the primary power induces a voltage in the secondary winding. This secondary winding is connected to a neon bulb, which glows, indicating a blown fuse.

11-1330. VOLTAGE AND RESISTANCE MEASUREMENTS.

11-1331. The quickest method of locating opens and shorts in synchro units and their associated wiring is by resistance measurements. Since most synchros work in pairs, it can be assumed that the resistances of both the rotors and the stators will show the same reading, within close tolerances. If the resistances should vary widely, the trouble may be easily located. Typical resistance values for synchros may run from a fraction of an ohm for the large synchros to a few hundred ohms for the smaller ones. Do not measure resistance without first shutting off all excitation voltage to the synchro rotors.

11-1332. An excellent method for detecting open or shorted stator windings is to connect a voltmeter across any two of the stator windings. As the angle of the transmitter is varied, a smooth variation between 0 and 90 volts should be indicated by the voltmeter (assuming a 110-volt synchro). Open- or short-circuited stator or rotor leads may be detected by measuring with a voltmeter or by measuring the resistance of the suspected part.

11-1333. SYMPTOMS OF INCORRECT WIRING.

11-1334. In new installations and after repairs or overhauls in synchro circuits, the crossing of buses or wires is frequently the cause of improper synchro operation. Reference to figure 11-199 should identify the wiring fault that is causing improper operation of the synchro circuit.

FOR ALL WIRING TROUBLES SHOWN HERE, RECEIVER GIVES WRONG INDICATION OR TURNS IN A REVERSE DIRECTION. TORQUE NORMAL. NO OVERLOAD. NO OVERHEATING.

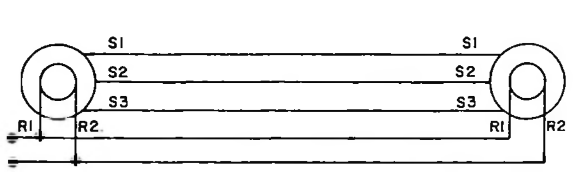
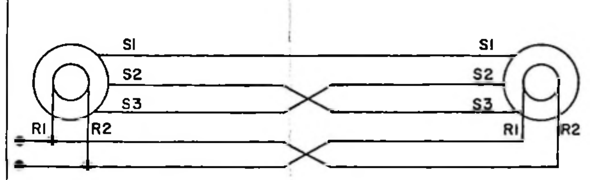
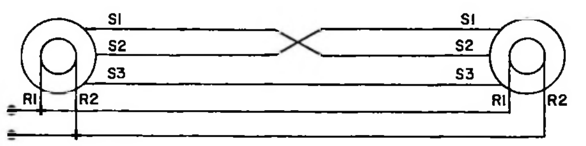
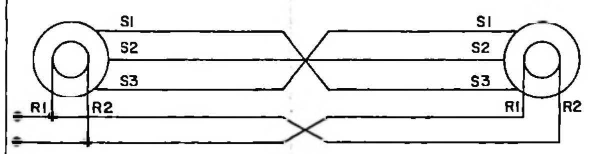
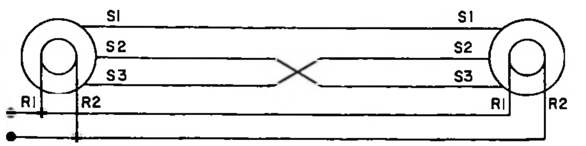
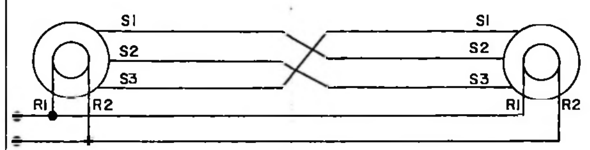
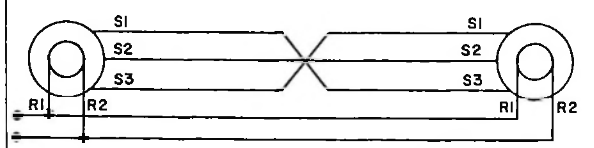
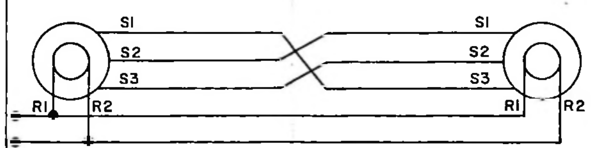
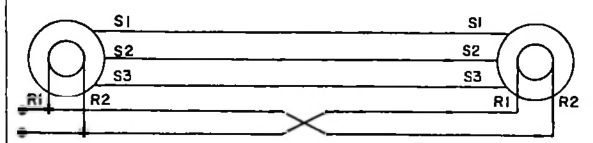
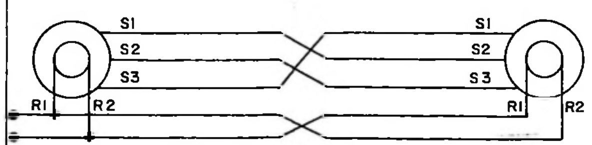
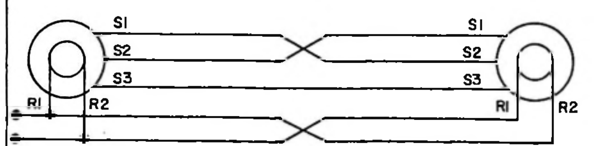
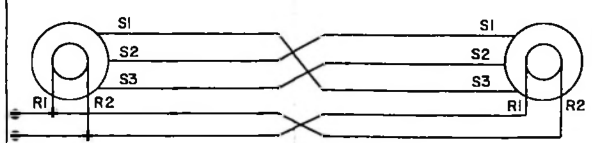
SPECIFIC SYMPTOMS		WIRING	TROUBLE	SPECIFIC SYMPTOMS		WIRING	TROUBLE
TRANSMITTER	RECEIVER			TRANSMITTER	RECEIVER		
SET ON 0° AND TURNED CCW	READS AND TURNS AS INDICATED			SET ON 0° AND TURNED CCW	READS AND TURNS AS INDICATED		
0°	0°		NO TROUBLE SYSTEM OPERATES NORMALLY	0°	300°		R1 AND R2 REVERSED S2 AND S3 REVERSED
0°	240°		S1 AND S2 REVERSED	0°	180°		R1 AND R2 REVERSED S1 AND S3 REVERSED
0°	120°		S2 AND S3 REVERSED	0°	120°		S1 TO S2 S2 TO S3 S3 TO S1
0°	0°		S1 AND S3 REVERSED	0°	240°		S1 TO S3 S2 TO S1 S3 TO S2
0°	180°		R1 AND R2 REVERSED	0°	300°		R1 AND R2 REVERSED S1 TO S2 S2 TO S1 S3 TO S1
0°	60°		R1 AND R2 REVERSED S1 AND S2 REVERSED	0°	60°		R1 AND R2 REVERSED S1 TO S3 S2 TO S1 S3 TO S2

Figure 11-199. Incorrect Synchro Connections Causing Receiver to Operate in Wrong Direction or Give Improper Indication

11-1335. SYMPTOMS OF OPEN- AND SHORT-CIRCUITED WIRING.

11-1336. Troubles involving open- and short-circuited wiring, with associated symptoms, are listed in table 11-10.

11-1337. SYNCHRO ZEROING METHODS.

11-1338. In any synchro circuit, it is important to have all the synchros electrically zeroed. Since different types of synchros must be zeroed by different methods, each type of synchrowill be discussed separately.

CAUTION

In some of the methods of electrical zeroing described in the following paragraphs, it should be noted that 115 volts is applied directly to the stator coils. Since stator coils are normally designed for a maximum of 90 volts (across any two coils), the line voltage should be applied for only short periods of time. When it is necessary to apply voltage for a considerable length of time, 78 volts should be used.

Table 11-10. Trouble-Shooting Synchro Circuits

SYMPTOMS	POSSIBLE CAUSE OF TROUBLE	REMEDY
Receiver rotor is either in correspondence with transmitter or 180 deg displaced, but follows in proper direction. Stator voltages vary from 0 to 90 volts. Both rotor voltages are 115 volts.	Rotor winding is open, connection to slip ring is open, or brush is not making contact.	If sure that trouble is not in ring connection or brush, replace the unit.
Same as above except that one rotor voltage is 115 volts and the other is 90 volts.	Supply line is open to rotor, and reading 90 volts; the 90 volts exists across the rotor by virtue of transformer action.	Locate open in supply line, and repair.
Voltage between one pair of stator wires is zero for all transmitter positions. Other stator-lead voltages read from 0 to 90 volts. Both rotor voltages are 115 volts.	Pair of stator leads which read 0 volts is short-circuited. For rotor behavior, see figure 11-199.	Remove short circuit from wiring or interconnecting switches. If the trouble is internal, the unit may require replacement.
Both transmitter and receiver units hum and heat excessively. Receiver either does not follow, or may spin.	The three stator wires are short-circuited together.	Locate defective wiring or switches, and repair.

Table 11-10. Trouble-Shooting Synchro Circuits (Continued)

SYMPTOMS	POSSIBLE CAUSE OF TROUBLE	REMEDY
Sudden change in transmitter rotor position causes oscillation at receiver or a spinning effect.	Inertia damper is jammed tight on receiver rotor shaft. Absence of damper indicates that transmitter unit is being used.	Free the damper if it is jammed. If transmitter has been used, replace with a receiver unit.
Intermittent operation.	Corroded rings, defective brushes, loose connections.	Clean the rings, install new brushes, tighten loose terminals, etc.
Torque normal. Receiver lags or leads transmitter, or may turn in proper direction or reverse direction.	Stator wiring incorrect. See figure 11-199 for specific symptoms.	Correct stator wiring.
Torque normal. Receiver follows transmitter, but is displaced 180 deg from it.	Rotor connections are reversed.	Correct wiring at proper unit.
Receiver shows large error, and lags transmitter. Connections normal, but excessive current flows, producing overload indication.	Bearings are frozen or partially frozen because of improper lubrication.	Replace unit, since bearing trouble usually damages other parts of unit.

11-1339. ZEROING RECEIVER SYNCHROS. Since the receiver synchro is usually free to turn, the jumper method of zeroing is usually employed. To zero a receiver synchro, the voltage between S1 and S3 must be made zero, and the phase of the voltage at S2 should be the same as the phase at R1. This is easily done by connecting S1 and S3, using a jumper wire, and connecting S2 and R1, using a jumper wire. This method is shown in figure 11-200. When the power is applied, the rotor will line up in the zero position. If the indicator does not point to zero on the dial, loosen the synchro in its mounting and rotate it until its dial reads zero.

11-1340. A second method, using a voltmeter, may also be employed for electrically zeroing the receiver synchro. Since this method is the preferred method for zeroing transmitter synchros, the procedure is given in paragraph 11-1341.

11-1341. ZEROING TRANSMITTER SYNCHROS. To zero a transmitter synchro, connect an ac voltmeter between S1 and S3, as shown in part (A) of figure 11-201. Rotate the energized rotor until a zero reading is obtained on the voltmeter. Since the rotor at zero-degree and 180-degree positions will produce this zero reading, it will be necessary to determine whether the phase of S2 is the same as that of R1. Make the connections shown in part (B) of the figure. If the proper polarity relationships exist, the voltmeter will indicate less than the line voltage being applied to the rotor. If the indication is greater than the line voltage, the rotor must be rotated 180 degrees and the previous step, as shown in part (A) of the figure performed again; the pointer connected to the rotor should be adjusted to indicate zero.

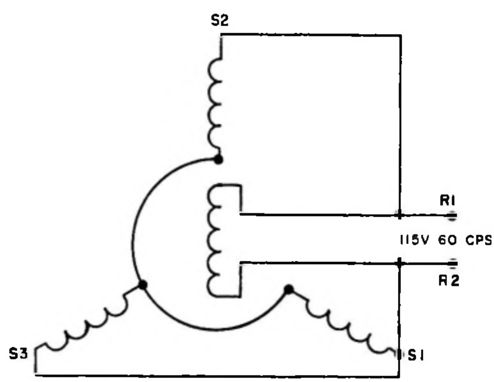


Figure 11-200. Electrically Zeroing a Receiver Synchro, Using the Jumper Method

11-1342. ZEROING DIFFERENTIAL TRANSMITTER SYNCHROS. Because the differential transmitter synchro is usually used to insert a correction voltage into a synchro circuit, it is normally driven either directly or through a gear train. Before zeroing the differential transmitter synchro, the unit whose position the differential synchro transmits should first be zeroed. After this has been done, connect the differential synchro as shown in part (A) of figure 11-202. Turn the synchro in its mounting until the voltmeter shows a minimum indication. After completing this step, make the connections shown in part (B) of the figure. Again turn the synchro slightly in its mounting until minimum voltage is indicated by the voltmeter.

11-1343. ZEROING DIFFERENTIAL RECEIVER SYNCHROS. To zero a differential receiver synchro, make the connections shown in figure 11-203. As soon as the power is applied to the synchro, the rotor will assume a position of electrical zero. The dial can then be set at zero and the unit reconnected to its circuit.

11-1344. ZEROING CONTROL TRANSFORMER SYNCHROS. To zero a control transformer synchro, connect it as shown in part (A) of figure 11-204. Apply power

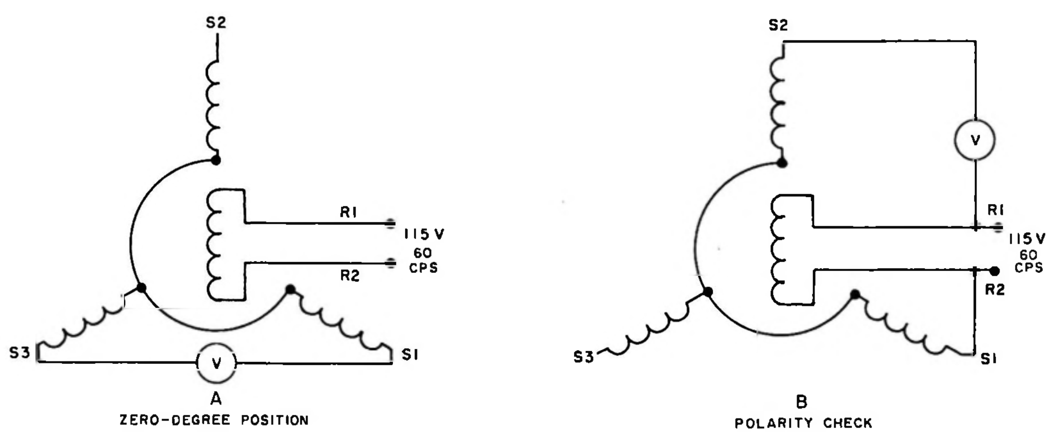


Figure 11-201. Electrically Zeroing Transmitter and Receiver Synchros, Using the Voltmeter Method

and turn the synchro in its mounting for minimum reading on the voltmeter. Then connect the control transformer synchro as shown in part B of the figure, and again turn the synchro slightly in its mounting in either direction, for minimum indication on the voltmeter.

11-1345. STANDARD TEST SYNCHROS.

11-1346. A standard test synchro is used for performing various operational tests on synchro circuits, and may be used for various kinds of checks and for troubleshooting. A standard test synchro is a small, precision synchro, mounted in an instrument case. It is equipped with a standard dial (numbers increasing in the clockwise direction), which moves past an engraved index. When the synchro is being used as a transmitter, a braking arrangement applies friction to the shaft. When the synchro is being used for receiving, the brake is released to allow the shaft to turn freely.

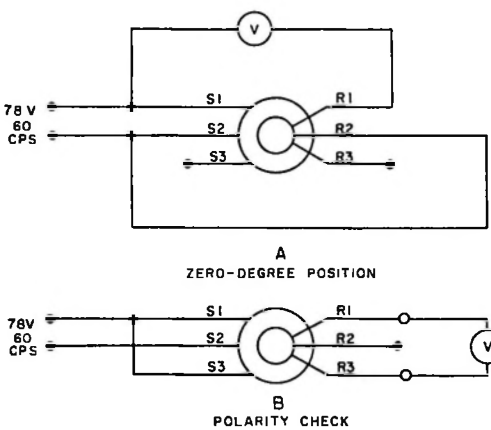


Figure 11-202. Electrically Zeroing a Differential Synchro Transmitter, Using the Voltmeter Method

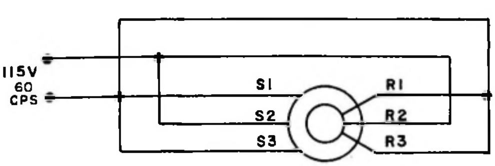


Figure 11-203. Electrically Zeroing a Differential Synchro Receiver

11-1347. <u>ELECTRONIC CONTROL METHODS.</u>

11-1348. GENERAL.

11-1349. There are a great number of servo circuits (servomechanisms) in which power is supplied to a driving motor from the output of a servo amplifier. Electronic control methods are generally employed where large amounts of torque are not required. Both dc and ac servomotors may be controlled directly from the output of electronic servo amplifiers.

11-1350. DC SERVOMOTOR METHOD.

11-1351. A schematic layout of the elements constituting a simple positioning servo circuit is shown in figure 11-205. The dc servomotor has a permanent-magnet field, a type which may be used when the load on the output shaft is not too heavy. In the circuit shown, the load is positioned without the use

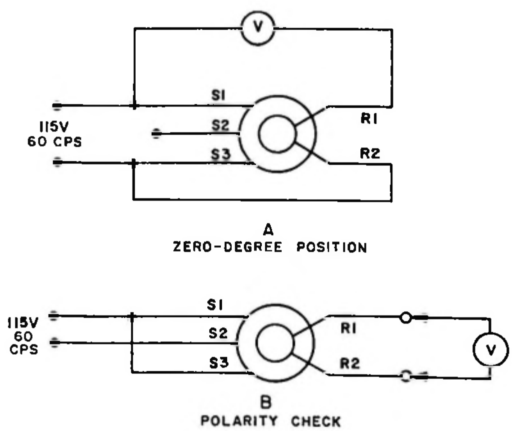


Figure 11-204. Electrically Zeroing a Control Transformer Synchro

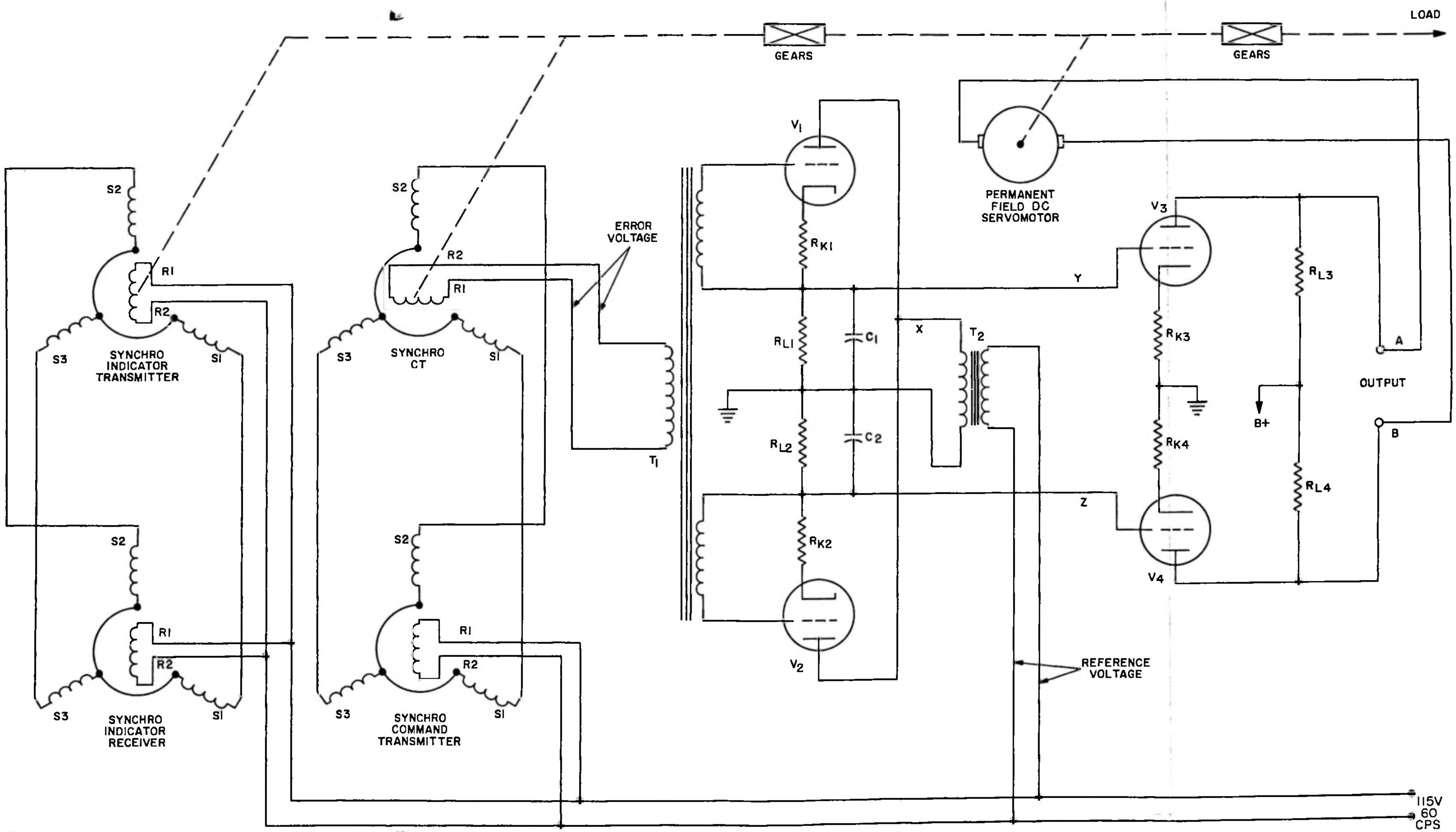


Figure 11-205. Servomechanism Control of DC Servomotor

of electromechanical amplification, the electronic amplifier alone supplying sufficient power to cause motor armature rotation.

11-1352. The servo amplifier in the positioning circuit consists of a phase-sensitive detector-amplifier, V1 and V2, coupled to a dc amplifier, V3 and V4, by means of a cathode-loading arrangement. The error voltage from the control transformer synchro rotor is coupled to the amplifier through transformer T1. It should be noted that the plate supply for V1 and V2 is an alternating voltage obtained from the secondary of transformer T2. The windings of T2 are connected so that the plate voltages of V1 and V2 are in phase, both voltages swinging positive or negative at the same time. In order that the grid and plate returns may be grounded, load resistors R1 and R2 are placed in the cathode circuits rather than in the plate circuits. The grids of the ac amplifiers are fed from the high sides of R1 and R2. Thus, each is supplied with a dc voltage, filtered by shunt capacitors C1 and C2. The remainder of the dc amplifier circuit is conventional, with the output of a rectified power supply applied to the junction of load resistors R7 and R8.

11-1353. With no error voltage applied, both V1 and V2 conduct equally when point X is positive. As a result, both outputs, Y and Z, are at positive dc potentials of equal magnitude. These outputs cause V3 and V4 to conduct the same amount and to produce equal voltage drops across R7 and R8. Consequently, the output to the servomotor armature is zero. When point X swings negative, neither V1 or V2 conducts, the grids of V3 and V4 are balanced by the equal potentials from C1 and C2, and the output voltage to the servomotor armature remains at zero. When an error voltage is present, so that the grid of V1 is positive at the same time as point X, the plate current through V1 increases, while that through V2 decreases. Since point Y is now more positive than point Z, V3 conducts more heavily than V4, effecting an unbalance in the voltage drops across R7 and R8. This means that the output to the servomotor is negative at A with respect to B. When the error voltage is reversed in phase (180 degrees difference), V2 and V4 become the heavier conducting tubes, causing the polarity at A with respect to B to reverse from the previous condition. The phase relationship between the error and reference voltages determines the direction of servomotor rotation. The speed of motor rotation depends upon the magnitude of the error voltage. The output shaft is geared to the ct synchro rotor, which turns in the proper direction to reduce the error voltage to zero. The position of the load is reported to the command position by the synchro indicator system at the left-hand side of the diagram.

11-1354. In cases where the load to be positioned is very light, a small, low-power servomotor may have its armature connected directly to point Y and Z, thus eliminating the dc amplifier, V3 and V4.

11-1355. AC SERVOMOTOR METHOD.

11-1356. One method for controlling an ac servomotor by electronic means is shown in figure 11-206. The circuit shown is used in certain radar facilities in which an ac servomotor is geared to the deflection coil of a ppi scope. The synchro circuits used with this servomechanism may be duplicates of those shown in the dc example of figure 11-205, and are therefore not included in the diagram.

11-1357. The error voltage is taken from the synchro control transformer and fed to a conventional RC-coupled ac amplifier. R1, R3, C1, and C2 are components of an error-rate damping network, the operation of which is discussed in a later section of the manual. The uncontrolled phase winding of the 2-phase ac servomotor is connected to a 115-volt, 60-cycle source, which also serves as the reference voltage. Capacitor

C6 is used to provide a 90-degree phase shift (necessary for two-phase motor operation) between the motor winding and the reference voltage. A portion of the output voltage is applied as degenerative feedback to the cathode circuit of V2, with the amount of feedback controlled by the servo gain potentiometer.

11-1358. With zero error voltage, the controlled phase of the servomotor is not energized, and there is no load rotation. When an error signal is present, the amplified signal, which appears across the secondary of the output transformer, will be either in phase, or 180 degrees out of phase, with the reference voltage. Thus, the direction of servomotor rotation depends upon the phase relationship existing between the error and reference voltages. As in other methods discussed, the servomotor shaft, which is geared to the load, is also geared to the ct synchro rotor, providing rotation in the proper direction to reduce the error.

11-1359. THYRATRON CONTROL.

11-1360. In considering armature current control methods for servomotor operation, an amplifier using vacuum tubes is found to be satisfactory for lightly loaded circuits. However, when load requirements increase, calling for more powerful servomotors and accompanying larger load currents, vacuum-tube control becomes inefficient. This is because the vacuum tube has an inherently high plate resistance, which imposes serious limitations in cases of high current demands. Of course, where current demands are heavier, larger vacuum tubes could be employed, necessitating higher operating voltages. However, the additional expense and the increased size and weight of such amplifier units are disadvantages which cannot be ignored. One solution to this problem lies in the use of the thyratron tube.

11-1361. The thyratron is a gas-filled, grid-controlled tube, capable of handling much greater load currents than a vacuum tube of

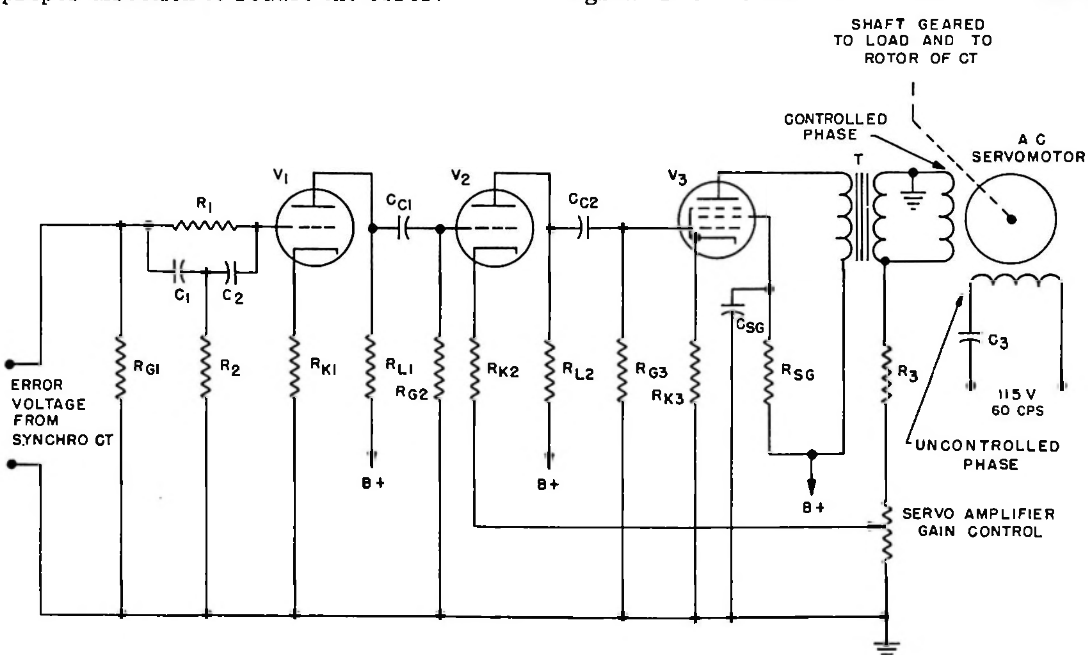


Figure 11-206. Servomechanism Control of AC Servomotor

equivalent size. It differs from the vacuum tube in three important respects. First, its plate-to-cathode resistance is so low in the conducting state that the voltage drop across the tube rarely exceeds 15 volts for full-rated current. Second, while the grid of the thyratron controls the point at which the thyratron fires, or ionizes, once ionization takes place, the grid loses all control and cannot stop plate-current flow. As a result of this thyratron characteristic, the tube can be extinguished only by lowering the plate voltage to the deionizing level. Third, the thyratron passes a current determined mainly by the series resistance and applied voltage. Therefore, while both types of tubes act as rectifiers, the vacuum tube may be regarded as a rheostat in series with the load, whereas the thyratron action is that of an on-off switch. Since it is not desirable for a servomotor to jump from no-speed to full-speed, circuits are devised whereby the average current flow is controlled by shifting the firing point of the tube.

11-1362. CRITICAL GRID BIAS CURVE. By varying the bias on the grid of the tube, the point at which the thyratron "switch" closes may be controlled. That is, the grid bias determines the value of plate-to-cathode voltage at which ionization takes place. For any value of plate voltage, there is a maximum negative grid potential, called the critical grid bias, below which the thyratron cannot fire. For a constant dc plate voltage, there is a constant critical grid bias. For an ac plate voltage, varying sinusoidally from zero to peak voltage values, the critical grid bias varies in inverse proportion, as illustrated in figure 11-207. Part (A) of the diagram shows that firing occurs at the point where the applied grid bias voltage level intersects the critical grid bias curve. If the grid bias voltage level is made more negative than the peak of the critical grid bias curve (part (B) of the figure), they do not intersect and the thyratron cannot fire.

11-1363. When an ac signal, in phase with, and of the same frequency as, the plate-supply voltage, is superimposed upon the dc grid-bias level, the thyratron may be made to conduct during different portions of the positive half-cycle. Thus in figure 11-208 it is seen that the firing point may be delayed by varying the amplitude of the ac grid signal. Examination of the curves makes it apparent that this amplitude control method cannot delay the firing point of the tube beyond 90 degrees. Within the 90-degree limit, varying the grid voltage allows the thyratron to control the speed of the servomotor, in addition to its rectifying and switching functions.

11-1364. THYRATRON SERVO AMPLIFIER FOR DC SERVOMOTOR CONTROL. A circuit for a thyratron servo amplifier, using the amplitude method of control, is illustrated in part (A) of figure 11-209. Unlike most input transformers, the secondary windings of T1 are connected so that

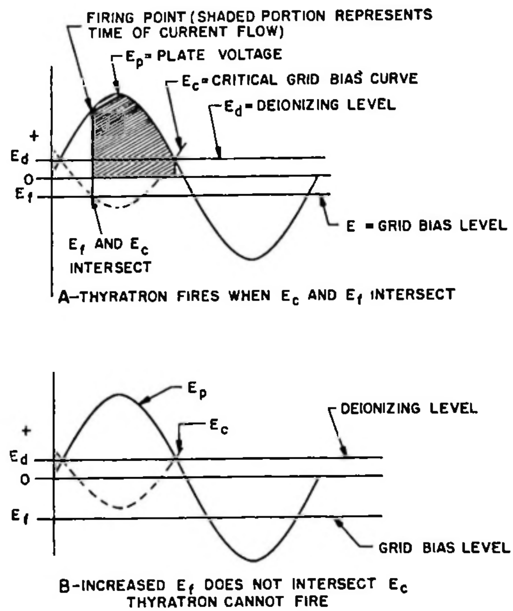


Figure 11-207. Grid Bias Thyratron Control

the grid voltages of V1 and V2 are in phase. The two secondaries of T2, on the other hand, are arranged so that the plate voltages of V1 and V2 are out of phase, as indicated by the instantaneous polarities marked on the diagram. Part (B) of the figure shows the plate-voltage curves for V1 and V2, with the corresponding critical grid bias curves and dc grid-bias levels. In the example shown, the error voltage is zero, and the curves indicate that values of grid-bias voltage have been chosen so that E_f does not intersect the critical grid bias curves. Consequently, neither tube can fire and no current is supplied to the armature of the servomotor.

11-1365. Figure 11-210 shows the amplifier action when an error signal is present. In this instance, E_{g1} is in phase with E_{p1}, and intersects the V1 critical grid bias curve (E_c) as indicated. V1, therefore, conducts during a portion of each positive alternation of E_{p1}. When the V2 plate voltage swings positive, E_{g2} is negative-going and does not intersect E_c. Thus, V2 cannot conduct. Current flows from Y to X through the servomotor armature, causing rotation in one direction. If the error voltage from the control transformer shifts 180 degrees, so that it is in phase with E_{p2} (figure 11-211), V2 conducts and V1 becomes the nonconducting thyratron. Current now flows from X to Y through the armature, and the servomotor rotation is in the opposite direction.

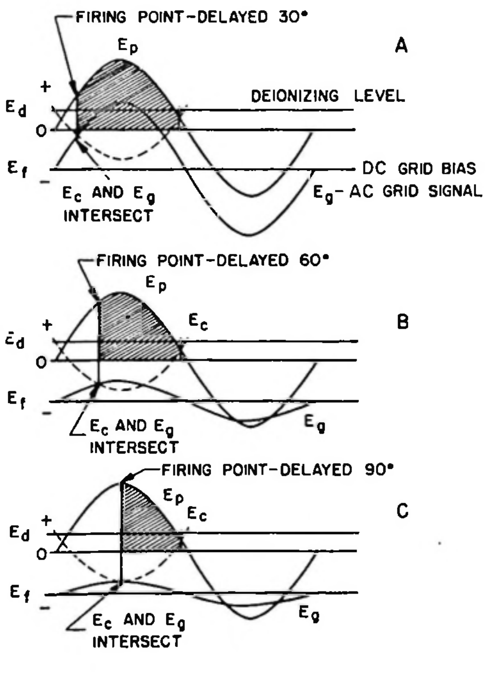


Figure 11-208. Thyratron Firing Point as Controlled by Amplitude of In-Phase Grid Signal

11-1366. PHASE-SHIFT CONTROL. It has been noted in the previous discussion that amplitude control of the thyratron firing point is limited to a 90-degree delay. It is obvious that if by some means the time of firing is extended beyond the 90-degree point, a more proportional control will be realized, resulting in an improved servomotor response. A method whereby the firing point may be varied over almost a 180-degree range is known as phase-shift control. This method consists of adding a third, or reference, voltage to the two voltages shown in the earlier examples. In part (A) of figure 11-212, the primaries of reference transformer T1 and plate-supply transformer T5 are connected to a three-phase power source, so that the two voltages are 120 degrees out of phase. Control of the thyratron firing point is accomplished by combining the reference voltage with the error voltage, which is fed into transformer T2. The phase-shift amplifier illustrated is set up so that the following phase conditions exist: (1) the voltages at the plates of V1 and V2 are out of phase; (2) the error voltages at the grids of V1 and V2 are in phase; and (3) the voltages at the grids of V1 and V2 are out of phase with the reference voltage. For explanation purposes, it is assumed for the first example that the error voltage, E_s, is zero. Therefore since there is no E_s component, E_g in this case must be E_r, which is known to be 120 degrees out of phase with E_p. The curves for this condition, part (B) of the figure show that both V1 and V2 conduct equally during their respective positive plate alternations. Thus, equal and opposing current pulses are applied to the armature, and there is no rotation.

11-1367. In the next example, an error voltage, E_s, in phase with E_{p1}, is introduced. Part (C) of figure 11-212 shows that for V1, the out-of-phase summation of E_s and E_r acts to decrease the phase lag of the effective grid voltage, E_{g1}, causing V1 to fire early in the positive alternation of E_{p1}. Conversely, the in-phase summation of E_s and E_r for the V2 thyratron shows that the phase lag of grid signal E_{g2} is increased so that V2 fires very late in its conducting half-cycle. Thus, the heavier V1 current predominates, flowing through the armature from Y to X.

11-1368. When the error-voltage phase is shifted 180 degrees, so that it is in phase with E_{p2}, the conditions in part (D) of figure 11-212 prevail, causing the servomotor to reverse its direction of rotation.

11-1369. For the phase-shift type of operation, it can be seen that the amount of phase shift obtained is dependent upon the magnitude of the error signal, E_s. The greater the E_s magnitude, the greater the inequality in the V1 and V2 average currents. As the error voltage falls toward zero, the thyratron currents approach the balanced condition for which the servomotor comes to rest.

11-1370. ANTI-HUNT CIRCUIT. The basic circuit of figure 11-209 may be modified to include an anti-hunt circuit, as shown in figure 11-213. In the no-error condition, the fixed grid bias is adjusted by means of R3 so that V1 and V2 cannot fire. When an error voltage appears which bears a phase relationship to the reference voltage as

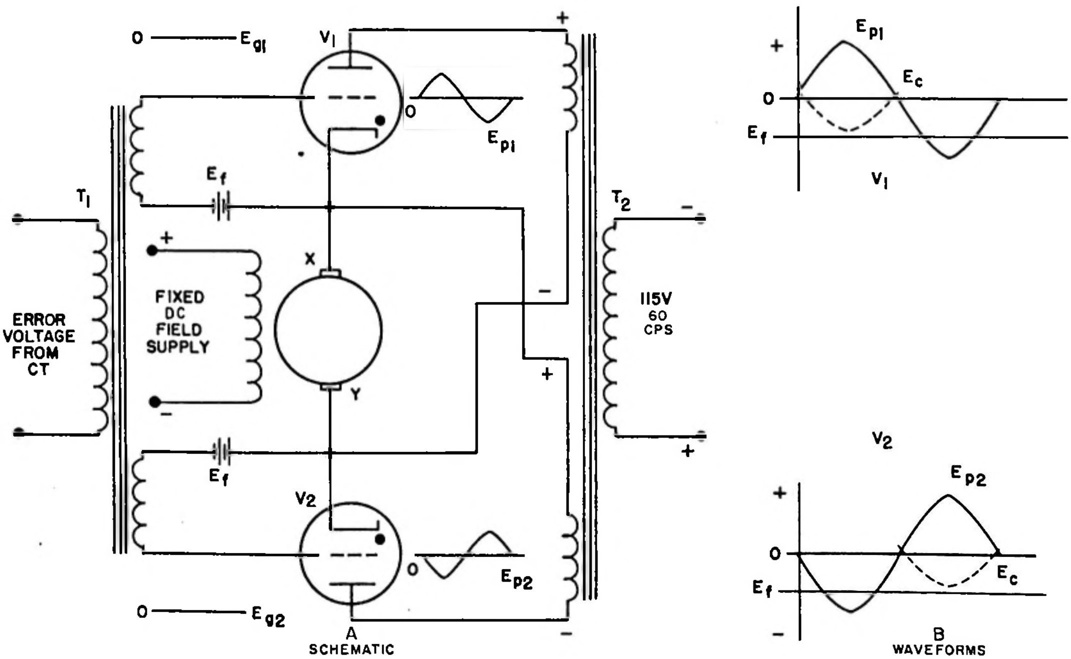
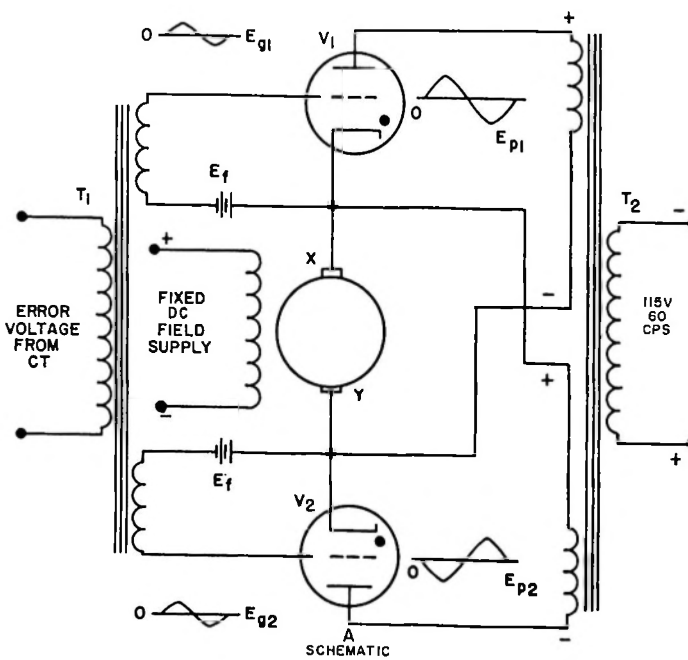


Figure 11-209. Thyratron Motor Control Circuit with Zero Error Signal

indicated by the voltage polarity symbols, V1 conducts. Current flows through the dc motor armature in the direction indicated by the arrows, causing armature rotation. In the rotating state, the armature tends to act as a generator, and develops a counter electromotive force with a polarity as indicated. Thus, the current produced by the counter emf opposes the applied current. Since the series combination of R1 and R2 is across the armature, the voltage across R1 tends to increase the negative bias on V1, thus providing for degenerative feedback. This action tends to stabilize the armature current to a substantially constant value. As the error voltage decreases, V1, being biased by the fixed supply plus the drop across R1, is cut off just before the load reaches the ordered position. At this instant, V2, which is biased by the fixed supply minus the drop across R2, conducts momentarily. The reversed current to the armature acts as a brake, which quickly checks the load motion. Thus, by means of proper circuit constants and accurate adjustment, the device functions to eliminate overshoot and hunting.

11-1371. THYRATRON CONTROL OF SPLIT-FIELD DC SERVOMOTOR. For servo circuits using the split-field type of dc servomotors, a thyratron arrangement similar to that of figure 11-214 may be used. In the simple circuit shown, the voltages at the grids of V1 and V2 are in phase with the error voltage, while the voltages at the plates of the tubes are out of phase with the reference voltage. With no error voltage, V1 and V2 conduct alternately and equally on their positive half-cycles, sending current pulses first from X to Y, and then from Z to Y, through the split field and armature



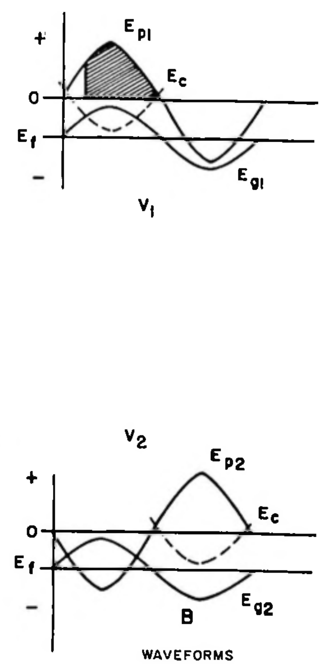


Figure 11-210. Thyratron Motor Control Circuit with Error Signal in Phase with E_{p1}

windings of the servomotor. The net current is zero, and there is no rotation. For an error voltage in phase with E_{p1}, V1 conducts heavily when the plate and grid are positive. On the following alternation, when the V2 plate voltage swings positive, its grid signal swings negative. Thus, V2 either does not conduct at all (large error voltage), or conducts a small amount (small error voltage). The current flow for this condition is seen to be from X to Y. When the error voltage reverses, so as to be in phase with E_{p2}, the opposite set of conditions is established.

11-1372. THYRATRON SERVO AMPLIFIER FOR AC SERVOMOTOR CONTROL. There are several circuit variations in which the thyratron is used to control the operation of an ac servomotor. A representative servo amplifier for this purpose is diagrammed in figure 11-215. T3 acts as the plate-supply and reference-voltage transformer, with the secondary windings arranged so that the voltages at the V1-V4 plates are in phase, as are the voltages at the V2-V3 plates. The error voltage is applied to input transformer T1. T2 is the output transformer to the controlled-phase winding of the servomotor. If no error signal is present, none of the thyratrons fires, because the negative dc grid-bias level is such that it does not intersect the critical grid bias curve. If an error voltage appears, with an instantaneous positive polarity at the top of the T1 secondary at the same time the V1 plate swings positive, V1 fires. V2 cannot fire because its plate is negative, and V4, having an additional negative bias, remains cut off. As long as the error voltage maintains this phase relationship, V2 and V4 cannot fire. On the first alternation, then, current flows from X to Y through the output transformer. On the following

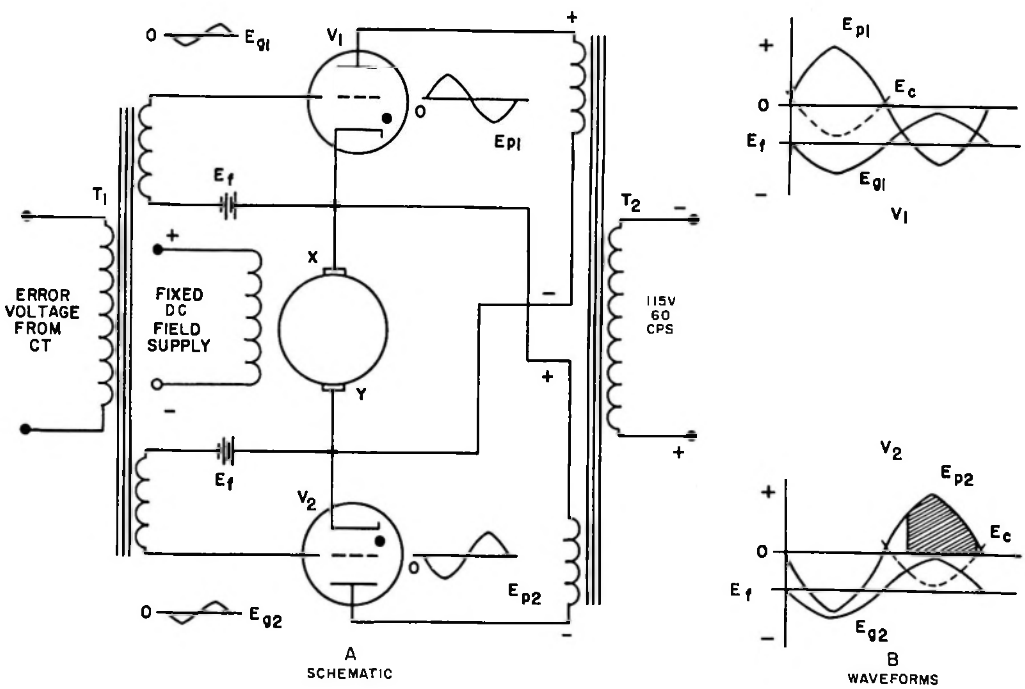


Figure 11-211. Thyratron Motor Control Circuit with Error Signal in Phase with E_{p2}

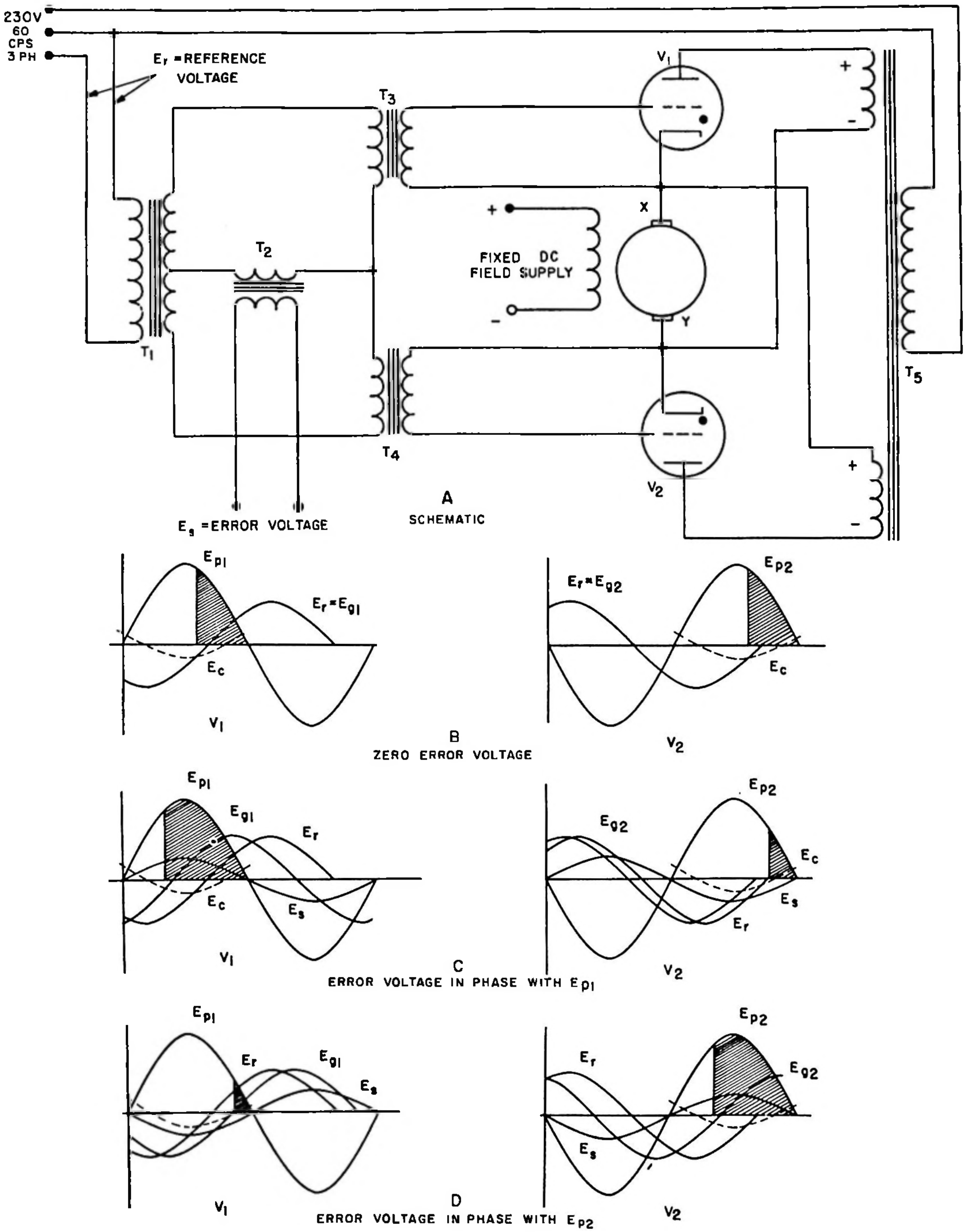


Figure 11-212. **Thyratron Phase-Shift Motor Control Circuit**

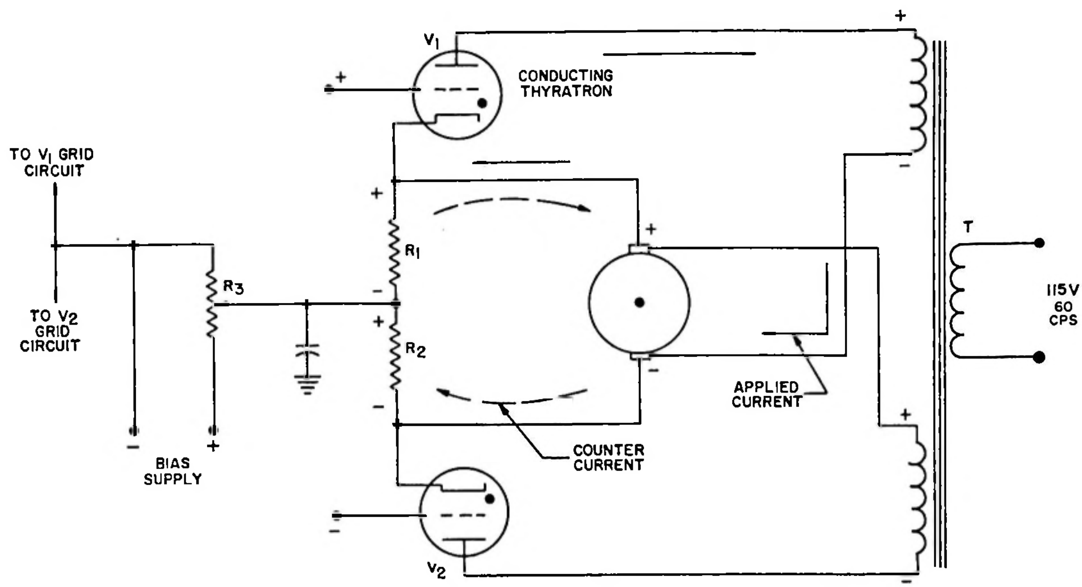


Figure 11-213. Anti-Hunt Circuit

alternation, both the grid and plate of V3 swing positive, and V3 fires, with plate-current flow from Z to Y in T2. Thus, V1 and V3 conduct on alternate half-cycles, causing an ac voltage to be induced into the T2 secondary. This voltage may be either in phase or out of phase with the reference voltage. The servomotor now turns in the ordered direction. Reversal of the error voltage phase causes V2 and

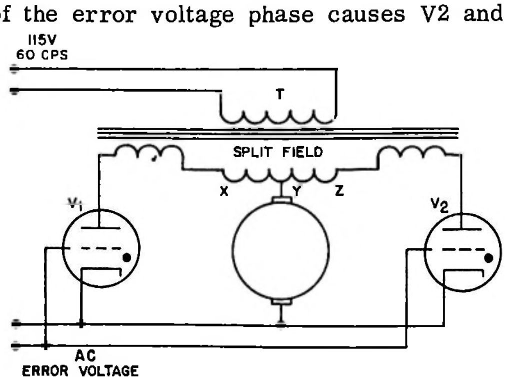


Figure 11-214. Thyratron Amplitude Control of Split-Field DC Servomotor

V4 to become the conducting thyratrons, and shifts the controlled phase 180 degrees with respect to the reference voltage. Hence, the servomotor reverses its direction of rotation.

11-1373. THYRATRON-SATURABLE REACTOR CONTROL. Another circuit for the control of ac servomotors is illustrated in figure 11-216. Two thyratrons are used here in conjunction with a pair of saturable reactors. The phasing requirements in this example are such that the V1-V2 grids are in phase with the error signal, while the V1-V2 plates are out of phase with the plate-supply voltage. One side of the servomotor control field is connected to the center tap of the T3 secondary. The other terminal connects to both ends of the T3 secondary through the two saturable reactor secondary windings, X1 and X2, as shown. In the no-error condition, V1 and V2 conduct equally on their alternate positive half-cycles, and the reactors are balanced. Thus, points B and D are at the same potential, and the servomotor control field is not energized. When an error voltage is introduced which is positive when the V1 plate is positive, V1 conducts heavily, and V2 does not conduct at all, because its plate is negative. Consequently, reactor X1 is saturated by the dc output of V1, reducing the inductance of the X1 secondary winding to a low value. Effectively, therefore, point D is connected to point A through a low reactance, and the motor control field is energized in one direction. When the error signal from T1 is reversed in phase, T2 fires on its positive alternations, saturating X2. Point D may now be considered as connected to point C, reversing the servomotor direction of rotation.

11-1374. C1 and C2 are shunted across the primary windings of X1 and X2, respectively, and capacitance values are chosen so as to maintain the flow of direct current in the control winding of the saturable reactor. Thus, if V1 is the controlling thyratron, C1 acts to supply current to the X1 primary when the thyratron is in its nonconducting half-cycle, and to prevent voltage surges during the conducting half-cycle. This circuit is of particular interest, since it lends itself to a frequency flexibility not usually encountered. Examination of the circuit diagram shows that the error and reference voltages may be of one frequency, while the servomotor and reactors may be supplied by a voltage of entirely different frequency, without affecting the operation of the control circuit.

11-1375. AMPLIDYNE CONTROL METHOD.

11-1376. The amplidyne motor-generator consists of a constant-speed ac drive motor and a two-stage electromechanical power

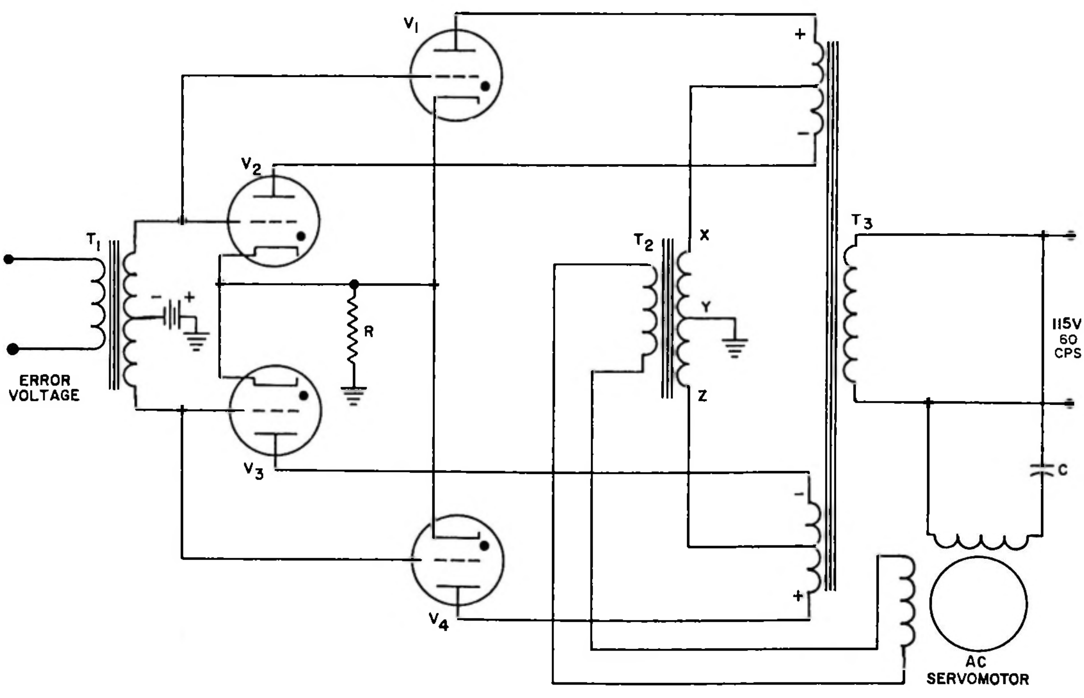


Figure 11-215. Thyratron Control for AC Servomotor

amplifier, contained in a single housing. The drive motor, which may be of the squirrel-cage induction type, has its rotor shaft coupled to the armature of the generator section. Since this motor drive mechanism is similar to that of other servo drive mechanisms, it can be considered as conventional and, therefore, self-explanatory. The amplidyne section, however, is radically different from the conventional generator in the unusual method employed to obtain high power amplification. The step-by-step development of the amplidyne principle is illustrated in figures 11-217 through 11-221.

11-1377. A cross section of a conventional dc generator is diagrammed in figure 11-217. In this representation, a load is shown drawing a current of 60 amperes from the generator armature. In order to meet this demand, the armature must have induced in it a voltage of sufficient magnitude to produce the necessary current. To provide the proper flux density to produce this current, a field excitation current of three amperes was found necessary. The generator may now be considered as an amplifier with a current gain of 20. Since the direction of the excitation flux, ϕ_e, is from north to south, the flux will pass through the armature core in a horizontal direction, as indicated by the arrows. When the external armature circuit is completed, causing a current flow of 60 amperes, the armature, being wound

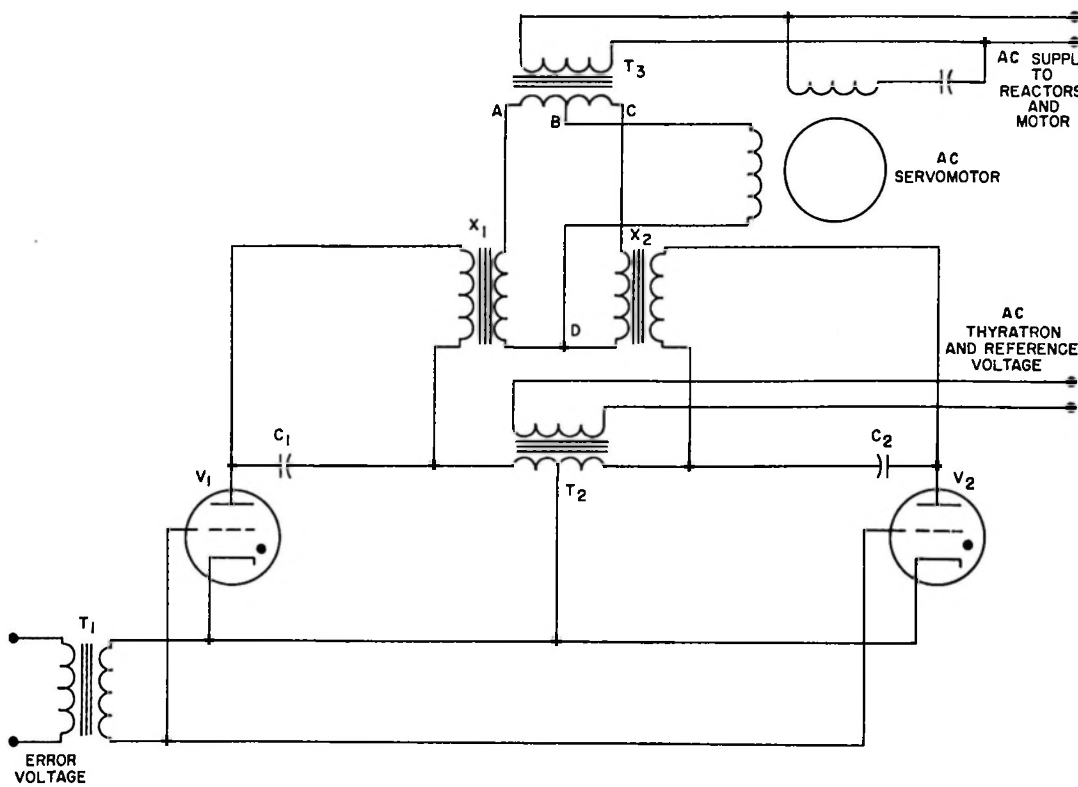


Figure 11-216. Thyratron Saturable Reactor Control for AC Servomotor

on an iron core, acts as an electromagnet. This action gives rise to an armature reaction flux, designated in the diagram as ϕ_a. If the left-hand rule is applied to the current flow in the armature conductors, the armature flux, ϕ_a, is shown to be at right angles to the excitation flux, ϕ_e. A simplified version of the circuit is shown in part (B) of figure 11-217 with the flux directions and magnitudes indicated.

11-1378. Figure 11-218 is the same as figure 11-217 except that, in this case, the load has been removed and the armature leads have been short-circuited. Since the resistance of the load is no longer a factor, the only opposition offered to current flow is the low resistance of the armature windings, plus the negligible resistance of the short-circuit wiring. The immediate result of such a short-circuiting procedure would be to increase the armature current to an abnormally high value. The end result would be a burned-out armature. However, one way of reducing the enormous armature current would be to reduce the excitation flux to a much lower level. It is apparent that this flux could be made weak enough so that the short-circuit current in the armature circuit could be reduced to 60 amperes, which it was in the previous example. Moreover, since the

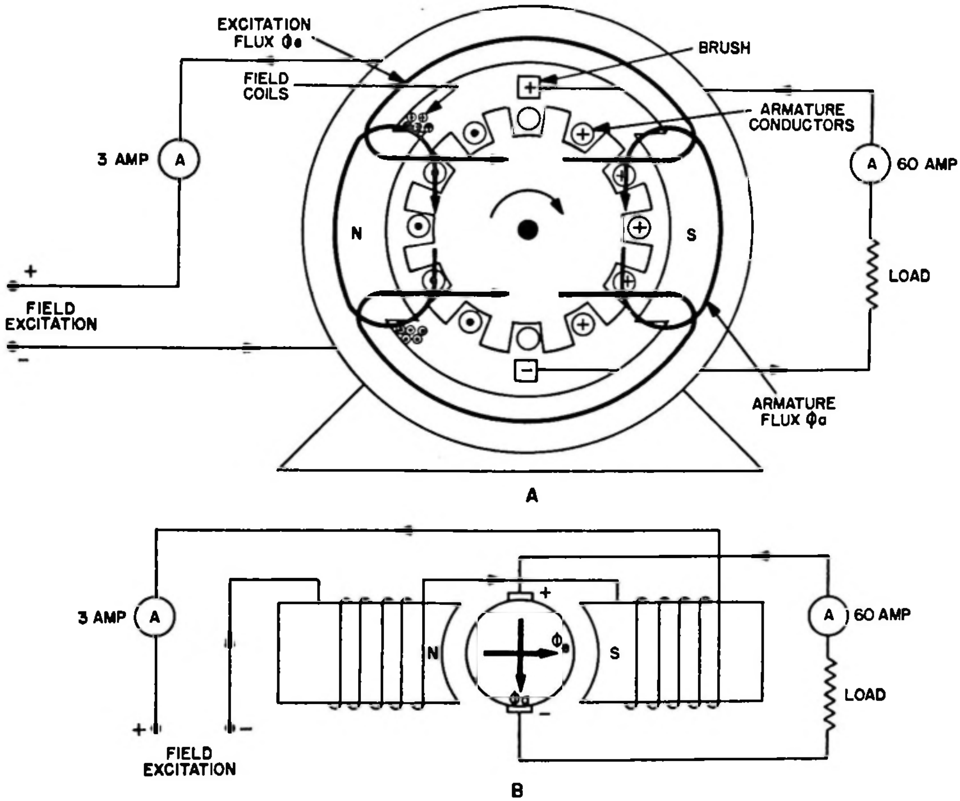


Figure 11-217. Magnetic Field and Current Relationship in Conventional DC Generator

armature easily handled a 60-ampere drain in the loaded condition, a short-circuit current of the same value cannot cause damage to it. A reduction in field excitation current will weaken the flux to the proper strength; in this case, the current has been dropped to 0.03 ampere. Consequently, it may be seen that a field current of 0.03 ampere controls a short-circuited armature current of 60 amperes, whereas it took 3 amperes to control the same amount of output current in the loaded state. It is clear, then, that the generator current gain has been increased to 2000.

11-1379. The problem now arises as to how the increased current gain can be put to use. Obviously, the load cannot be placed in series with the short circuit, since this would mean a return to the original status. The short circuit, therefore, must remain intact. From part (B) of figure 11-218, it is evident that two fluxes exist—a weak field flux, ϕ_e, and a strong armature flux, ϕ_a, the latter created by a heavy current of 60 amperes. The cross sections show that the armature conductors are evenly spaced around the core; therefore, the conductors will cut the heavy armature flux, ϕ_a, at the same rate as they cut the excitation flux, ϕ_e. But the maximum voltage induced in the conductors as they cut the armature flux appears across the armature at right angles to the voltage induced by the excitation flux.

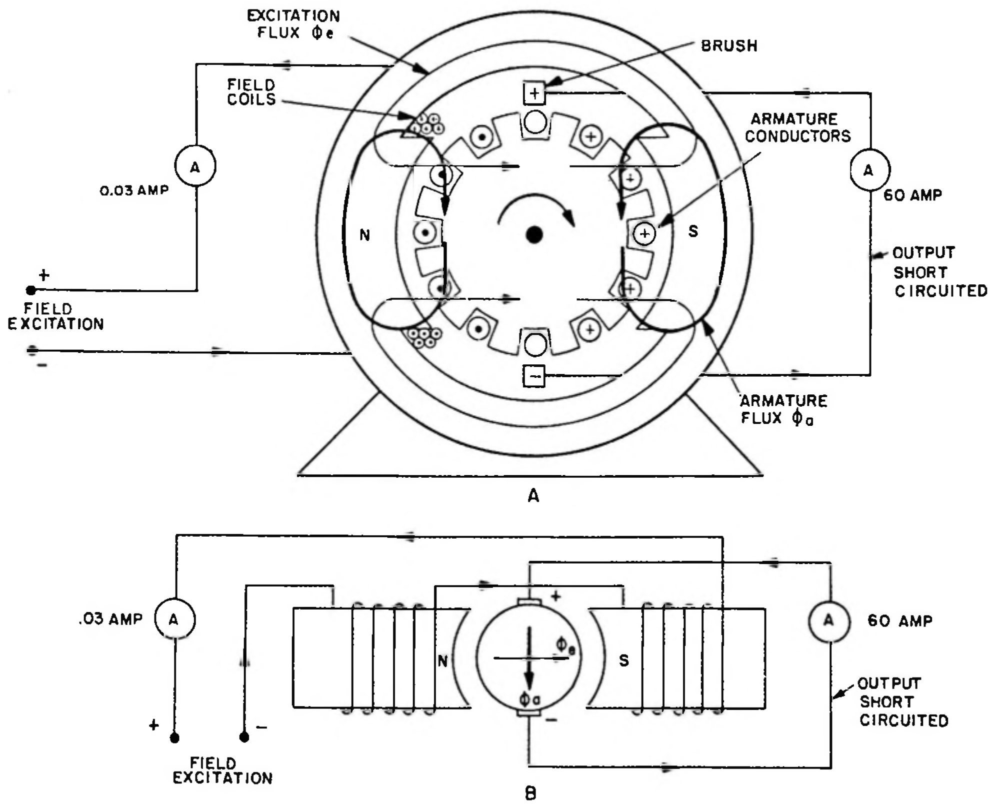


Figure 11-218. Magnetic Field and Current Relationship in Short-Circuited DC Generator

To take advantage of this new voltage, caused by the fact that the armature conductors cut their own reaction flux, a second pair of brushes (figure 11-219) is placed on the commutator at right angles to the short-circuited brushes, and connected to the load. Since a high voltage is realized as the armature conductors cut the strong reaction flux, ϕ_a, the voltage developed across the output brushes is sufficient to supply a large current, e.g., 60 amperes, to the load, despite the resistance of the load circuit.

11-1380. Another problem is encountered here, as shown by the diagram. Just as the armature current in the short-circuited section creates a flux at 90 degrees to the excitation flux, so does the current in the output circuit set up a flux at 90 degrees to the armature flux. This new reaction flux, ϕ_b, is therefore removed 180 degrees from the original excitation flux, ϕ_e. Moreover, reaction flux ϕ_b is very much stronger than excitation flux, ϕ_e, and, since it is in opposition to ϕ_e, the excitation flux could no longer control the output. To overcome this obstacle, a compensating winding is placed on the field pole pieces, and is connected in series with the output to the load. In design considerations, the number of turns in this compensating winding is calculated

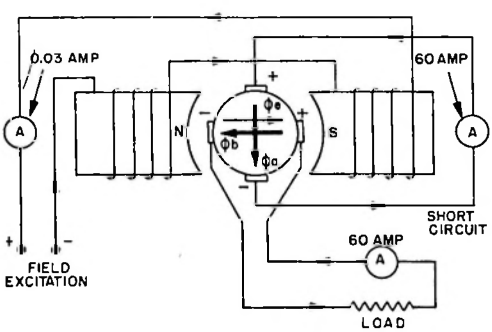


Figure 11-219. Short-Circuited DC Generator Supplied with Additional Brushes

and the direction of current flow determined so that a compensating flux, ϕ_c, exactly equal and opposite to flux ϕ_b, is developed. The compensating winding and the four flux components are represented in figure 11-220. Inasmuch as ϕ_b and ϕ_c cancel each other, the resulting fluxes are ϕ_e and ϕ_a, as indicated in the amplidyne equivalent circuit of figure 11-221.

11-1381. The original build-up of a voltage in a conventional self-excited dc generator depends upon the presence of a certain amount of residual magnetism in the field pole pieces. Some residual magnetism remains in the amplidyne field poles after the field excitation current is reduced to zero, which is the proper value when the input and output shafts of the servomechanism are in correspondence. The presence of residual magnetism would cause a weak field flux, resulting in the induction of an appreciable voltage in the armature, because of the amplidyne's high gain. This, in turn, would have the undesirable effect of causing rotation of the servomotor where no error exists. In order to eliminate the residual magnetism, a small ac generator is used. This generator may consist of a permanent magnet mounted on the end of the amplidyne armature, and revolving in a small field winding. The output of the ac generator is applied to two sets of opposed windings placed on the field pole pieces. The effect of these windings is to neutralize the residual magnetism which exists when the field excitation current is zero.

11-1382. The amplidyne generator can be compared to a two-stage vacuum-tube amplifier. The development of the short-circuit current, with its accompanying armature reaction flux, by means of a small excitation current, constitutes the first state of amplification. The use of the strong flux developed by the short-circuit current, to produce a voltage high enough to cause an equally large current to flow through a load circuit, represents the second state of the system. The second reaction flux, ϕ_b, is analogous to negative feedback, or degeneration, in a tube circuit. The compensating winding can be regarded as a regenerative circuit, designed to exactly balance the degeneration so that the net feedback in the amplidyne is zero. Because of its power gain of approximately 10,000, the amplidyne has extensive use in servomechanisms.

11-1383. HYDRAULIC CONTROL METHOD.

11-1384. GENERAL.

11-1385. Hydraulic servomechanisms are used in some radar equipments for antenna positioning of guns and other ordnance equipment. Smaller hydraulic servos find extensive use in specialized military control equipment. The hydraulic servomechanism is chosen for many applications where a rapid response, combined with smooth operation, is required. For purposes of explanation, the hydraulic servo components mentioned in the following description are simplified. In actual practice such devices embody many complex refinements. Some of these include dither mechanisms, error correctors, coarse-fine data assemblies with accompanying contact-ring-relay transfer arrangements, limit switches, etc.

11-1386. VARIABLE-FLOW PUMP.

11-1387. Figure 11-222 shows a type of variable-flow pump similar to that illustrated in the block diagram of figure 11-223. Figure 11-222 is a cutaway side view of pump mechanism A, a bottom view of cylinder-piston assembly B, and a bottom view of valve plate C. The solid-line portion of A shows the pump housing depressed 30 degrees below the drive shaft. In the dotted-line position, the housing is raised 30 degrees above the shaft axis. Only two pistons and cylinders are indicated in the side view of the sketch. However, there are actually six pistons (in this example), equally spaced in the cylinder block, as shown in B.

11-1388. In analyzing the operation of the pump, it is assumed that the ac motor is turning the circular drive plate and the rotating cylinder assembly in the clockwise direction, as indicated by the arrow. At the instant shown, piston 1 is at the top of its

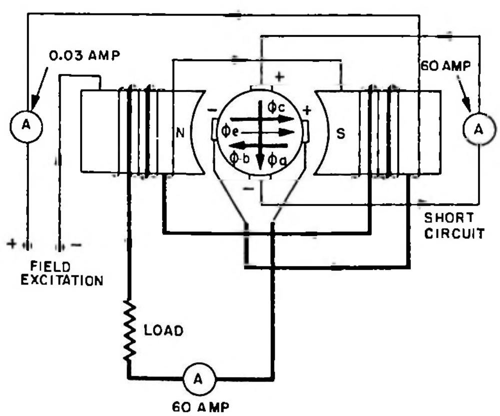


Figure 11-220. Short-Circuited DC Generator with Additional Brushes and Compensating Windings

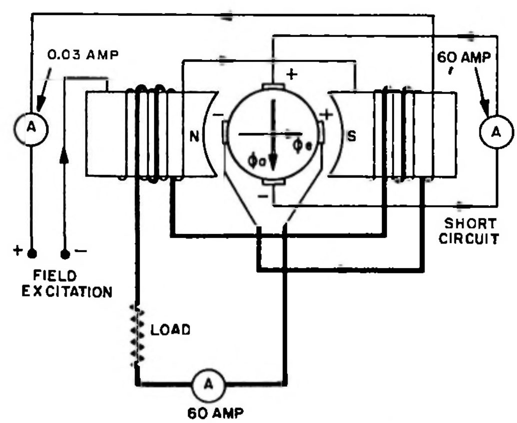


Figure 11-221. Amplidyne Generator Equivalent Circuit, Showing Effective Magnetic Field and Amplification

stroke and its cylinder is filled with oil, while piston 4 is at the bottom of its stroke, having already pumped its store of oil. By referring to drawing-plate projections B and C, it will be noted that neither piston is pumping at this instant. However, pistons 5 and 6 are over port Y, and, since these pistons are moving downward and pushing into their cylinders, oil is pumped into port Y and out of oil line Y. At the same time, pistons 2 and 3, which are aligned with oil port X, are moving upward and pulling out of their cylinders, thereby sucking oil from oil line X into cylinders 2 and 3 through oil port X. In the solid-line pump position, then, Y is the output oil line and X is the input line.

11-1389. When the control arm pulls the mechanism through the zero-degree position into the top 30-degree position (dotted lines), the pumping action takes place as the pistons move upward. Thus, as the cylinder assembly rotates, the oil is pumped into oil port X. In this condition, X becomes the output oil line, oil is brought into the pump through input line Y, and the direction of oil flow is reversed.

11-1390. When the pump housing is set in the neutral position, so that it forms no angle with the drive shaft, it is apparent that the cylinder assembly rotates without piston action, since each of the pistons is in the center of its cylinder. Without angular displacement, then, there is no in-and-out, or pumping, action of the pistons. A condition now exists where each piston cylinder is partially filled with oil, but there is no oil flow, either in or out, through X or Y. It must be stressed that the ac motor continues to turn the cylinder and piston assembly at all times, even when no pumping action takes place. The ac motor turns the variable-flow pump at a constant speed in one direction only. The amount of oil flow

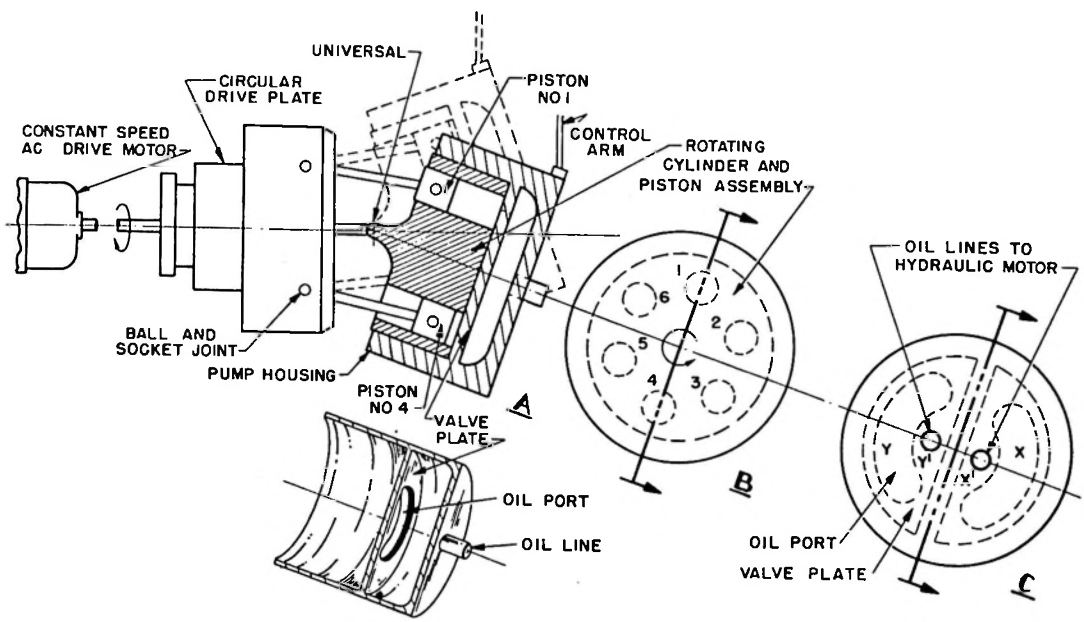


Figure 11-222. Hydraulic Variable-Flow Pump

is controlled by the angular tilt of the pump housing, maximum flow being realized at 30 degrees. The direction of oil flow may be changed by reversing the pump-housing angle.

11-1391. HYDRAULIC MOTOR.

11-1392. The hydraulic motor is a mechanism exactly like that of the variable-flow pump, except that the pump housing is held fixed at a 30-degree angle to the shaft axis. In the following analysis of the hydraulic-motor action, the cylinder and piston assembly is assumed to be in the position indicated in figure 11-224. Here again, B represents a bottom view of the rotating assembly, and C is a bottom view of the

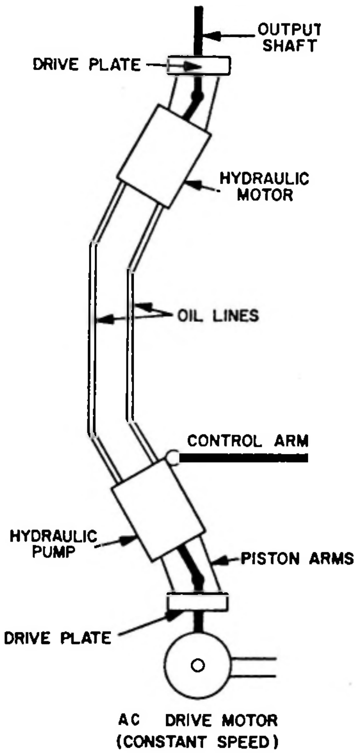


Figure 11-223. Basic Hydraulic Servomechanism

rotating assembly and the valve plate. If oil is pumped into the motor through oil line X and oil port X, pressure is exerted on pistons 2 and 3, which, in turn, apply an upward force on the edge of the circular drive plate toward the reader. A clockwise rotation of the load results, as viewed from the end of the output shaft. Oil returns to the variable-flow pump through oil line Y. When the pump angle is reversed, so that oil is pumped through Y into the motor, pistons 5 and 6, which are aligned with port Y, are under pressure. An upward force is again exerted on the drive plate, but, since the force is now applied to the edge away from the reader, the load rotates in a counterclockwise direction.

11-1393. OIL PRESSURE.

11-1394. In the variable-speed transmission unit just described, the oil in the unit must be held under a certain pressure. This is accomplished by the combination of an oil-replenisher pump and valves which maintain an oil-supply pressure of approximately 75 psi to the transmission unit. The maximum hydraulic pressure in the unit, controlled by a high-pressure relief valve, is approximately 1400 psi.

11-1395. ERROR MEASUREMENT.

11-1396. It has been shown that the pump is in the neutral, or zero-degree, position when the gun is aimed according to the director order and a no-error condition exists. Furthermore, it is obvious that the angle assumed by the pump housing is a function of the control arm through an error-measuring device. As in the previous instance, there are several error-measuring devices that may be used. In one method the position of the gun is ordered by a synchro transmitter, located at the gun director and wired to a synchro control transformer at the gun. Completing this type of loop, the shaft of the synchro control transformer is geared to the gun, and the rotor winding feeds the error voltage to a servo amplifier. The amplifier then energizes a servomotor, which actuates the pump-housing control arm by means of mechanical or hydraulic linkage. In the variation of this error circuit, a synchro differential is used as an error detector, supplying enough torque to mechanically position the input member of a hydraulic-booster arrangement. The output piston of the booster is then used to move the arm which controls the angular position of the pump housing. In this circuit, the differential receives its rotor and stator voltages from the synchro transmitters at the director and gun, respectively.

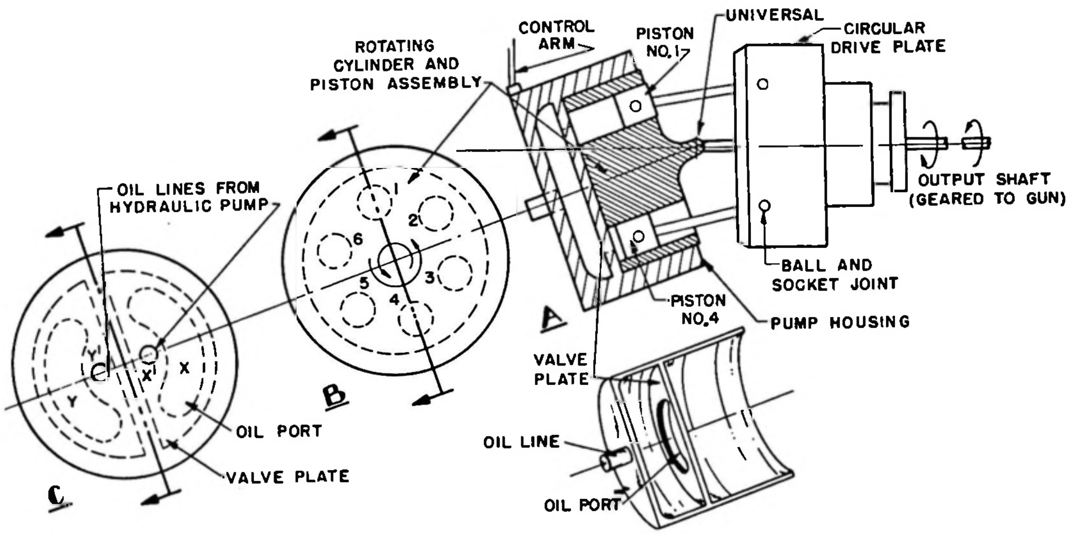


Figure 11-224. Hydraulic Motor

SECTION VIII

POWER SUPPLY TESTING

11-1397. GENERAL.

11-1398. Power supplies used to provide the dc voltage sources for operation of electronic equipments are of various types, such as half-wave, full-wave, and bridge circuit rectifiers. The type employed for a particular application depends on such factors as current and voltage load requirements, available space, weight, degree of voltage regulation required, etc. Electronic systems are generally complex, and contain one or more power supplies which provide several voltage outputs. The design performance of these equipments depends upon the proper operation of the power supplies. For this reason, power supply tests are included in the preventive maintenance test procedures prepared for most equipments. Many equipments are provided with easily accessible test points for ease in performing these tests. The most common type of power supply test is a simple voltage reading taken under load (with the equipment in operation). When a power supply filter is suspected, the ripple content may be observed to see that it is within the percentage of tolerance.

11-1399. Power supplies operate either from an ac or a dc power source. Rectifiers, usually in the form of vacuum tubes, gas tubes, or semiconductors, are used to convert the ac input voltage to a pulsating dc voltage. Semiconductor rectifiers are frequently used in power supply circuits. These devices use the rectifying characteristics of substances such as copper oxide, copper sulphide, selenium, germanium, silicon, etc. Semiconductor rectifiers have the advantages of light weight, high efficiency, ruggedness, and long life. The pulsating dc voltage is smoothed to a steady dc voltage by means of a suitably designed filter. Most power-supply circuits employ a power transformer, thereby allowing a step-up or step-down in the ac voltage in order to produce the desired dc output voltage. In some power supplies, however, where weight, space, and cost are important considerations, the transformer is omitted and the rectifier connected directly to the ac source. Rectifiers designed to furnish power greater than 1 kilowatt generally use three-phase ac power and a polyphase arrangement of tubes. Such multiphase rectifiers produce a smoother output waveform than single-phase circuits, because the ripple frequency is higher and, therefore, relatively easy to filter.

11-1400. HALF-WAVE, FULL-WAVE, AND BRIDGE CIRCUIT RECTIFIERS.

11-1401. HALF-WAVE POWER SUPPLY.

11-1402. Half-wave power supplies are the simplest of the various types of power supplies. The half-wave power supply uses a single rectifier (electron tube shown in part A of figure 11-225, and semiconductor rectifier shown in part B of the figure, which conducts essentially in only one direction. The rectifier conducts once during each cycle of the input voltage (during the interval when the plate is positive with respect to the cathode). A pulsating dc voltage is produced, and the pulsation frequency is the same as the frequency of the input voltage. Since the amplitude of the rectified voltage remains at zero for an appreciable interval between successive dc pulses, the output of a single-phase half-wave rectifier is difficult to filter, and requires a good filter network to produce a steady dc output voltage (with little ripple). This type of circuit has the additional disadvantage that the current flowing in one direction through the secondary magnetizes the core of the transformer. Because of these disadvantages, the half-wave power supply is rarely used. The circuit shown in part A of the figure employs a power transformer with two secondary windings; the low-voltage winding, S_2, supplies the correct filament voltage to the rectifier tube.

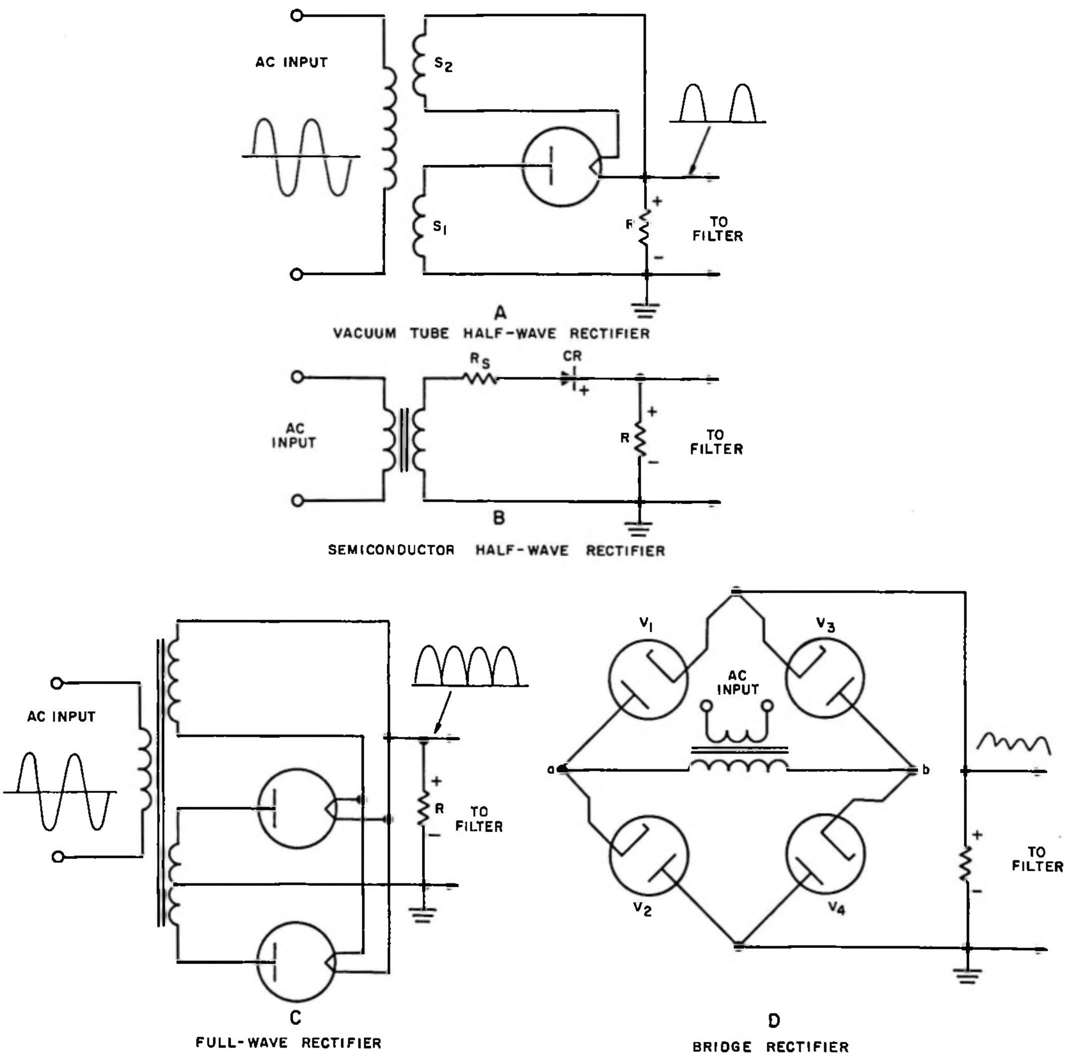


Figure 11-225. Basic Power Supply Circuits

11-1403. Part B of figure 11-225 shows a resistor, R_S, in series with the rectifier. This resistor, called the <u>surge</u> <u>resistor</u>, serves to limit the peak current that flows through the rectifier, and is composed of the transformer resistance and any necessary external resistance. When the circuit is first energized, the input capacitor in the filter circuit is in a discharged condition. Thus, if power were applied with no resistor in the circuit, a very heavy current would flow to establish the charge on the capacitor (this is not the case in electron-tube rectifiers because of the time required to heat the cathode and the relatively high internal resistance). Since this current might damage the rectifier, R_S is included. For 380-volt (peak inverse) rectifiers, the value of R_S is between 1 and 50 ohms, depending upon the peak current rating of the unit.

11-1404. FULL-WAVE POWER SUPPLY.

11-1405. A full-wave power supply, shown in part C of figure 11-225, produces an output which is easier to filter and which has a higher average value than that of a half-wave power supply. However, this circuit requires two rectifiers (commonly in one envelope), and the transformer must have a center-tapped, high-voltage secondary with a step-up voltage approximately equal to twice the desired dc voltage output. The plate of one rectifier is connected to one side of the high-voltage secondary winding, and the plate of the other rectifier is connected to the other side.

11-1406. The voltage induced in each half of the secondary causes each tube to conduct on alternate half-cycles of the input voltage. Hence, two dc pulses (of the same polarity) occur during each cycle of the ac input voltage, thereby producing an output which has a fundamental frequency which is twice that of the input frequency. Since the dc pulses in the output of a full-wave rectifier are more closely spaced in time than those obtained from a half-wave circuit, the full-wave output is less difficult to filter. In other words, for the same type of filter, the full-wave circuit has a lower percentage of ripple than the half-wave circuit. In addition, dc saturation is not present in the core of the transformer, because the dc magnetization in the two halves of the transformer secondary are opposed to each other and therefore cancel the magnetizing effects.

11-1407. BRIDGE-TYPE POWER SUPPLY.

11-1408. A bridge-type power supply, shown in part D of figure 11-225, produces an output similar to that of the conventional full-wave circuit (the ripple frequency is twice the input frequency). In the bridge circuit, four rectifier elements are needed; however, the power-transformer secondary need not be center-tapped.

11-1409. During each half-cycle of the ac voltage that appears across the secondary, two tubes conduct in series, and produce one dc pulse in the output. During the interval that point "a" is positive with respect to point "b", V_1 and V_4 conduct, and produce an output pulse. Similarly, during the next half-cycle, when point "b" is positive with respect to point "a", V_2 and V_3 conduct, and produce another output pulse. Two dc output pulses are therefore produced for each input cycle (full-wave output). In a bridge-type rectifier, the peak inverse voltage appearing across each tube does not exceed the peak transformer voltage. Hence, the bridge-type circuit can be used to obtain a higher output voltage than a center-tapped, full-wave rectifier using equivalent rectifiers. This circuit is widely used with silicon and copper-oxide rectifiers, and less often with electron tube rectifiers. This is because the cathodes are at different potentials, and thus cannot be connected in parallel and supplied from a single filament transformer winding.

11-1410. VOLTAGE MULTIPLIERS.

11-1411. GENERAL.

11-1412. Voltage multipliers are used to produce a higher dc output voltage than can be obtained from a conventional rectifier; they are generally used in circuits requiring low current, and also when it is desired not to use a power transformer. Because of its ability to produce high voltage without the use of a power transformer, the voltage-multiplier circuit is frequently used in high-voltage applications in which the cost of construction, weight, and size of the power supply must be held to the minimum. If used with a transformer, it can be designed to provide a much higher voltage than is possible with a conventional power supply.

11-1413. FULL-WAVE VOLTAGE DOUBLER.

11-1414. This circuit, shown in part A of figure 11-226, uses two rectifiers, each of which conducts on alternate half-cycles of the input voltage. On the positive half-cycles of the input voltage, V_1 conducts and charges capacitor C_1 to the peak value of the input voltage, with the polarity as shown. Likewise, on the negative half-cycles of the input voltage, V_2 conducts and charges capacitor C_2 to the peak value, with the polarity as indicated. The polarities of the charges on the two capacitors are such that the voltages add, and produce an output equal to their sum. Therefore, the voltage available at the output is approximately double that for a half-wave circuit.

11-1415. CASCADE VOLTAGE DOUBLER.

11-1416. The operation of the cascade voltage doubler, shown in part B of figure 11-226, is slightly different from that of the full-wave doubler, shown in part A of the figure. During the negative half-cycle of the input voltage, V_2 conducts and charges C_1 to the peak value of the input voltage, with the polarity as indicated. The voltage at point X will then consist of the ac input in series with the charge on C_1, and will vary in amplitude from zero to twice the peak value of the ac input voltage. On the positive half-cycle of the input voltage, twice the peak value of the ac input voltage is effectively applied to V_1, causing it to conduct and charge C_2 to twice the peak value of the input voltage. The cascade voltage doubler has the advantages that no transformer is required and that there is a common connection between the input and output so that they may both be grounded.

11-1417. VOLTAGE TRIPLER.

11-1418. The voltage tripler produces an output voltage approximately equal to three times that produced by an equivalent half-wave power supply. This circuit, shown in part C of figure 11-226, consists essentially of a cascade voltage doubler whose output is in series with the output of a half-wave rectifier. The portion of the circuit that includes CR_1 and CR_2 is a cascade voltage doubler, which charges C_2 to produce an output equal to twice the peak value of the input voltage. The portion of the circuit that includes CR_3 is a conventional half-wave circuit, which charges C_3 to the peak input value. Since C_2 and C_3 are connected in series (aiding), the output voltage is equal to the sum of the two capacitor voltages, and is three times the peak input value. The voltage regulation of this circuit is poorer than that of a voltage doubler if the same filter capacitor values are used.

11-1419. VOLTAGE QUADRUPLER.

11-1420. The voltage quadrupler, shown in part D of figure 11-226, produces an output which is four times the peak value of the input voltage. This circuit consists of two cascade voltage doublers whose outputs are in series. (Refer to paragraph 11-1416.) When V_4 conducts, capacitor C_1 charges to the peak input value. This charge on C_1, combined with the input voltage, causes twice the peak value of voltage to be applied to V_3, which conducts and charges C_4 to this value. The conduction of V_2 charges C_2 to the same value. The voltage on C_2 is then applied to V_1, which conducts and charges C_3 to twice the peak value of the input voltage. Since C_3 and C_4 are in series, the voltages across them add to produce the output voltage, which is equal to four times the peak value of the input.

11-1421. MULTIPHASE RECTIFIERS.

11-1422. THREE-PHASE HALF-WAVE POWER SUPPLY.

11-1423. The three-phase half-wave rectifier, shown in part A of figure 11-227, uses a power transformer with y-connected

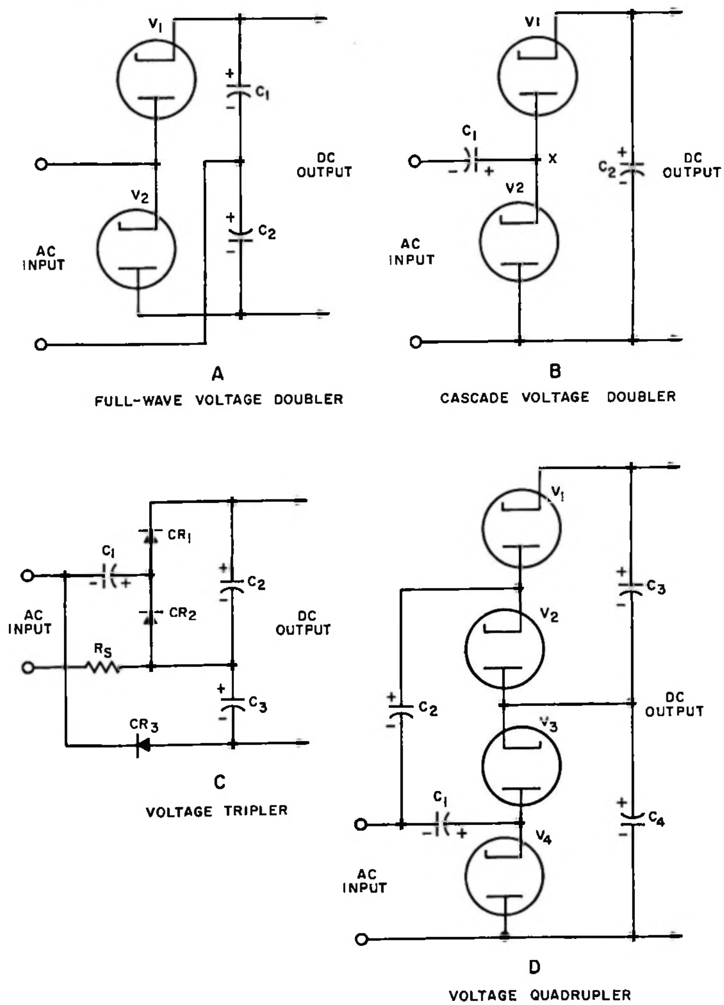
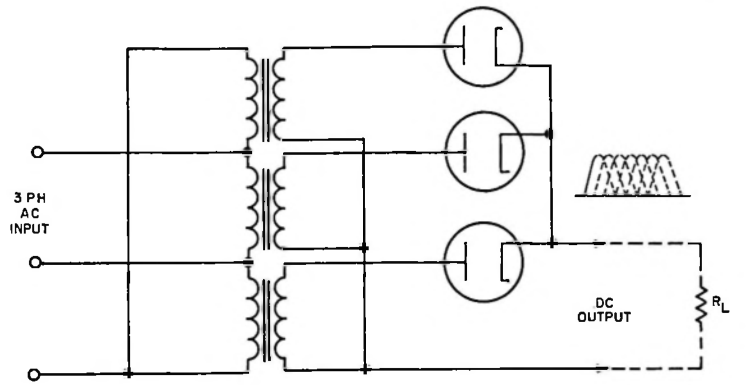


Figure 11-226. Voltage-Multiplier Circuits

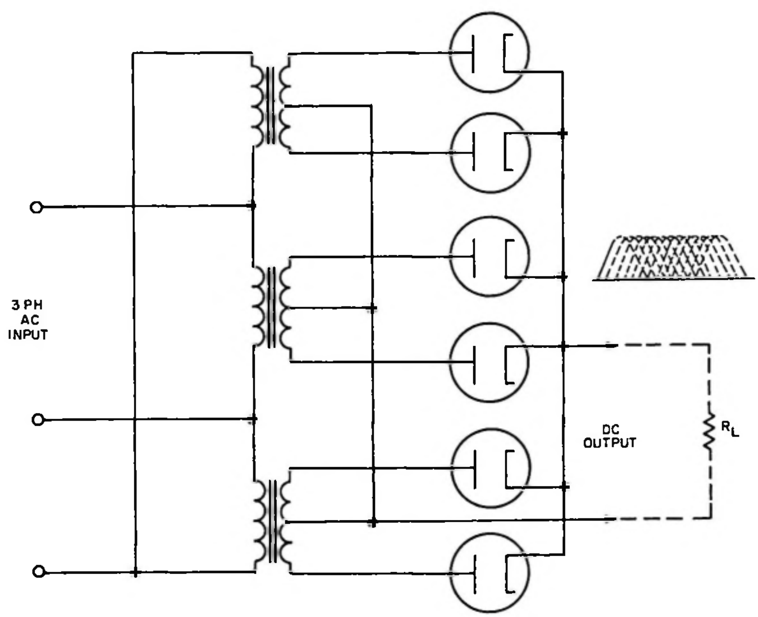
secondaries. The primary connections can be either y or delta. The voltages induced in the secondary windings differ in phase by 120 degrees. Hence, each diode conducts for a 120-degree interval of a complete input cycle, and must carry one-third of the average current supplied to the load. The frequency of the dc pulses in the output is three times the fundamental power frequency.

11-1424. THREE-PHASE FULL-WAVE POWER SUPPLY.

11-1425. The double-three-phase rectifier, shown in part B of figure 11-227, employs a



A
HALF-WAVE RECTIFIER



B
FULL-WAVE RECTIFIER

Figure 11-227. Three-Phase Power Supply Circuits transformer having delta-connected primaries and star-connected secondaries. The use of this transformer connection has the advantage over the delta-y connection shown in part A of the figure, of producing less saturation of the transformer cores. The voltages induced into the star-connected secondaries differ in phase by 60 degrees. Therefore, each diode conducts for 60 electrical degrees, and passes a current that is one-sixth of the average load current. The output of this circuit contains a ripple frequency that is equal to six times the fundamental power frequency. Because of this high ripple frequency and low percentage ripple content, filtering is greatly simplified. In multiphase rectifiers that deliver very large values of current, mercury-pool rectifier tubes are generally used.

11-1426. RF HIGH-VOLTAGE POWER SUPPLY.

11-1427. The rf high-voltage power supply is a half-wave rectifier using a conventional LC power oscillator; it is used principally in cathode-ray-tube applications where extremely high dc second-anode potentials at low current are required. The circuit shown in figure 11-228 is typical. Tube V1 is connected in a modified Armstrong oscillator circuit, with tuning accomplished in the plate circuit. The oscillator output is transformer-coupled to the plate of V2 (connected as a half-wave rectifier). Since the step-up rf transformer (air core) has a tuned primary and usually a self-resonant secondary, high efficiency in terms of high voltage with low current is realized. Filament power for V2 is obtained from the oscillator tank circuit by the use of a small pickup loop. This method of obtaining filament power isolates the filament circuit from near-ground potentials, as a precaution against voltage breakdown. Regenerative feedback for the oscillator circuit is provided by capacitive coupling between a metallic spring mounted around the glass envelope of the rectifier tube, V_2, and the plate of the tube. The output voltage level may be varied by C_3; however, C_3 is generally tuned to provide maximum stability rather than maximum voltage. The oscillator is normally tuned to operate at some frequency between 50 and 500 kilocycles.

11-1428. The output of the circuit is applied to an RC filter network. Since the ripple is in the radio-frequency range, small filter constants may be used, with a consequent saving in space, weight, and cost. Also, the shock hazard to persons coming in contact with this high potential is reduced because of the small amount of energy stored by these small filter components.

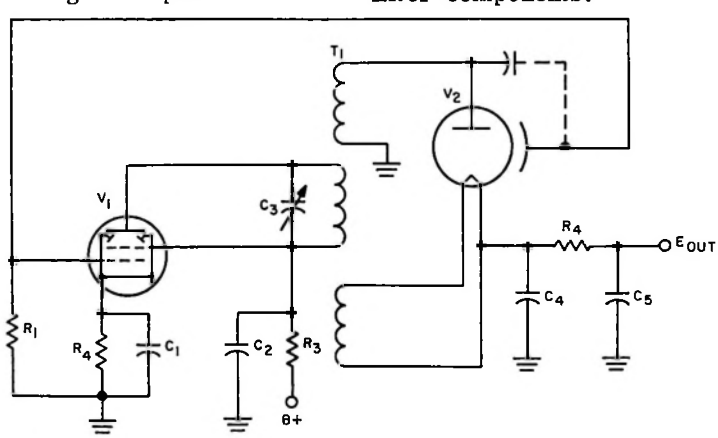


Figure 11-228. RF High-Voltage Power Supply

11-1429. The advantages mentioned above are the principal reasons for the use of this circuit in applications where its poor regulation and stability can be tolerated. Since the high dc voltage output is a result of the heavily stepped-up rf, failure of this circuit is often a result of failure of the oscillator. A fast check of this circuit can be made by measuring the grid voltage of V1, using an electronic voltmeter; oscillator operation is indicated by a negative voltage reading.

11-1430. VOLTAGE REGULATORS.

11-1431. GENERAL.

11-1432. Voltage regulators are used to maintain a constant voltage to the load, even though changes in input voltage or changes in load current occur. They operate on the principle of the voltage divider, the voltage divider consisting essentially of a fixed resistance and a variable resistance.

11-1433. GAS-TUBE REGULATOR.

11-1434. The gas-tube regulator, shown in part A of figure 11-229, is the simplest type of voltage regulator; it consists of a resistor in series with a gas tube, with the regulated output being taken from across the tube. This circuit operates satisfactorily to provide a definite output voltage (dependent upon the type of tube) if the variation in load current is within a certain range. Gas tubes are rated according to the voltage drop that appears across them, and according to the maximum current that can be allowed to flow through them. For example, a VR-105/30 gas tube maintains a 105-volt output, and has a maximum current of 30 ma. The internal resistance of the gas tube depends upon the amount of current flowing through it. If a large current flows, the gas is highly ionized and hence has a low resistance. If a small current flows, the gas is only partly ionized and has a higher resistance. Therefore the voltage drop across the tube is relatively constant over its entire operating range.

11-1435. In part A of figure 11-229, both the load and the gas-tube currents flow through series resistor, R. If the supply voltage decreases, the voltage across the gas tube tends to decrease also. This effect causes the gas in the neon tube to deionize slightly, and causes less current to flow through the tube because of the increased resistance. The resulting decrease in current through resistance R lowers the voltage drop across R, and thereby raises the output voltage to its normal value.

11-1436. If the output of a single voltage-regulator tube is too low in value for certain applications, several gas tubes may be connected in series and the output taken from the combination. The circuit shown in part B of figure 11-229, is designed to provide three different regulated voltages (with low-current drain). If VR-105/30 tubes are used, the circuit will apply 105 volts to load 1, 210 volts to load 2, and 315 volts to load 3.

11-1437. For a gas tube to conduct when a voltage is applied, the voltage must exceed the striking or firing level of the gas, which is the potential required to ionize the gas. For example, a 105-volt regulator must be provided with at least 133 volts to produce positive firing or starting. Such a tube will produce a 4-volt change over the current range of 5 to 40 ma, representing an ac impedance of about 115 ohms. If the gas tube were replaced with a resistor designed for 105 volts at the mid-current rating of the tube (20 ma), the resistor would have an ac impedance of 5000 ohms. Thus, it can be seen that the vr tube acts to prevent changes in voltage. Since it is possible to get negative-resistance effects in a vr tube, it is

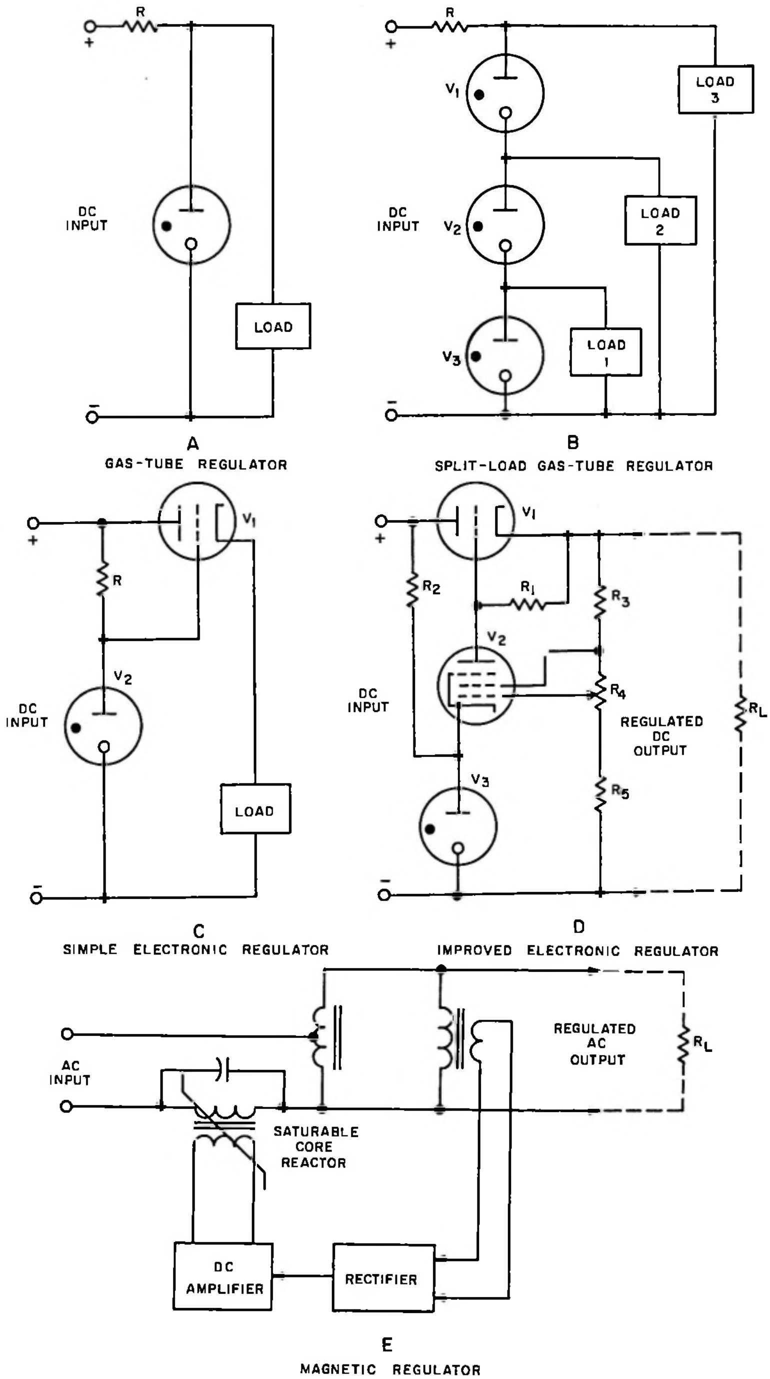


Figure 11-229. **Voltage Regulator Circuits**

recommended that the shunt capacitance for a tube of a given type be limited to the specific values indicated in a tube manual for that type of tube. The maximum value of capacitance that can be used is usually less than 1 microfarad.

11-1438. ELECTRONIC REGULATOR.

11-1439. The electronic regulator, shown in part C of figure 11-229, is superior to a gas-tube regulator in that it maintains the output voltage more nearly constant, and is capable of operating over a much greater load-current range. Regulation is accomplished by using the plate-to-cathode resistance of an electron tube in series with the output of the power supply, which functions as a variable resistance, and thus provides the voltage drop necessary to compensate for any change in the output voltage. Since the plate-to-cathode resistance depends upon the plate current, the amount of resistance can be controlled automatically by connecting the circuit so that changes in the output voltage affect the grid bias. The value about which the resistance varies is determined by the initial bias level established for the tube.

11-1440. To understand the operation of the electronic regulator circuit, assume that the output voltage supplied to the load is correct. See part C of figure 11-229. The cathode of the series tube, V_1, is positive with respect to ground by the voltage across the load, while the grid is held somewhat less positive by the voltage drop across the gas tube, V_2. The difference between these two voltages, of course, is the bias voltage, which is set at the proper value to cause the tube to have the required amount of plate-to-cathode resistance to produce the correct output voltage. If the output voltage increases by a small amount, the cathode potential of V_1 increases too, but the grid potential changes very little because of the regulating action of the gas tube, V_2. Hence, the bias on V_1 increases, and the tube resistance becomes greater. Therefore, a greater voltage drop occurs across V_1, to compensate for the increase in the output voltages.

11-1441. A regulator with increased stability can be designed by using a pentode with a high amplification factor to control the resistance of the series regulator tube. The improved electronic regulator illustrated in part D of figure 11-229 produces an output voltage that is relatively independent of input voltage and load changes over a wide range. The output voltage is developed across resistors R_3, R_4, and R_5 in parallel with the load impedance. The load current flows through the regulator tube, V_1, which is in series with the output. The remaining elements in the circuit (R_2, V_2, and V_3) are used to control the resistance of V_1. The plate voltage of the control tube, V_2, is taken from the output of the regulator. The cathode potential of V_2 is taken from the output of gas-tube regulator V_3, and is therefore a constant voltage. The control-grid voltage is determined by the setting of potentiometer R_4, and is such as to provide the correct bias for V_2, which, in turn, determines the amount of plate current. The plate current of V_2 flows through plate load resistor R_1, across which the produced voltage drop provides the bias for V_1. Therefore, the adjustment of R_4 determines the resistance of V_1. This adjustment is initially set for the desired regulated output.

11-1442. If the output voltage tends to increase (because of either a decrease in the load current or an increase in the rectifier voltage), the potential at the control grid of V_2 increases. Since the cathode potential remains constant, a decrease in bias occurs, which allows a greater plate current in the tube. The increased plate current produces a greater drop across R_1, which increases the bias on V_1. This action, in turn, increases the voltage drop across V_1. Therefore, a larger portion of the rectifier voltage appears across V_1, and the regulator output is returned to the correct value. If the output voltage decreases, the action of the circuit is reversed. Since a pentode is used to control the resistance of V_1, small variations in the output voltage are amplified sufficiently to operate the circuit. Because of its high sensitivity to voltage changes, an electronic regulator of this type will reduce any ripple that remains in the filtered output of the power supply. The anode of the gas tube is returned to B+ through R_2 to ionize the gas when the power supply is first turned on.

11-1443. MAGNETIC VOLTAGE REGULATOR.

11-1444. Part E of figure 11-229 illustrates the circuit of a magnetic voltage regulator. This type of voltage regulator employs the changes that occur in the impedance of a saturable-core reactor, as the reactor core is subjected to different degrees of saturation, to regulate the ac voltage applied to a power supply or other electronic equipment. The degree of saturation of the reactor core is controlled by a dc amplifier, the bias of which is provided by a rectifier that receives its input voltage from the regulated ac output.

11-1445. The circuit functions as follows: An increase in the ac output results in a greater dc output from the rectifier, and hence in a greater bias on the dc amplifier. This action reduces the plate current flowing through the dc amplifier, and results in a decrease in the degree of saturation of the reactor core. The reactor presents a greater impedance to the ac input voltage, thus lowering the voltage applied to the autotransformer. Therefore, the ac output voltage is returned to the correct value. A capacitor is often connected across the reactor to compensate for the distortion introduced by the saturating element.

11-1446. When this device is used for regulating a dc power supply, the rectifier shown in part E of figure 11-229 is omitted, and the power supply is used in its place to furnish bias voltage for the dc amplifier. Thus, the dc output of the power supply controls the ac input voltage to the power supply, to provide regulation.

11-1447. VIBRATOR POWER SUPPLIES.

11-1448. GENERAL.

11-1449. Vibrator power supplies are used to produce a high dc voltage from a low-voltage dc supply, usually from a 6- to 24-volt battery source. These power supplies are generally used to provide a low-current plate-voltage supply for portable or mobile electronic equipment. The high ac voltage is either rectified by the vibrator itself (synchronous) or by a separate rectifier circuit (non-synchronous), to produce the dc output. Vibrator circuits require good rf filtering and shielding. As compared with a dynamotor, a vibrator is less expensive and lighter, but has a shorter useful life and lower efficiency.

11-1450. NONSYNCHRONOUS VIBRATOR.

11-1451. Part A of figure 11-230 shows a typical non-synchronous vibrator supply. When switch S is closed, current flows through the lower half of the transformer primary and the magnet winding, "m". The armature "a", is attracted to the contact "b", thereby short-circuiting the magnet winding. This permits the armature to be released, and it springs back to make contact with point "c". Since the short circuit is now removed from the electromagnet, the core is again magnetized, and the armature is pulled down until it again makes contact with point "b". Thus the armature keeps vibrating at its own natural frequency (usually around 100 cps), and alternately touches points "c" and "b". During the time that contacts "a" and "b" are closed, the lower portion of the primary is connected across the battery, and current flows through the lower half of the primary. This induces a voltage of a certain polarity in the secondary. Likewise, when contacts "a" and "c" are closed, current flows through the upper half of the primary, because this portion is now connected across the battery. Since this current is in the direction opposite that through the lower half of the primary, a voltage of the opposite polarity is induced in the secondary. Therefore, an alternating voltage (essentially square in waveform) is developed across the secondary. The output of the secondary is then rectified and filtered in the conventional manner. Capacitor C_1 (commonly called a buffer capacitor) is connected across the secondary to absorb the high-voltage inductive effect produced by the sudden cessation of current through the primary when the contacts are broken. The rf choke (rfc) and C_1 comprise a "hash filter" (as do C_2 and its rfc), and are included for the purpose of preventing noise (or hash), produced by the vibrating contact, from being carried out through the dc circuits.

11-1452. The vibrator is a plug-in device provided with two forms of shielding. An

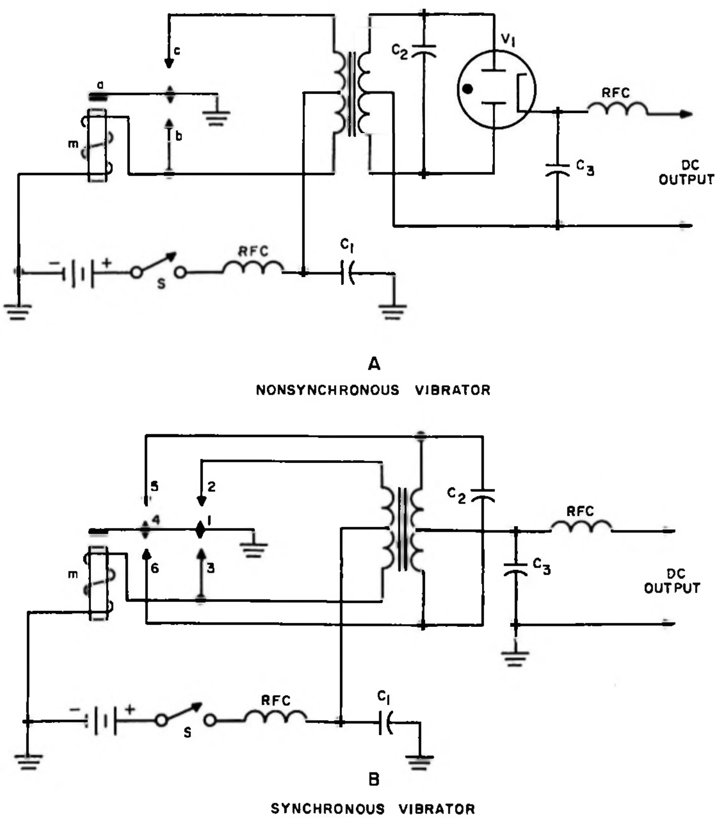


Figure 11-230. Vibrator Power Supply Circuits

inner shield, usually made of foam rubber, acts as an acoustical absorber to prevent radiation of audible vibrations, and an outer metallic shield prevents radiation of rf energy. As an added precaution, the entire power supply is usually completely surrounded by a second metallic shield to further reduce rf radiation.

11-1453. SYNCHRONOUS VIBRATOR.

11-1454. The synchronous vibrator circuit shown in part B of figure 11-230, produces rectification without the use of a separate rectifier circuit. When the battery voltage is applied by closing switch S, coil "m" of the vibrator becomes energized, and causes contacts 1 and 3 to close. During this interval, current flows through the lower half of the primary winding of the transformer and induces a voltage of a certain polarity in the secondary. When contacts 1 and 3 close, the coil becomes shorted and is de-energized, and spring action of the armature causes contacts 1 and 2 to close. For this condition current flows through the upper half of the primary and induces a voltage of the opposite polarity. As soon as contacts 1 and 3 are broken, coil "m" again becomes energized and the process is repeated.

11-1455. During the time that an ac voltage is being produced in the secondary, a second set of contacts (4, 5, and 6) on the vibrator rectify the output of the secondary. These contacts accomplish rectification by connecting alternate sides of the secondary winding to ground in synchronism with the reversal in polarity of the induced voltage. This action causes one-half of the secondary to supply each alternate half-cycle, and thus results in a unidirectional current. A conventional filter connected to the center tap of the secondary smoothes out the pulsations to produce a steady output. Capacitor C_2 functions as a buffer, and the rf chokes, along with C_1 and C_2, serve as hash filters.

11-1456. The output polarity of a synchronous vibrator power supply depends upon the polarity of the dc input voltage (the polarity of the nonsynchronous vibrator supply is not affected by dc input-voltage polarity. For this reason, means are often provided for reversing the dc output polarity. In some cases the vibrator base is constructed so that it has two plug-in positions; in other cases a reversing switch or flexible jumper leads on a terminal board are used to reverse either the primary or secondary winding of the transformer.

11-1457. MAINTENANCE NOTES.

11-1458. Vibrator power supplies normally have a relatively short useful life. Contact failure is the most common type of trouble, and is usually indicated by lowered or erratic rectified voltage. Often, vibrator failure is caused by the failure of the buffer capacitor, C_1 in part A, and C_2 in part B of figure 11-230. Vibrators are ordinarily manufactured in sealed units, and defective vibrators are usually replaced as a unit. It is considered good practice, when a vibrator is replaced, to also replace the buffer capacitor. The capacitance and voltage rating of the buffer capacitor are critical, and the replacement unit should be identical with the original.

11-1459. DYNAMOTORS.

11-1460. A dynamotor is a combined dc motor and generator operating in the same magnetic field. It may have two armatures on one shaft or two sets of windings and two commutators on one armature. Dynamotors are employed to generate high dc voltage when driven from a low-voltage supply, and are used in mobile equipment when high-current stability and high-voltage stability are required. Current is supplied to the low-voltage commutator, and flows in the low-voltage winding on the armature. At the same time, this current flows in the field coils, producing flux that reacts with the flux in the low-voltage armature winding. This results in torque which causes the armature to rotate. The secondary winding in the armature has a voltage induced in it, because it is rotating in the common field flux. Since the secondary winding has a larger number of turns than the low-voltage winding, the induced voltage is higher in proportion to the turns ratio. The current in this high-voltage winding flows through the high-voltage commutator, the brushes, and the load. When both windings are on one armature, the armature reactions neutralize each other, so that neither interpoles nor brush shifting is required to prevent commutation difficulties. Since fewer mechanical parts are required, and only one set of bearings is needed, higher efficiency results. Both windings in the armature cut the magnetic field at the same time and rate; therefore; the induced emf is directly proportional to the number of turns. The emf ratio cannot be changed by increasing the field strength, because this causes the motor armature, as well as the generator armature, to cut more magnetic lines. This slows down the armature, but the emf remains unchanged. If the field is weakened, the speed of the armature increases, but the induced emf does not change. Not all dynamotors are rated continuously at full load; therefore, it is important to check the specifications before operating the dynamotor over a longer period than one-half hour. If a dynamotor is rated as "intermittent", it may sometimes be run continuously, if necessary, if the covers are removed to allow more air to circulate through the windings. Observe the temperature carefully; if it starts to climb above the rated temperature, the dynamotor must be shut down and allowed to cool. If this is not done, the windings may burn out. Be sure that guards are placed around a dynamotor operating without its proper protective covering so that personnel cannot come in contact with exposed high-voltage components. If the dynamotor brushes are removed and replaced, make certain that each brush is replaced in the proper holder. The brushes and holders are usually marked with a plus or minus sign. The high-voltage brushes of a dynamotor should be fitted with exceptional care, as these contacts can cause a loss in output voltage. Do not attempt to remove the discoloration from commutators, as this is a copper oxide that is an aid to commutation. Since a dynamotor is a form of motor and generator, the procedure for testing is the same as for conventional dc machinery; that is, the two windings are treated separately. (Refer to paragraph heading 11-1513 for further test information.)

11-1461. TIME DELAY CIRCUITS.

11-1462. Hot-cathode gas tubes such as transmitters, public address amplifiers, etc., are used where high power is required. Mercury-vapor tubes contain mercury vapor in equilibrium with liquid mercury. The mercury vapor is formed in these tubes when the small amount of liquid mercury within the tube is vaporized by the hot cathode. When a plate voltage in excess of the ionization voltage is applied to the rectifier, and circuit conditions permit, the plate current will increase to the full value of the cathode emission. This is a result of neutralization of the space charge within the tube, from positive ions produced by the ionization process. The tubes may have either filaments or heated cathodes, which are invariably oxide-coated and subject to disintegration by ion bombardment. Tubes of this type must be brought to normal operating temperature prior to the application of plate voltage, or damage to the tube will result. High-vacuum recitifiers are much more rugged than gas tube rectifiers, since the cathodes are ordinarily made with a coating or thoriated tungsten. Under normal operating conditions, no damage to the tube occurs during the warm-up time. However, in certain applications, especially when high plate voltages are involved, preheating the tube will extend its life.

11-1463. In order to protect and lengthen the useful life of power supply rectifiers, a time-delay circuit to delay the application of plate voltage is incorporated in many equipments. The power transformers used with these circuits ordinarily employ separate filament and plate transformers. The most common time-delay circuits use a time-delay mechanism, usually a motor- or escapement-driven relay. The filament transformer and the mechanism are energized simultaneously by the application of input power. The delay mechanism incorporates electrical contacts, which close the input power to the plate transformer after a predetermined time (about 20 seconds).

11-1464. Another type of time-delay circuit employs a special power supply with rectifier and plate-circuit relay for the purpose of controlling the input power to the equipment power supply. The filament windings of the rectifiers are connected in common with the special power supply. Input power is first applied to this power supply; as the rectifiers reach normal operating temperature, the plate current from the special rectifier actuates the time-delay relay in the plate circuit, thereby applying power to the equipment power supply. Variations of these delay circuits may employ components like thermally actuated relays, holding relays, etc.; however, the basic purpose is the same. Often the heater voltage and plate voltages are controlled by separate, manually operated switches. In many situations, such as in locations of high humidity, the heater voltage may be applied even though the equipment may not be in operation. When starting equipment which is so equipped, it is good practice to apply filament power 30 seconds or more before applying plate voltage, since longer heating times extend the tube life.

11-1465. Another type of effect that can shorten tube life results from connecting the filaments of different types of tubes in series. If the cold resistance values of the tube filaments vary substantially from each other when power is first applied, those tubes having the least resistance are subject to surge currents far in excess of their ratings. The surge current, for series-filament-connected equipment, is often limited by a negative coefficient resistor connected in the filament circuit. This resistor has maximum resistance when cold, and minimum resistance when hot. When power is applied to this equipment, the surge current is limited by the high resistance of the cold resistor. By the time the filaments reach operating temperature, the resistor has reached a low quiescent value which allows the rated current to flow through the circuit.

SECTION IX

ELECTRICAL MACHINERY

11-1466. GENERAL.

11-1467. Rotating electrical machinery (motors, generators, dynamotors, amplidynes, etc.) are used for many purposes in almost every large electronic equipment. Proper service and repair of these components is essential for reliable operation of the equipment. All motors and generators consist of a rotatable member (rotor), usually an armature, enclosed in a magnetic field (stator). Various types of construction and electrical connections produce the different characteristics necessary for specific applications of the machine.

11-1468. BASIC GENERATORS.

11-1469. A generator is a machine that converts mechanical energy into electrical energy. A basic generator consists of a single-turn coil (armature) suspended in a magnetic field. When the coil is rotated at a constant speed, a sinusoidal voltage is induced in the coil. An alternating-current generator has a slip ring connected to each end of the coil. The slip rings are mounted on the shaft rotating the coil and are insulated from each other. A stationary brush, made of either carbon or metal, rides on each slip ring and transfers the resulting voltage to the generator load. One rotation of the single-turn coil produces one cycle of alternating current (360 electrical degrees). The basic direct-current generator is similar to the basic ac generator, except in the method of taking the output from the rotating coil. In the dc generator, the rotating coil is connected to the external circuit by means of a device called a commutator. This is effectively a switch which converts the alternating current from the rotating coil to current which flows in one direction through the external circuit. In general, a commutator consists of a metal ring, usually of copper, which is divided into a number of segments. These segments are insulated from each other and from the shaft upon which they and the rotating coil are mounted. For the basic generator described, the commutator consists of only two segments, with one segment connected to each end of the coil. Two brushes, mounted on opposite sides of the commutator, bear on its surface, so that electrical contact is made between the coil and the external circuit. One rotation of the single-turn coil (360 electrical degrees) produces two pulses of direct current. The output waveform looks like the output of a full-wave rectifier.

11-1470. BASIC MOTORS.

11-1471. A motor is a machine that converts electrical energy into mechanical energy. A magnetic field exists about any current-carrying conductor; the strength of this field is dependent upon the amount of current. When a current-carrying conductor is placed in a magnetic field, a force is exerted on the conductor, tending to move it out of the field. This is the fundamental principle of motor action. When current is passed through the coil of the rotor of the simple generator described above (and the rotor is in the proper position of the magnetic field) a turning of the rotor results. If the current of the coil is reversed at the proper time, the turning can be sustained to cause continuous rotation. Current reversal is accomplished by the nature of alternating current to operate ac motors; this is accomplished by means of a commutator in operation of dc motors. Motors can be considered the reverse of generators. Manufactured motors and generators are complex versions of the basic motor and generator described previously. It is not intended in this text to enter into the theoretical principles of operation of these machines. The subject is covered in physics text books and other easily obtainable books.

11-1472. DC GENERATORS.

11-1473. ARMATURE CONSTRUCTION.

11-1474. GENERAL. To increase the output of the basic generator to a usable value, the number of coils in the rotating member and/or the number of turns in the coils must be increased. Up to this point, a rotating coil of only one turn has been considered. In a few types of heavy-current generators, single-turn coils of heavy wire are used, but in most types, the coils are made up of many turns of lighter-gauge wire. High efficiency of the generator requires a strong magnetic field. Placing a core of magnetic material in a coil increases the density of the magnetic flux. Therefore, to increase the flux in the rotating member of the generator, the coil is wound on a core of magnetic material. The rotating member of a generator is called the armature. If the single coil of wire and two-segment commutator of the basic generator are replaced by two coils and four segments, the number of flux linkages per revolution is doubled, with a consequent increase in ripple and increase in the average output voltage. The output can also be increased by adding to the number of turns in the coils. By increasing the number of coils and/or the number of turns in the coils, the output of the generator can be increased to any reasonable value, as long as the core material does not approach the saturation point.

11-1475. An armature with 12 slots for coil wires, and six segments on the commutator is shown in part A of figure 11-231. The diagram of this winding shown in part B is known as a "developed" view. With the commutator end of the armature toward the reader, the coil positions in part A can be matched to their respective positions in part B. Starting at commutator segment A, and proceeding clockwise around the armature, the first coil passes to the rear through slot 1, crosses the back, and returns through slot 6 to segment B. The next coil starts at segment B, passes through slot 3, and returns through slot 8 to segment C. The remaining coils are wound in this manner until the final coil ends at segment A. In the basic conception of a coil rotating in a magnetic field, drawings usually show the return side 180 degrees away from the entering side, although this condition is not necessarily observed in practice. For example, in part A of figure 11-231, it is readily seen that slot 1 is not diametrically opposite slot 6. Actually, the starting slots and the return slots are in chordal positions with respect to one another (1, 6, 3, 8; etc.), so that the coils may be placed in a symmetrical manner around the armature. If an attempt were made to return the coils through diametrically opposite slots, as shown in part C, it is easy to see that an unbalanced condition would result, leaving segments and slots unused. On the other hand, there are many cases where the return slots fall opposite the starting slots naturally, permitting a symmetrical winding. This is the condition for the armature shown in part D, which has 14 slots and seven commutator segments. In multipolar generators, the coils should return through the slot located under the adjacent unlike poles (four poles, return 90 degrees away; six poles, 60 degrees; etc.). For each complete coil on the armature, there must be at least one segment on the commutator. For instance, in part B of figure 11-231, there are six coils and six segments, and in part D, there are seven coils and seven segments. In some cases, particularly with larger machines, there are more segments than coils.

11-1476. DRUM TYPE ARMATURES. The drum-type armature is the only type in common use today. The core is made of steel lamination (to reduce eddy current losses) of the same type used in the construction of field poles. In small generators, the laminations are keyed discs which are pressed onto the shaft, and held in place by rings or nuts. In larger machines, only the outside of the ocre is made up of laminations that are bolted or welded into a spider. This spider is either welded or cast and forms the center of the core. Regardless of the size of the armature, the coils are wound in parellel slots cut in the outer surface of the core, as shown in figure 11-232. Armatures for large multipolar machines usually have form-wound coils. These coils are finished separately, and are then placed in the armature slots and secured by keepers or wedges. The wedges are made of fiber, wood, micarta, or metal. They are held in place by binding wire made from bronze or steel. The wire is tinned, so that solder may be puddled on the binding, to help in balancing the armature. Some armatures have distance pieces in the lamination. These are corrugated, serrated, or slotted pieces of steel of the smae shape as the laminations. The irregular surface holds the laminations apart so that air can circulate, for cooling purposes.

11-1477. LAP WINDING. A lap winding is so called because the coils overlap one other. A typical armature with two lap-wound coils in place is shown in figure 11-232. In figure 11-231, the windings in parts A, B, and D are lap windings; the winding in part C does not close on itself, and therefore, a complete

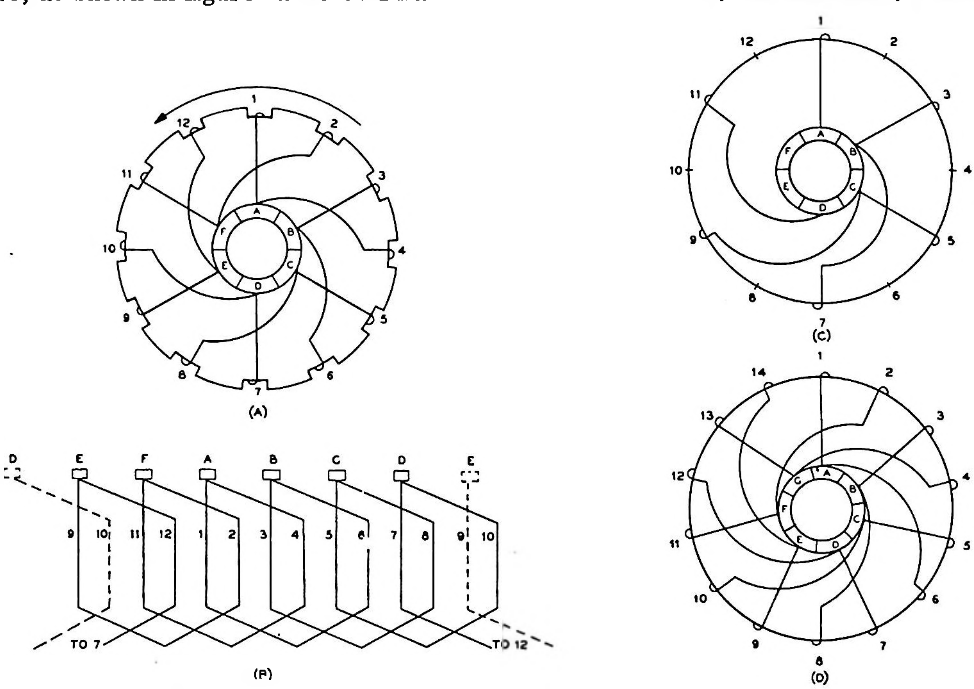


Figure 11-231. Armature Winding Development

circuit through this armature does not exist. This type of winding is adaptable to generators designed for high current output at relatively low voltage.

11-1478. A brush is required for each pole of a lap-wound armature. Since there are always as many paths as there are poles, the current must be taken off, or commutated, at the ends of each path. The voltage in each path is the sum of the voltages induced in the coils of that path; therefore, the total generated voltage is limited by the number of coils in a path. The number of coils is, in turn, limited by the number of slots it is possible to use, since core material must be cut away for each slot.

11-1479. In any set of parallel-connected coils, it is desirable to have the same amount of emf developed in each coil, to prevent currents from circulating between the coils of the set. In practice, the unavoidable inequalities of coil emf result in loss of energy and increased heating of the coils. In small generators, these currents are negligible, and do not cause detrimental heating of the brushes. In large machines the currents can be excessive. To overcome this condition, several points in the armature which are theoretically at the same potential are connected together by heavy conductors called equalizers. Theoretically, all coils of like potential should be inter-connected, but it has been found sufficient, in actual practice, to equalize every third or fourth coil.

11-1480. WAVE WINDING. A wave winding derives its name from the shape of the coils before they are placed on the armature core. In this type of winding, illustrated in figure 11-233, the series-connected armature conductors are cutting flux under adjacent unlike poles at the same time. The winding progresses around the armature a sufficient number of times to connect all the conductors in series. This winding always has two parallel paths through the armature, regardless of the number of poles. Therefore, it is used only in machines requiring a limited current for a given output voltage.

11-1481. A wave winding has an advantage over a lap winding in that it produces a higher voltage for a given number of poles and armature conductors. This is true because the coils in each path are series-connected, and the number of paths is always two, regardless of the number of poles; therefore, the voltage produced at the brushes is the sum of the voltages produced around the entire armature. In a lap winding, the various paths are parallel-connected; there are as many

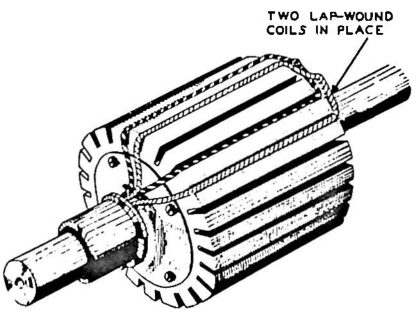


Figure 11-232. Drum Armature, with Two Lap-Wound Coils in Place

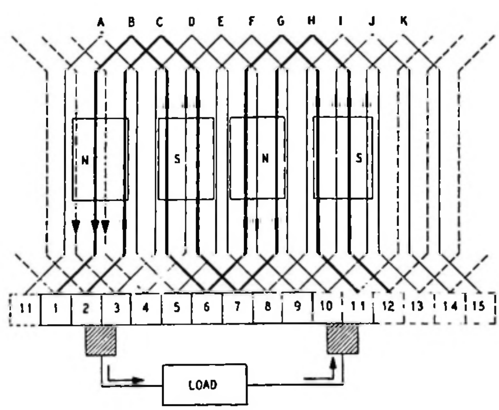


Figure 11-233. Developed View of Wave Winding

paths as there are poles. For a given number of poles and armature conductors, a lower voltage is produced, although the load can be greater than with the wave winding.

11-1482. MULTIPLEX WINDINGS. The windings described in the foregoing text are known as simplex lap or simplex wave windings. Simplex refers to the arrangement in which only one complete winding of either type if placed on an armature. Multiplex windings are those having two or more complete windings, and are classified as duplex, triplex, etc. A multiplex winding is capable of greater output current than a simplex winding of the same general type.

11-1483. COMMUTATORS. Commutators are made up of segments of hard-drawn copper. In small generators these segments are molded in a metal sleeve filled with insulating compound, and in larger machines, they are clamped in frames. The segments must be insulated from the frame and from each other. One of the main requisites of commutator insulation is that it shall not be affected by moisture or changes of temperature. It must have enough elasticity to fill in the spaces when the commutator bars expand or contract. Although many substitutes have been tried, mica has been found to be the most satisfactory material for this purpose. Most of the substitutes were either eaten away by sparking or were unsatisfactory because of moisture absorption. The connection of the coil to the commutator is made through radial strips called risers. In some cases the risers are an integral part of the segment while in others they are used as cooling vanes and are separate pieces of copper. The coil leads may be bolted, riveted, or soldered into the risers. Usually when the commutator diameter is close to that of the core, no risers are used.

11-1484. The number of segments required for a commutator is determined by several factors. The most important are the voltage between segments, the amount of fluctuation in the output, and the physical size limitations in the construction. Since an emf in excess of 25 volts across adjacent commutator bars can lead to a breakdown of the insulation or a flashover between the bars, the average design potential across commutator bars is approximately 15 volts. Machines that have a high output voltage require a large number of segments in the commutator. The number is determined by dividing the output voltage by 15 (the maximum voltage desired across each pair of segments). Since there must be a coil element, or leg, for every segment, the number of coils is dependent upon the number of segments. Therefore, the number of segments required is determined basically by the output voltage, and the number of coils required is determined by the number of segments. The number of segments in a commutator must be equal to, or a multiple of, the number of slots in the armature.

11-1485. COMMUTATION. The emf induced in any one coil of a direct-current generator is alternating. For the current in the load circuit to flow in one direction, rectification is necessary. The rotating commutator provides this rectifying action. The ideal condition is for the current flowing in a coil, as it passes under a brush located in the load neutral plane, to decrease from its maximum strength so that it reaches its minimum value at the instant the coil is passing through the commutating plane. Immediately after passing through the plane, it should start building up in the opposite direction as it recedes from the commutating plane. In the coil, the current reversal from maximum positive to maximum negative value is essentially uniform. At the same instant that the current induced in the particular coil undergoing commutation is at its minimum value, current flows through the coil from the adjacent coils (which are cutting flux lines), and then on through the brush to the external load circuit.

11-1486. Commutation under the conditions just described rarely exists, since the neutral load plane shifts with the load, while the position of the brushes is fixed. As a result, some sparking at the brushes will almost always be present. The brushes are seldom, or never, in actual contact with a commutator surface. This fact is demonstrated by the simple process of measuring the voltage drop across the contact point. The voltage drop is a constant value, regardless of the amount of current being drawn through the contact. This indicates that conduction must be occurring through an ionized gas path since constant voltage drop is a characteristic of such conducting mediums. The condition indicates that the current generated in the armature passes from the commutator to the brushes in numerous minute arcs. This arcing produces an ionized condition between the surfaces. This arcing also causes the condition known as high mica. High mica is the term in common use, though in reality, the condition is low copper. It was previously thought that the high-mica condition occurs because the copper is softer than the mica, and therefore, wears down much faster. This theory is disproved by the fact that serious wear of the copper continues long after the copper height is reduced to the extent that the brushes ride almost entirely on the mica. Obviously, then, the condition cannot be attributed to friction. The present theory is that the copper is carried away by ionization caused by the arcs. Arcing, which is always present at the brushes, causes pits or depressions in the surface of the commutator, and these pits prevent proper contact. As the contact becomes poorer, the arcing increases in magnitude, causing more copper wear and higher mica; this cycle continues until the commutator becomes inoperative. The occurrence of high mica can be prevented by undercutting the mica whenever the copper becomes reduced sufficiently to warrant this step. It is also essential that the brushes in any machine be carefully fitted to the commutator, to assist in retarding the high-mica condition as much as possible.

11-1487. BRUSH TYPES. Any brush, regardless of its quality, may give satisfactory performance on one machine and completely fail on another. There is not sufficient knowledge of the properties of the material composition of brushes, insofar as they affect current collection, to permit exact prediction of results. The ultimate test of suitability must be made on the machine itself. It is possible to overcome minor machine defects by trying brushes of various characteristics until satisfactory results are obtained. Brushes are available in a wide variety of types and grades of material to meet all normal requirements. Brushes are made from one of four general classes of material. Each type is superior when used on equipment for which it is designed. Carbon brushes are restricted to use on machines of low current density and speeds. Carbon gives excellent results where cleaning action is required at low peripheral speeds, where mica is flush and it is impractical to undercut, and where strong oxidizing fumes are present. Electrographitic brushes are made from a base material such as carbon, but are processed at high temperatures in an electric graphitizing furnace. The treatment produces radical changes in the characteristics of the material, increasing its electrical and thermal conductivity, rendering it much softer, and at the same time very tough. These brushes have low friction, are non-abrasive, cool running, and capable of carrying more current than the carbon grades. Graphite brushes are readily distinguished from the black carbon grades by their silvery appearance and soft, flaky texture. The high-quality graphite used for brushes is a mined product, in contrast to carbon, which is manufactured from base materials. To manufacture graphite brushes, it is not necessary to subject the material to high temperatures; therefore, it is a simple matter to add any amount of cleaning action desired. The abrasive properties of this class of brush may be altered without making a change in the grade. Therefore, it is possible to manufacture a grade of graphite capable of producing any degree of commutator film desired. Copper-graphite brushes are made from a mixture of powdered copper and graphite, pressed together and baked at a low temperature. Copper-graphite brushes are primarily designed for use where high currents are to be carried.

11-1488. Brush shape is determined by the type of holder used and the angle of the brush with respect to the commutator. Brush holders are made up of sheet-metal stampings for smaller machines, and are cast in aluminum or bronze for larger generators. The holders must be constructed substantially, so as not to permit vibration or to allow distortion from heat. Holders in some measure help to conduct heat away from the brushes.

11-1489. FRAMES.

11-1490. Generator frames were at one time made of cast iron, but because of the great weight and inferior permeability of this material, it has given way to steel. The base or mounting brackets are usually an integral part of the frame as are the supports for the armature. Figure 11-234 shows that a section of each field core is flared at the ends. These sections, called the pole shoes, serve to reduce the reluctance of the air gap. They also serve as mechanical supports for the field coils, to prevent them from being pulled off the poles by the great mechanical stresses on the coils resulting from magnetic force.

11-1491. Cast steel is used for the frames of large machines, while rolled sheet steel is used to fabricate smaller generators. In the large machines, the pole pieces are usually separate, and are made of laminations that are bolted or riveted together. A spring flange forms a metallic sleeve between the field coil and the pole piece. This flange holds the coil tightly against the frame, thus preventing any chafing of the insulation. An insulating pad of waterproof oiled asbestos between the coils and the frame provides additional protection for the coil. The field coil is insulated with mica and glass wrappings, and is vacuum-treated with waterproof insulating compound. The metallic sleeve exerts a spring action in the same direction as that of the magnetic pull on the coil, thus preventing shifting and vibration of the coil.

11-1492. The number of poles in a field structure is determined by the required output of the generator and the speed of the armature. The voltage generated rises to its peak value twice in one revolution of a two-pole generator, and four times in one revolution of a four-pole generator. The

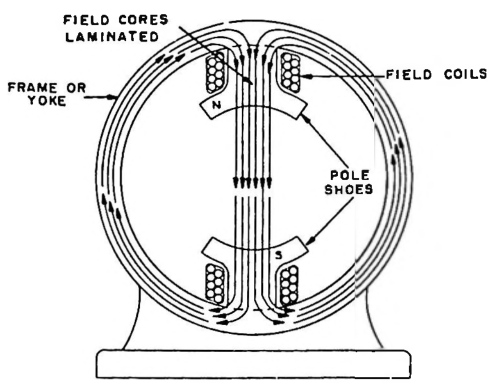


Figure 11-234. Typical Field Frame

number of poles has a direct bearing on the output of any generator, because, for a given armature speed, it determines the rate at which the coil legs cut the flux. The flux for the magnetic circuit in the field poles is provided by coils which are wound on the cores and connected in series; the resultant circuit is called the field circuit. The field circuit can be connected in parallel with the armature or in series with the armature. Field coils which are connected in parallel with the armature are wound with small wire to provide high resistance, and are called a shunt field. The series-connected coils consist of a few turns of heavy wire to provide low resistance, and are called a series field. The field coils are usually wound on micarta forms for large generators, but are simply wrapped with insulating tape in the smaller machines.

11-1493. Permanent-magnet fields have a limited strength, and find use commercially only in machines that do not require a large current output, such as ignition magnetos for internal combustion engines, meggers for testing insulation, and electrical measuring instruments.

11-1494. TYPICAL GENERATORS.

11-1495. GENERAL. Figure 11-235 illustrates a generator which is typical of stationary machines in sizes from one half to 150 kw. All the major components are visible. This type of generator is usually driven at a constant rate of speed, and is built to operate in ranges from 500 to 3600 rpm. The armature core is built up of laminations mounted on a spider. This assembly is, in turn, mounted on the shaft. The cooling spaces between the laminations are visible in a vertical position, in the center of the generator. Commutator bars are mica insulated, and are locked in place by a ring nut that exerts an even pressure on them, thus preventing high bars. A one-piece frame of rolled steel is used with feet welded to it and provides the foundation for mounting the generator components. The field coils are wound on micarta forms, and are held in place by the pole pieces which are bolted to the frame. In this type of machine, the bearings may be of the ball or sleeve type.

11-1496. Generators with oil lubrication devices should not be mounted at an angle greater than 15 degrees. The brushes, brush rigging, and commutators are readily accessible on generators of the open-frame type. The bearing shown is of the ring-oiled type, and consists of a heavy-walled bronze bearing and a bronze oil ring. An oil cup is provided for filling. Its location determines the oil level and prevents over-lubrication. Since the oil rings rides on the shaft, it turns continually, and carries oil from the reservoir to the top of the shaft, where it is distributed to the bearing surfaces. These generators are often equipped with yarn-packed bearings. The yarn-packed bearing has a long bundle of long-fiber wool yarn looped on the shaft with both ends extending to the bottom of the oil well. The yarn draws the oil upward, by capillary action, to the revolving shaft. Packing or seals are provided at each end of the bearings to prevent the oil or grease from escaping along the shaft or to the internal structure of the generator.

11-1497. SERIES GENERATOR. When all the field windings are connected directly in series with the armature, as in part A of figure 11-236, the generator is series-connected. The no-load voltage of this generator can only result from the residual magnetism, since there is no current in the field, and therefore, no excitation. Theoretically, any voltage increase depends upon the load current through the series field, and the voltage is directly proportional to the load. Actually, the generated voltage is proportional to the excitation of the ampere turns

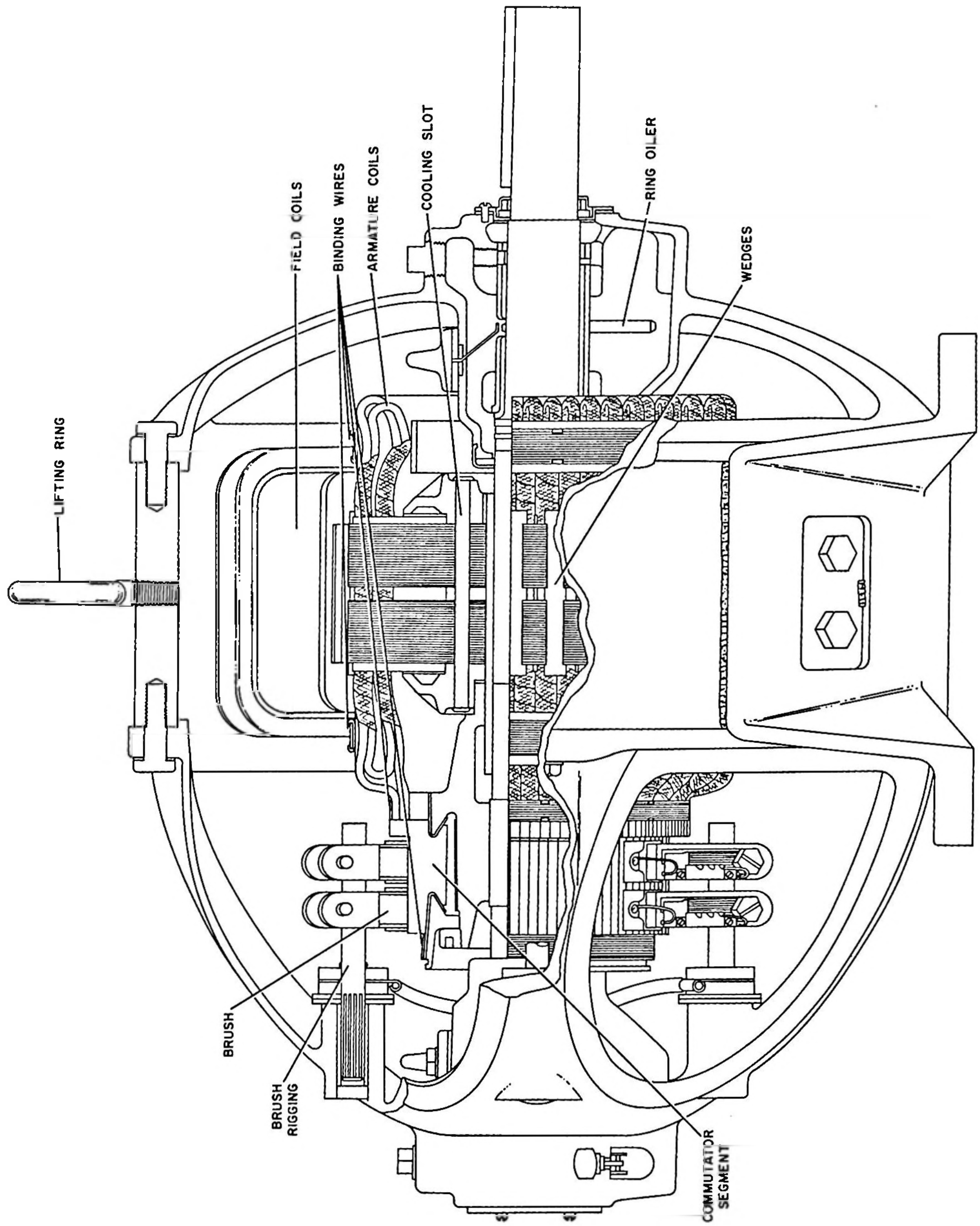


Figure 11-235. Cutaway View of Stationary Generator

of the field, less the counter emf, less the losses from resistance in the coils of the armature and the field. The resultant voltage is called terminal voltage. Whereas the shunt generator maintains nearly constant potential under varying load conditions, the voltage output of the series generator varies considerably under such conditions. Consequently, the series generator is limited to constant-current applications. Closer regulation is made possible by shunting the field with a resistance, which may be either manually or automatically controlled. The wire used in the construction of series field coils is of a relatively large size, as it must carry the full-load current of the generator. This feature serves to identify series fields.

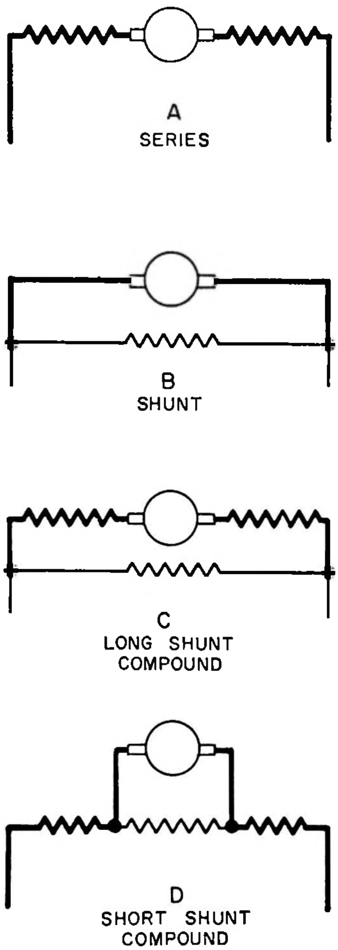


Figure 11-236. Field and Armature Connections for Series, Shunt, and Compound Generators and Motors

11-1498. SHUNT GENERATOR. When the field windings are connected in parallel with the armature, as in part B of figure 11-236, the generator is shunt-connected. Shunt-connected generators are readily adapted to applications where the speed of the prime mover cannot be kept constant, as in the aircraft and automotive fields. In these uses it is not desirable to have a constantly fluctuating output voltage. It is necessary, therefore, to vary the shunt field resistance (which controls the field current), to compensate for changes in the speed of the prime mover. The control of generator output by varying the field resistance is the basis of practically all voltage regulation.

11-1499. COMPOUND GENERATOR. A compound generator is a combination of the series and shunt types. When the shunt winding parallels the series winding and the armature, it is called a long shunt compound. When the shunt winding parallels the armature only, it is called a short shunt compound. These types are illustrated in parts C and D of figure 11-236. The characteristics of long and short shunt compounding are very similar, although the short shunt finds wider use. The purpose of the series winding is to control the output of the generator in relation to the load; it is connected so that it helps to increase the emf of the shunt field. When it is helping the shunt field, it is said to be "cumulative compound." When the series field is connected to oppose the emf of the shunt field, it is said to be "overcompound," or "differential compound." If the rated load voltage is greater than no-load voltage, the generator is overcompound. If the series winding could be connected so that the no-load voltage and full-load voltage were identical, the generator would be "flat compound." Figure 11-237 illustrates the relationship between these types of connections. The series field is distinguished by a small number of turns of large wire, and the shunt winding by a large number of turns of small wire. The graph in figure 11-237 shows the relative output voltage under various load conditions for different types of compound generators.

11-1500. DC MOTORS.

11-1501. GENERAL.

11-1502. Since a dc motor is fundamentally the reverse of the dc generator just described, the construction details applying to the generator apply similarly to the motor. When the motor is running, the armature develops mechanical energy at the expense of electrical energy. If the armature and the field windings are connected in series, the motor is a series motor; if they are connected in parallel, it is a shunt motor. In some motors the field windings are in two parts, one in series with the armature and the other in parallel with it. If the motor is so connected, it is compound. The corresponding electrical circuits of the various types are shown in figure 11-236.

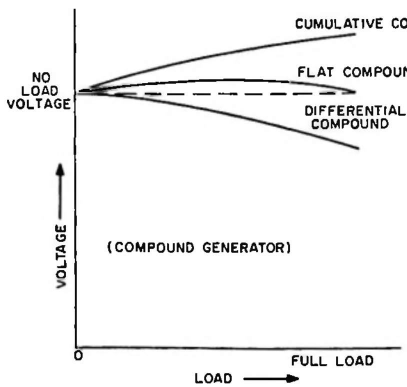


Figure 11-237. Relative Output Voltage Under Various Load Conditions for Different Types of Compound Generators

11-1503. SERIES MOTORS.

11-1504. The series motor is so named because the armature and field are connected in series, as shown at part A of figure 11-236. This method of connection makes it necessary for the field to be heavy enough to carry the armature current. Consequently, the wire in a series field is of large size, and the winding contains relatively few turns. The one serious disadvantage of the series motor is that the speed of the motor varies inversely with the load. Should the load be removed, the increase in motor speed could result in destruction of the motor. Therefore, this type of motor is not adaptable to a constant-speed application, although it is ideal for use where it comes continuously under the control of an operator, and where a high starting torque is required. The torque of a series motor varies as the square of the current. (The torque will increase as the square of the current up to the point where the core material reaches saturation.) Because of its speed characteristics, the series motor is not suitable for any load where the required torque might drop below 15 percent of full-load torque.

11-1505. SHUNT MOTORS.

11-1506. In the shunt-type motor, the field winding is connected in parallel with the armature, as shown in part B of figure 11-236. There are two circuits through the shunt motor, one through the armature and one through the field. The field coils are wound with relatively small wire, and have a large number of turns. This type of winding is necessary to maintain a high flux value and to prevent coil overheating. Because the field is connected across the line, the density of the magnetic field remains constant. Therefore, the torque of a shunt motor must vary with the current in the armature; that is, if the armature current is doubled, the torque is also doubled. Since the field strength is constant, the motor speed will be relatively constant (± 10 percent) from no load to full load. The shunt-wound motor is the type used for constant-speed applications, but, because of the fixed field current, it has a lower starting torque than the series motor.

11-1507. COMPOUND MOTOR.

11-1508. When the series and shunt-type windings are combined in one motor, the result is a machine that has both high starting torque and constant speed. In some applications, the series winding is in the circuit for starting only, and is switched out when the motor is up to normal speed. The compound-type motor does not have a flat constant speed under load, but its variation is not great (between 15 and 20 percent). This variation is dependent upon the number of ampere-turns in the series winding; the greater the number of ampere-turns, the greater the starting torque, and the greater the speed variation (unless the series winding is switched out when the motor reaches normal speed). As the ampere-turns of the series winding are decreased, the speed variation under changing load is decreased. With a weakened shunt field, the series-field flux constitutes the greater portion of the total flux; hence, changes in load may produce unstable speed.

11-1509. UNIVERSAL MOTORS.

11-1510. A universal motor is a series-connected motor that may be operated on alternating or direct current with approximately the same speed and torque characteristics. The armature and field coils of the universal motor are connected in series. A change in current direction, due to the ac input, or commutation action of direct current, results in an in-phase relationship between the armature and the field flux. As the direction of current flow and the direction of magnetic flux in the field and armature are changed, the rotation of the armature will remain the same. Even though the current is reversing 120 times per second (60-cycle input), the universal motor will continue to run in the same direction, because the field flux is always in phase with the armature flux. When a universal motor is run on direct current, the current flowing in the circuit is limited only by its resistance and the self-induced armature voltage. For alternating-current operation, the reactance due to the inductance of the coils absorbs some of the line voltage, resulting in a lower speed on 60-cycle ac than on dc, for a given value of current. In large universal motors, reactance losses are compensated for by an auxiliary winding. This winding is displaced 90 electrical degrees from the main field winding, and the field of this compensating winding counteracts the effect of armature reaction, and also tends to improve commutation. In universal motors, the field and the frame must be laminated, to reduce heating by eddy currents when the motor is used on alternating current.

11-1511. DYNAMOTORS.

11-1512. A dynamotor is a combined dc motor and generator operating in the same magnetic field. It may have two armatures on one shaft or two sets of windings and two commutators on one armature. Dynamotors are usually employed to generate high dc voltage when driven from a low-voltage supply. They are usually employed in mobile equipment when high-current and high-voltage stability are required. Further information regarding the construction and maintenance of dynamotors is given under paragraph heading 11-1459.

11-1513. MAINTENANCE OF DC GENERATORS AND MOTORS.

11-1514. PRELIMINARY CONSIDERATIONS.

11-1515. Before placing into service a new generator, or one that has been shut down for a protracted period, a routine check should be made, according to the recommendations given below.

11-1516. The generator should be clean, inside and out. The interior can be cleaned with a vacuum cleaner, or by the use of clean, dry compressed air at low pressure (25 psi). The nozzle of the cleaning device should be nonmetallic. If a cleaning device with a nonmetallic nozzle is not available, a clean, dry brush should be used. Avoid personal injury while blowing out electrical equipment with air under pressure. Use goggles to avoid eye injury. Greasy dirt may be removed by an approved cleaning fluid. However, cleaning fluid should not be used on brushes or commutators.

11-1517. Any dents or cracks indicating rough handling should be investigated. If moisture appears to be present, it should be removed to avoid the possibility of insulation failure. Infrared lamps may be used for this purpose. In case of emergency, when no drying devices are available, the generator may be run with its armature short-circuited, and with just enough field current to warm the machine if possible (do not exceed 200 percent of the full load value); this should be done at a reduced speed. After drying, remove the armature short; then apply the load gradually, to prevent any sudden current surge that might result in a breakdown of the insulation.

11-1518. Disconnect the coupling between the generator and the power source, and rotate the generator shaft by hand. It should turn freely. Note the end play of the shaft, and the tightness of the mechanical and electrical connections. If the generator is rigidly connected to the power source, turn the crank or shaft, and check for binding. With the coupling connected, check again for freedom of rotation, as there may be misalignment between the two units. A belt-driven unit should have the pulleys in exact alignment, and the belts should depress 3/4 to 1 inch, depending upon the center distance and the pulley size. In no case should the belts be tight. A general visual inspection should be made for loose parts, frayed wire, cracked insulation, loose setscrews, frayed or cracked belts, etc.

11-1519. Electrical equipment operating in a salt atmosphere should be inspected daily for signs of corrosion or salt growths. Open or drip-proof generators or motors should not be used outdoors, or where splashing or hosing down may occur. For operation under water, pressurized equipment should be used, although totally enclosed motors may be used for short periods when flooded or submerged.

11-1520. Electrical equipment operating in dust-laden atmosphere should be inspected daily for signs of dust accumulation. Usually, desert operation does not shorten the life of electrical equipment if the equipment is protected from sand and dust storms. Severe damage can be caused by dust or sand which is carried in large amounts in the atmosphere, for the presence of these particles between moving surfaces is almost sure to cause failure. In some locations, atmospheric dust that is peculiar to that locality can cause corrosion or breakdown of insulation. Coral dust is a conductive material, and may cause arcing. Volcanic dust, in addition to its abrasive qualities, also causes corrosion when it is combined with moisture.

11-1521. Equipment for use in the tropics is usually tropicalized at the time of manufacture. Unfortunately, the tropicalization treatment given to electrical equipment is not effective over a long period of time, and it is necessary to renew the fungicidal varnish or lacquer at frequent intervals. Since extremely humid atmosphere is the condition that causes fungus growth, it is necessary to bake the equipment, in order to remove the moisture, before revarnishing. This fungicidal varnish does not take the place of the usual insulating varnish applied to coils, but is intended to supplement it. The presence of a heavy growth of fungus on or in equipment does not mean that the machinery is ruined. Usually, removal of the fungus, followed by baking and varnishing, will restore the power-generating equipment to service. Do not rebake equipment after the fungicidal coating has been applied, as heat destroys its effectiveness. Certain types of fungus do not require tropical conditions for growth; some types flourish even in arctic regions. If the electrical equipment is to be used where termite infestation is probable, chlorinated naphthalene should be added to insulation varnish. Cellulose-base insulating materials are especially susceptible to attacks by termites.

11-1522. LUBRICATION.

11-1523. The importance of proper lubrication cannot be overemphasized. Insufficient lubrication can cause serious wear in moving parts. Too much lubrication is also damaging, as the lubricant may get into the commutator and brush rigging, where it will serve as a binder for copper and carbon dust collection, thus causing leakage paths and short circuits. Determining the correct amount of lubricant for ball bearings is one of the most important considerations in the maintenance of rotating machinery. Too much lubricant can cause troubles just as readily as too little. Many machines have ball bearings designed to be greased with a pressure gun. Only a high grade of grease, having the following general characteristics, should be used for ball bearing lubrication:

a. Consistency a little stiffer than that of vaseline, maintained over operating temperature range.

b. Melting point preferably about 150 degrees.

c. Freedom from abrasive matter, acid, and alkali.

11-1524. Do not oil a sleeve-bearing generator while it is running, as this will give a false indication of the oil level. Ball-bearing generators should be greased while running, with the relief hole open. Approved cleaning fluid may be used to remove excess lubricant from the interior of the generator. Under ordinary conditions, greasing once a year is sufficient, while under severe service, every 3 months may not be too often.

11-1525. If the bearings have oil cups or reservoirs, any good oil with a viscosity approximating SAE 10 may be used. Turbine oil is excellent for this purpose, and enough should be added to bring the level to the proper point. See figure 11-238. On some generators, prepacked lifetime-type bearings are used, and no cups or fittings are provided; in such cases, the bearing caps are usually so marked.

11-1526. TEMPERATURE CHECKS.

11-1527. Excessive temperature rise is usually the first indication of trouble. A temperature rise of 55 degrees to 65 degrees centigrade is considered normal, with 40 degrees centigrade being considered the ambient temperature. Feeling with the hand is not an accurate means of determining the temperature, but it will serve to indicate a hot spot on which to place a thermometer. A mercury or alcohol thermometer can be used to measure generator temperature if the bulb can be placed against the hot area. The thermometer can be insulated from the surrounding air by means of friction tape or other insulating material. It is advisable to check the temperature frequently over a 3-or-4 hour period, since a generator (if not overloaded) will not come up to, or exceed, its operating temperature for several hours.

11-1528. An excessively high temperature is always an indication of trouble. Overload should be suspected first. The load condition should be checked by means of an ammeter, and the reading compared with the current rating of the generator. Worn brushes and "high mica" contribute to a temperature rise. Clogged cooling vents, bent fan or impeller blades, and dirt are other causes. In two-pole generators, a frequently overlooked cause of temperature rise, brought about by sudden load changes, is oscillation of the armature about its axis. This causes excessive bearing load, and in a sleeve-type bearing, will cause bell-shaped wear of the bearings. Trouble of this type usually produces a rise in bearing temperature. Rubbing between the field pole and the armature may also be caused by sudden load changes, but such rubbing is usually indicated by a knocking noise. This condition may exist even

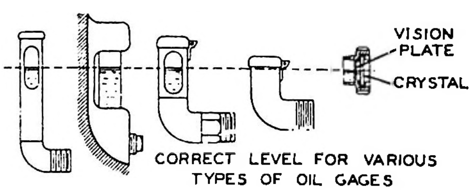


Figure 11-238. Proper Oil Level for Various Lubrication Sight Gages

when there is apparently sufficient clearance between the two moving parts. Knocking with sudden changes in load is due to a design fault, assuming that the shaft has not been sprung. The proper corrective measure is to reduce the load on the generator, and avoid imposing sudden, heavy loads on it.

11-1529. Commutators may overheat if the brushes are not on the proper load plane, if they are making poor contact, or if dirt is clogging the risers (in large commutators). Shorted turns in the coils or shorts to the metal parts will also cause heating.

11-1530. NOISE.

11-1531. Noise is another indication of impending trouble. High mica, incorrect brush angle, out-of-round commutator, excessive surface film, or too much clearance between the commutator and the brush holder will all cause abnormal noise. Noisy bearings may be caused by worn balls or rollers or by lack of lubrication. Too much grease under pressure in a bearing, especially a semi-solid grease, will cause excessive noise. If sleeve-type bearings are worn out-of-round, they will make a knocking or pounding noise. Foreign objects or loose parts inside the generator may cause a clicking noise. Loose laminations or other parts will usually cause a humming or buzzing sound.

11-1532. TEST INSTRUMENTS.

11-1533. A milliammeter can be used to conjunction with a voltmeter to test an armature for shorted coils. The use of an ohmmeter is impracticable for this test, because the resistance of the coils is of such a low value. An ohmmeter is useful mainly for checking the continuity of shunt field coils.

11-1534. A voltmeter can be used to check voltages in any part of the circuit, and can also be used for finding open coils. A megger is used to test insulation resistance to ground. The reading of insulation resistance must be considered in conjunction with other conditions if it is to be of any value. When certain machines come continually under the care of one person, it is advisable to keep records of insulation tests, since only in this way is it possible to have an accurate indication of insulation conditions. The original test may be plotted on graph paper at 15-second intervals for 60 seconds, or readings may be plotted once each minute for 10 minutes, using a motor-driven megger. Figure 11-239 shows typical readings; from these curves it can be seen that insulation resistance increases as the time of dc voltage application is increased. With a motor-driven megger, a steady reading should be obtained in 10 minutes. If a winding is wet or dirty, the steady reading will be reached sooner, as shown by curve 2. After the windings have been cleaned and dried, the reading will probably approximate that indicated by curve 3. It should be noted that insulation resistance varies inversely with the temperature, and that resistance values are given for the basic temperature of 40 degrees (C). The resistance of insulation doubles for each 12 degrees (C) drop in temperature, within the range normally encountered in the operation of these machines.

11-1535. MOTORING TEST.

11-1536. For motoring tests, the generator must be disconnected from its prime mover. To test, connect the generator to a dc voltage source of some value near the rated output voltage (for example, a 150-volt generator to 100-volt source); include an ammeter of suitable range in the circuit. A generator that is in good condition will run as a motor with a current drain of approximately 10 percent of its rated output current. A shunt generator with an open shunt field will not run, and may draw over three times as much current. A generator with a shorted or grounded armature will run unsteadily, and the current will fluctuate from 5 percent to 50 percent of its normal rated output current.

11-1537. TESTING FOR INSULATION BREAKDOWN.

11-1538. Disconnect all electrical grounds in the generator. If the generator is of the open type, or if the brush rigging is accessible and the commutator can be reached, the generator can be tested for grounds without disassembly. A megger may be used, or a voltage source and a voltmeter or lamp may be used. Connect the ground lead of the megger to the shaft or the frame of the generator. Connect the test lead of the megger to the armature copper; then crank the megger and note the reading over a 60-second period. When testing with a megger, if the initial

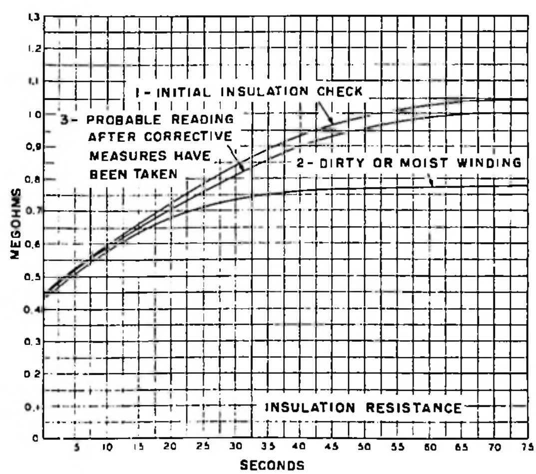


Figure 11-239. Insulation Resistance vs Time

resistance is lower than approximately 1 megohm or is about 1 megohm or more, and then suddenly drops to a low value, an insulation breakdown is imminent. A dead short (zero reading) indicates insulation breakdown.

11-1539. If a breakdown is indicated, isolate the armature from the fields by raising the brushes and placing a piece of insulation under them, to insulate them from the commutator. Connect one lead of the megger to frame, and one to the commutator copper. If a breakdown is now indicated, the fault is in the armature. If not, the lead of the megger should be moved from the armature copper to a field-coil lead; the megger should then give an indication of field-coil insulation breakdown. Generators or motors having defective armatures or field windings must be sent to an electrical repair shop for disassembly and repair.

11-1540. CHECKING COMMUTATOR.

11-1541. Check the commutator for high mica, badly burned or pitted copper, etc. If these conditions exist, it will be necessary to remove the armature, "turn down" the commutator, and undercut the mica. Note whether solder has been thrown out of the risers; this indicates excessive heat at the bars.

11-1542. Sometimes the mica between the segments of a commutator will burn out. There are two possible causes for this condition: (1) the mica may have become oil-soaked; (2) the segments may have loosened, allowing foreign conducting material to become imbedded between them. Repairs can be made by scraping out the burned mica and filling the holes with commutator cement. If high mica, flats, burned bars, or burned mica are in evidence, the generator or motor should be sent to an electrical repair shop for overhaul.

11-1543. The negative brushes of a generator continually deposit material, which is removed by the friction of the positive brushes. (In a motor, the action of the brushes is the opposite; the positive brushes deposit the material, and the negative brushes tend to remove it.) At some point a balance is reached, and the surface film is stabilized. A good commutator film is desirable and necessary. However, some grades of brushes may leave a deposit which is too thick and heavy for quiet operation. The degree of film thickness may be estimated by observing the color of the brush path on the commutator. The desired color when electrographitic or carbon brushes are used is glossy brown or light chocolate. A film that is too thick has a dull-black appearance. With this condition, the brush friction increases, the temperature rises, and there is high-frequency brush chatter and sparking. If this condition is allowed to persist, rapid brush wear or brush destruction will follow.

11-1544. The film thickness may often be controlled by applying a more abrasive grade of brush. This remedy is not always practicable because brushes with added abrasive usually cause greater friction or have unsuitable commutating characteristics. However, when properly applied, they provide a means of film control. When brushes have too much abrasive action, brush chatter may result. In such cases, the brushes will not allow a film to form; it is rubbed off as fast as it is deposited. This is often true when graphite brushes are used. In this case, the commutator will have a dull, brassy appearance. This should not be confused with the desirable lacquered-copper color formed by the proper grade of graphite brushes.

11-1545. CHECKING GENERATOR UNDER LOAD.

11-1546. If the fault is not obvious, and is not readily corrected, it will be necessary to shut down at the first opportunity, for further inspection. If these checks indicate that the generator is normal, start it up and slowly apply a load 25 percent in excess of the name plate rating. Continue at this load for 1-1/2 hours, and observe the machine carefully for signs of trouble. Throughout this test, check the temperature carefully at any questionable spot. After 1-1/2 hours, reduce the load to normal or less. If trouble appears, shut the generator down immediately, and remove for disassembly and repair.

11-1547. TESTING AND SERVICING BRUSHES.

11-1548. GENERAL. In a direct-current generator, brushes usually give the first visual indication of trouble, both in the machine itself and in the equipment to which it is connected. A daily visual inspection of the brushes, while the generator is under load, will often highlight trouble long before it becomes too serious.

11-1549. The proper location for the brush rigging is usually marked with punch marks and paint by the manufacturer. The brushes should not be moved from this position until all other methods of correcting excessive sparking have failed. Brushes may have been displaced from their proper locations with respect to each other because the rigging or holder has been bent out of position. When trouble is indicated by excessive sparking at the brushes, these two conditions should be checked before proceeding further. The brush holder should not be more than 1/16 inch away from the commutator copper. Brush holders are usually staggered, and a slight longitudinal oscillation of the armature is desirable, so that the brushes will not always ride in the same path. This oscillation prevents ridges from forming on the commutator, and also exerts a scrubbing effect that helps to dislodge foreign particles.

11-1550. If the mark on the rigging does not match the mark on the frame, adjust the position of the holder assembly. Check the locations of the brushes with respect to each other by placing a piece of paper tightly about the commutator, and marking the positions of the brushes on the paper. When the paper is removed and laid out flat, the distance between the marks should be equal. Brushes that are binding in their holders cannot exert the proper pressure on the commutator, and the carbon in the brushes can be heated red hot by the lack of proper pressure; this often causes an incandescent arc. The brush holders and the brushes should be clean, and the brushes should fit freely, but not loosely. If the brush holders are clean, sanding of the brushes to make them fit is permissible. Under no conditions should any lubrication be used on the brushes or commutator. Some brushes have a waxy composition, which softens under heat and helps lower friction (but at the same time introducing contact resistance). It is permissible to lubricate slip rings and their brushes sparingly in ac machines.

11-1551. FIRE AND FLASHING. Oil will cause the mica between the segments to break down from bar to bar, or a shorted coil will cause a bar-to-bar flash. Bright flashes are caused by copper particles imbedded in the brushes; the lower resistance of this copper contact causes a heavy current to flow at this point, producing a severe, bright spark when the segment leaves the brush. Copper imbedded in the brushes by "copper picking" will also score the commutator. In such cases the brush should be replaced, or the copper picked out of the brush surface and the brush refitted to the commutator. (Copper picking is the action in which the negative brush in a generator, or the positive brush in a motor, takes on a deposit of copper from the commutator, because of electrolysis.)

11-1552. If the commutator risers designed for cooling are clogged, an excessive temperature will build up at the brushes. High commutator temperature brings about the conditions most conducive to brush disintegration, with excessive sparking and possible fire.

11-1553. Fire and flashing are not to be confused with normal brush sparking. Normal brush sparking can occur at either the leading or the trailing edges of the brushes, depending upon the load plane. Occasionally, when a generator is operating under a variable load, heavy sparking of short duration may be observed. Such effects are caused by sudden changes in the magnitude of the load; the removal of a load can cause sparking just as readily as the sudden application of a load. The removal of a load causes a momentary surge of current during the time the field is relaxing.

11-1554. Before removing any brush mounting, always be sure that it has been properly marked, and that it has a corresponding mark on the frame, so that it may be replaced in its original position. When a generator is being operated under overload conditions, the commutator and brushes should be checked every 5 minutes for signs of overheating; if severe sparking occurs, the load must be reduced, or commutator failure will result. Only under emergency conditions should a generator be operated continuously under overload. Brushes should be replaced when worn down to the mark molded in the surface of the brush. This may be an arrow or a line, or both. If the brush does not have a maximum-wear mark, and if there is any doubt as to its condition, it should be replaced.

11-1555. REPLACING BRUSHES. When replacing brushes, the following points should be observed. If possible, always use complete sets, as one or several abrasive brushes are usually included in a set, to aid in keeping the commutator film at the proper density. The abrasive brush is necessary for proper commutator action, but more than the required number of abrasive brushes would cause excessive commutator wear, and thus would remove too much of the film. Make certain that the pigtail connection to each brush is secure. Remove any roughness inside the brush holders with sandpaper; also, make sure that the brushes and the holders are clean and of the right kind. Insert the brushes in the holders, making sure that they move freely. The brushes should be worked down in the proper fit by the use of sandpaper. In large machines, fit the brushes, one at a time, with sandpaper wrapped half-way around the commutator, or with a seating stone. A seating stone is made of soft, abrasive material. When the stone is held against the commutator, in front of the brush, abrasive material is carried under the brush. In small machines, the sandpaper should be wrapped around the commutator, secured with a rubber band, and the brushes worked down by rocking the armature back and forth. You should use only sandpaper or a seating stone, since other abrasive papers may cause short-circuiting of the commutator.

11-1556. Make certain that the pigtails are securely connected to their terminals. Measure and adjust the brush spring pressure. The proper method of making the measurements is shown in figure 11-240. The following brush pressures are recommended for average use when the manufacturer's specifications are not available:

Electrographitic grade —	1-3/4 to 2-1/2 psi
Carbon and carbon-graphite grade —	1-1/4 to 2-1/2 psi
Graphite grade —	3 to 4 psi
Metal-graphite grade —	2-1/2 to 4 psi

(The tension on the lower brushes of large machines should be made slightly greater than the tension on the upper brushes, to compensate for the effects of gravity.)

11-1557. AC GENERATORS.

11-1558. GENERAL.

11-1559. In direct-current generators, the alternating current which is induced in the armature coils is rectified by the action of the commutator, to obtain the direct-current output. In alternating-current generators, the alternating-current output is obtained by simply taking off the output by means of a rotary collector, which does not rectify. The collector most commonly used for this purpose is a set of slip rings, which are mounted on, and insulated from, the rotating shaft. The brushes which contact the slip rings are mounted on the stationary frame. The armature coil is not closed, as each end is connected to a separate slip ring.

11-1560. A current is induced in a conductor (or coil) when the conductor is moved through a magnetic field; this action is the result of the action of the conductor relative to the field flux. Since the motion of the field flux with respect to the armature coil is relative, and since the armature has no commutator, most alternators have the armature coils mounted on a stationary frame, while the field is rotated. Alternator armatures often develop output voltages as high as 13,200 volts, while the field voltages seldom exceed 250 volts. By using stationary armature construction, direct connections can be made to the high-voltage coils, while the sliding contacts can be safely used to transfer the low exciting voltage to the rotating field. The low voltage required for the field permits the use of less bulky field-coil insulation, which greatly reduces the weight. These lighter coils develop less centrifugal force and have less wind resistance, which permits more efficient use of the driving power. The rotating field in an alternator is called the rotor, and the stationary armature is called the stator.

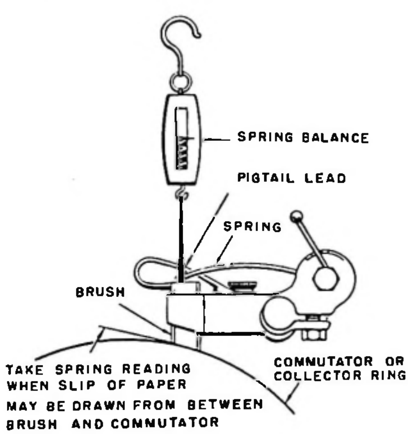


Figure 11-240. Method of Measuring Brush Tension

11-1561. ROTORS.

11-1562. Alternator rotors are divided into two classes, depending upon their operating speeds. If the alternator is to be driven at slow speed, a "projecting-pole," or "salient-pole," rotor is used. The field poles are formed by fastening a number of steel laminations to a spoked frame, or spider. The heavy pole pieces produce a flywheel effect on the slow-speed rotor, to help keep the angular speed constant. When an alternator is connected to a line which is being fed by other alternators, it is, of course, in parallel with these other machines. Unless the speed of the added alternator were controlled within exceedingly narrow limits, the speed would vary, and, consequently, the frequency and voltage of this alternator would vary in relation to the line frequency. To prevent this condition from occurring, alternators are provided with "amortisseur," or "damper" windings, located on the rotor to stabilize the speed of the alternator. These damper windings consist of a number of copper bars located in slots in the face of the rotor pole pieces. The bars are parallel to the axis of the rotor, and are connected together at the ends by means of a shorting bar or ring, as in a squirrel-cage winding.

11-1563. High-speed alternators are of the non-salient or smooth-rotor type. These rotors are made from a solid steel forging or a number of steel disks fastened together, with the field coils locked in slots which are milled in its surface. These coils are usually distributed field windings, which provide a sinusoidal flux distribution around the rotor. A smooth rotor is used in high-speed machines because it has less air-friction loss, and the windings may be so placed that they can withstand the centrifugal forces developed at high speeds.

11-1564. STATORS.

11-1565. In a rotating-field alternator, the armature windings are stationary, and are called the stator. The armature iron, being in a moving magnetic field, is laminated, in order to reduce eddy-current losses. In ac generators to be used with high-speed turbines, the stator laminations are ribbed, to provide sufficient ventilation, because the high temperature developed in the windings cannot be dissipated in the small air gap between the rotor and stator. In some of the large installations, the alternators are totally enclosed, and are cooled by hydrogen gas under pressure, which causes lower windage losses than air, and has greater heat-dissipating properties. The stator coils in a high-speed alternator must be able to withstand the heavy stresses caused by extremely large currents flowing in the coils. The coils must also be well braced, to prevent them from being pulled out of place when the alternator is operating under a heavy load.

11-1566. EXCITERS.

11-1567. A self-excited direct-current generator takes the rectified current from the commutator brushes to supply dc excitation to its own field. Alternators also require dc for their fields, but, since most of them do not have commutators, the dc field current must be obtained from an external source. The exciter used to supply this current is usually a flat-compound-wound dc generator designed to supply 125 to 250 volts. The exciter rotor may be mounted directly on the rotor shaft of the alternator, or it may be belt-driven. Large installations have separate motor-generator sets or separate exciter bus lines. The power for field excitation seldom exceeds 2 percent of the alternator rating, and larger machines require only 1 percent.

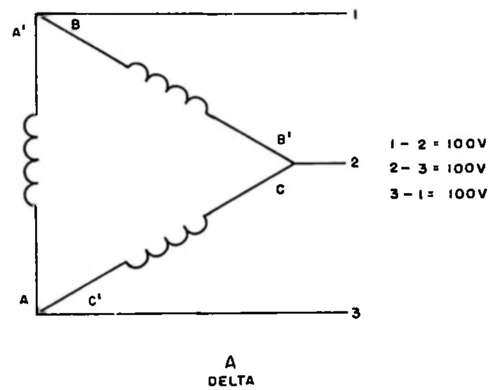
11-1568. REVOLVING-FIELD ALTERNATORS.

11-1569. GENERAL. Alternators are designed to produce voltages which have as perfect a sine wave as possible, because this output characteristic is the most desirable for general-purpose applications. The waveform of a voltage produced by an alternator is determined by the type of magnetic field and the type of stator winding used. Basically, the flux distribution in the air gap determines the shape of the voltage wave produced. Flux distribution is controlled by rounding off the pole pieces (projecting pole types) or by the distribution of the field windings (distributed field types).

11-1570. FIELD WINDINGS. There are two major types of field windings commonly used today: the distributed field type and the conventional projecting-pole-field type. If the machine is designed to deliver low power, the coil wire is cotton covered. Higher-output machines use flat copper straps, which are wound around the poles and insulated with high-temperature-resistant bonded mica strips. The field coils must be so connected that they have alternate north and south poles. The field coils in non-salient rotors are made of flat straps of copper fastened in slots that have been milled in the surface of the rotor. Since the pole pieces are a part of the forging itself, it would be difficult to round them off in the manner of a salient-pole rotor. To compensate for this, the windings are modified to produce the same effect. When the field flux of each set of coils is combined, this forms a resultant flux that is capable of inducing a sinusoidal wave in a conductor. In practice, more than three coils are used, and the combined output produces a nearly perfect sinusoidal field flux. This gives the same results as rounding the edges of the poles in a salient-pole rotor.

11-1571. ARMATURE WINDINGS. Although one, two, and three-phase alternators are in common use, the most popular is the three-phase alternator because it is the most efficient. Three-phase alternators have three separate sets of armature coils that are insulated from one another, and give the same effect, electrically, that three separate alternators would produce.

11-1572. Most three-phase alternators have six armature leads brought out to a terminal box. There are three pairs of leads, one pair for each coil or phase winding. The separate leads make it possible to connect the armature coils in a star or delta network. This is a method used to balance three-phase circuits so that all phase voltages are equal in magnitude and differ in phase by equal angles. The three phases are usually distributed between three lines, although some star connections have four wires when a common line is used. The common line is generally used to ground the neutral point of the windings. Figure 11-241 shows how a star or delta network may be connected; however, the circuit must be connected so that the proper phase relationships exist. If the alternator is to be delta-connected, as in part A of the figure, coils AA' and BB' should be connected together at A'B. The voltage across AB' should be equal to the voltage across a single coil. If the voltage is higher, the leads of one of the coils should be reversed. Now, one end of coil CC' should be connected to A, leaving junction B'C open. If the voltage across B'C is twice the voltage of a single coil, the leads of coil CC' should be reversed. If the voltage across B'C is approximately zero, then B' and C can be connected together. In the delta network, the voltage across any two lines will equal the voltage across a single coil. In the balanced delta method, the line current is $\sqrt{3}$ (1.73) times the current of a single-phase winding.



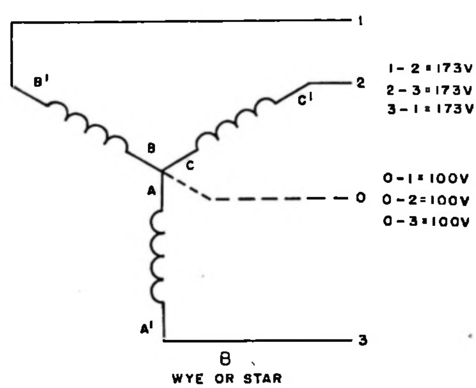


Figure 11-241. Delta and Wye (Star) Connections of Three-Phase Alternator

11-1573. Part B of figure 11-241 shows a three-phase alternator having a star, or wye connection. First, the two coil ends A and B should be connected together so that the voltage across A'B' will equal 1.73 times the voltage across one coil. If the voltage across A'B' is the same as the voltage across one coil, it will be necessary to reverse the leads of one of the coils. Next, C of the coil CC' should be connected to the junction AB, so that the voltage across A'C' and B'C' will equal $\sqrt{3}$ (1.73) times the voltage across a single coil. If the voltage is less, the leads of coil CC' should be reversed. In the star, or wye, network, the voltage across any two lines will be equal to the vector sum of the voltages of the two coils, which is the same as the voltage of one coil times 1.73. If a common lead is used, single-phase power can be obtained by connecting the external load to one of the outside lines and to the common line. The common lead in a star circuit is usually grounded.

11-1574. AC MOTORS.

11-1575. GENERAL.

11-1576. The induction motor has long been known for dependable, trouble-free service. This type of motor is used where small and medium-sized ac motors are needed. The speed is controlled by the frequency of the supply voltage, and remains constant over a wide range of loads. Most small motors operate on single-phase power, while the larger ones operate on two-phase or three-phase power. The induction motor, as the name implies, operates on the same principle as a transformer, with the stator acting as the primary winding and the rotor acting as the secondary winding. There are no connections between the stator and the rotor; all voltages in the rotor are created solely by induction.

11-1577. CONSTRUCTION.

11-1578. Most induction motors differ from other types of motors in that commutators and brushes are not used. The two basic parts of an induction motor are the stator and the rotor. The rotor can be wound with wire, or formed with solid bars, as in a squirrel-cage winding. The core is made up of a number of laminated steel discs with the coil slots punched in them. The stator core is made up of a number of laminated steel rings, with the coil slots punched in them around the inside bore. The type of slot used depends upon the size and application of the motor. The semiclosed slots provide higher efficiency, because they offer a greater stator surface area adjacent to the rotor than is possible with open slots, a condition which results in higher output. In semiclosed slots, the coils are placed in the slot through the narrow opening, one at a time. In this case, the coils must be insulated after they are formed, and placed in the slot, an operation that requires time and expense. Although semiclosed slots permit more efficient operation, open slots are usually used in the larger motors, in order that the heavy copper straps which are used for the coils may be machine-wound, and insulated, before being placed in the slots. In a squirrel-cage rotor, solid bars of aluminum, copper, or an alloy are forced into the slots, to insure a tight fit. The bars are then fastened to short-circuiting rings, which are placed at each end of the rotor core. In small motors, the bars and end rings are often made of a single casting.

11-1579. OPERATING PRINCIPLES.

11-1580. When an alternating current is applied to the primary of a transformer, a varying magnetic field is established, which induces a voltage into the secondary winding. When a voltage is induced in a coil situated in a magnetic field, a force is set up which tends to produce motion in the coil relative to the field. If the secondary windings were free to move, they would do so; however, they would come to rest as soon as they were outside the influence of the magnetic field. If the secondary is to continue moving, the primary must be moved so as to keep the secondary within the magnetic field.

11-1581. To explain how torque is produced, the induction motor may be considered as consisting of a horseshoe magnet and a disk. For purposes of explanation, however, assume that the disk and the horseshoe magnet are both free to rotate on a common axis. Since the pole faces of the magnet are separated from the surface of the disk by only a small air gap, the flux lines of the magnet pass through the disk. As the magnet is rotated, the magnetic field induces eddy currents in the metal disk. These induced currents follow definite paths in the disk, as though they were flowing through regular conductors. According to Lenz's law, the current induced in a conductor as a result of its motion in a magnetic field is in such a direction as to exert a mechanical force opposing the motion. In the case under consideration, the fact that the disk is initially stationary and the magnetic field is rotating is of no consequence, because all motion is relative. The eddy currents induced in the disk exert a force that opposes the motion of the magnet. But, since the disk is not held stationary, the opposing force in the disk causes it to revolve.

11-1582. In a constant magnetic field, motion of the field, or motion of a conductor in the field, will cause current to be induced in the conductor. The direction of the induced current in the conductor will be such that, by its electromagnetic action on the field, it will cause the conductor to move in the same direction as the magnetic field. The magnet and the disk will both be turning in the same direction; however, when the disk is turning more slowly than the magnet, the relative motion of the disk is in the opposite direction, with respect to the direction of rotation of the magnet. In other words, if relative motion were maintained between the magnet and the disk, and the magnet stopped rotating relative to an observer, the disk would be rotating in the opposite direction. The relative motion between the magnet and the disk, usually referred to as the slip, is important, because it is this motion which induces the current into the disk, to produce the operating torque. As the slip increases, and this action continues until a breakdown torque is reached, at which point the rotor comes to a standstill.

11-1583. A commercial induction motor operates in a manner very similar to that of the simple induction motor just described, with the exception that it is unnecessary to rotate the stator to obtain a rotating field. There are no pole pieces in the stator of an induction motor. Instead, a distributed winding, similar to the stator of a universal motor, is used. The coil groups in the stator are lap-wound, and these groups are connected so as to produce the desired magnetic poles. Any number of poles may be formed by connecting the coils together properly. The stator core remains stationary, but it produces a magnetic field which rotates as if the entire stator were turning. The ability of magnetic fields to add together or cancel out makes it possible to create smoothly rotating field poles.

11-1584. When the motor is running with no load, the rotor will increase its speed to nearly that of the rotating magnetic field. If the rotor speed equaled the speed of the rotating field, there would be no slip; consequently, no voltage would be induced in the rotor windings, and there would be no torque, because the conductors would be cutting no flux lines. Therefore, the rotor would slow down until there was sufficient slip to develop the necessary torque. At a no-load condition, very little torque is required; as stated previously, under no-load conditions the rotor speed nearly equals the speed of the revolving field.

11-1585. If a load is placed on the motor, the rotor slows down, thereby increasing the slip. As a result, the rotating field cuts the conductors at an increased rate, inducing higher rotor currents, and providing more torque. As the load increases, the rotor keeps slowing down, and, as the slip increases, more torque is developed. However, there is a limit to the amount of torque a motor can develop; this limit is called the breakdown torque. When the breakdown torque is reached, the motor comes to a standstill until the load is decreased or removed.

11-1586. As the percentage of slip increases, the frequency of the induced rotor current increases. For example, if the frequency is 3 cycles at 5-percent slip, it will be approximately 6 cycles at 10-percent slip. As the frequency of the induced voltage keeps increasing, the inductive reactance of the rotor winding increases, and this, in turn, increases the phase angle between the rotor current and the field flux. As long as the increasing rotor current has more effect on the torque than the increasing phase angle, the torque increases. At the breakdown point, the increased torque due to the increased rotor current is neutralized by a negative torque, which is the rsult of the increasing phase angle.

11-1587. The starting torque of a single-squirrel-cage motor is relatively low. As the motor is started, the frequency of the induced voltage in the rotor is nearly equal to that of the supply voltage. The inductive reactance is relatively high, because of the high frequency of the induced voltage; although the reactance may be only a fraction of an ohm, it is large as compared with the resistance of the rotor conductors. The rotor currents, therefore, lag the induced voltage and the field flux by a large angle. Although the current in the rotor is high, almost as much torque is produced in the negative direction as in the positive direction. The negative torque tends to cancel out the positive torque, leaving a low resultant torque. For given load conditions, adequate starting torque is obtained by rotor design which incorporates a suitable amount of resistance.

11-1588. The rotor in a double-squirrel-cage motor contains two sets of rotor bars. One set of bars, which has a comparatively small cross-sectional area, is placed in the slots close to the surface of the rotor; the other set, which has a larger cross-sectional area, is placed deeper into the slots. The set with the smaller cross-sectional area has a resistance of a few tenths of an ohm, while the other set has a resistance of a few thousandths of an ohm. The larger number of flux linkages around the lower conductor gives this set a higher inductance. The high frequency at starting speeds causes the inductive reactance of the low-resistance winding to be higher than the resistance of the top winding. In starting, therefore, most of the current flows through the top bars. The higher resistance of these bars tends to reduce the phase angle between the rotor current and the field flux, and this increases the starting torque. As the rotor comes up to maximum speed, the frequency of the induced voltage becomes lower, and the inductive reactance of the bottom bars is reduced. Since the reactance of both sets of bars is relatively low, the current flow is now limited largely by the ohmic resistance of the bars. Most of the current now flows through the bottom bars, since they have the lowest resistance.

11-1589. Under no-load conditions, the double-squirrel-cage motor operates as a normal single-squirrel-cage motor, with most of the current flowing through the bottom bars. Under varying load conditions, the current automatically divides between both sets of bars in the proper proportions to produce the required amount of torque.

11-1590. The double-squirrel-cage motor has medium-high starting torque and moderate speed characteristics. If very high starting torque and moderate speed control are desired, a wound rotor is recommended. Motors with this type of rotor are useful for air compressors and on other machinery where the load is connected at all times.

11-1591. Wound-rotor motors are always operated on three-phase power. The stator is wound in the same manner as the stator of a three-phase squirrel-cage motor; the rotor, however, is wire-wound, and is connected into three-phase groups. The leads from one end of the three-phase groups are star-connected, and the other three leads are connected to three slip rings. Three rheostats are star-connected, through the slip rings, to the rotor windings. All three rheostats are mounted on one shaft, so that they may all be adjusted simultaneously. The motor is started with the full resistance in the rotor circuit; as the motor comes up to full speed, the resistance is reduced until it is out of the circuit entirely. The starting current is not much greater than the full-load current. The coils in each phase group are connected so that the current flowing through the rheostats is small. The wound-rotor motor is a variable-speed motor, since its speed can be controlled by varying the external resistances. Although this type of motor has a higher starting torque than single- and double-squirrel-cage motors, it is not as efficient at running speeds, because it is not possible to have as low a resistance in a wound rotor as in a squirrel-cage rotor.

SECTION X

INFRARED EQUIPMENT TESTING

11-1592. GENERAL.

11-1593. PURPOSE AND USE.

11-1594. Infrared radiations extend from the limit of the visible red region of the electromagnetic spectrum to the upper edge of the radio microwave region. Considering that infrared waves are not visible to the human eye, especially designed equipment is required to detect these extremely useful radiations and provide a presentation for analysis. Infrared radiations have characteristics similar to those of visible light rays; they can be reflected by mirrors and refracted by lenses. In addition, these radiations can be transmitted through substances such as silicon and germanium, which are opaque to visible light. Therefore, they can also be compared to rf radiations.

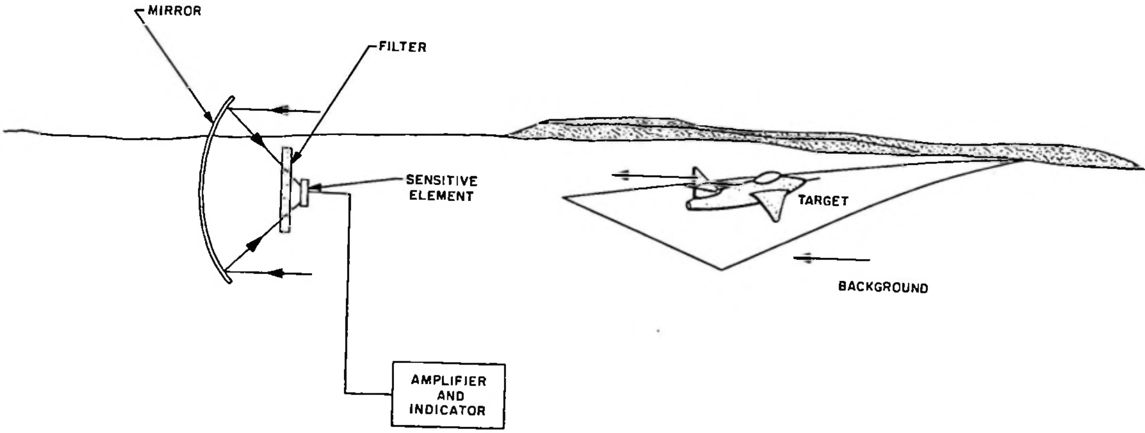
11-1595. Infrared equipment is designed to create, control, or detect invisible infrared radiation. The infrared transmitter (source) equipment is designed to produce and direct the radiations, whereas receivers are designed to detect and convert the radiations into either visible light, for viewing purposes, or into voice or code signals, for audible presentation.

11-1596. You can use infrared devices for weapon guidance, detection of enemy equipment and personnel, navigation, recognition, aircraft proximity warning, and communications. You can use this equipment for either passive or active applications. The active method employs both infrared transmitter and receiver equipment, whereas the passive method requires only receiver equipment. Figure 11-242 shows the basic components required for both active and passive methods.

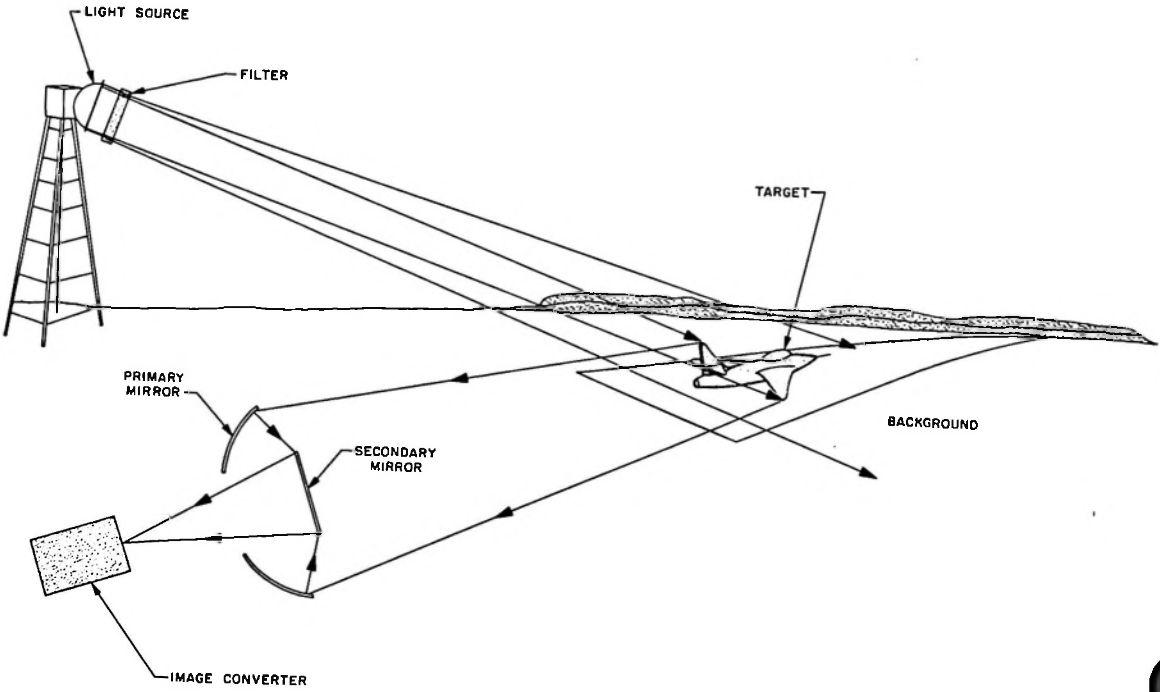
11-1597. INFRARED BANDS. The infrared spectrum is often divided into the far, middle, and near bands, as illustrated in figure 11-243. The devices used in the near and middle bands are employed for ranging, recognition, and communications. The usable distance range of near infrared equipment is normally between 6.5 and 10 miles. Equipment which operates in the far infrared band is used for ranging, missile guidance, and the detection and location of personnel, tanks, aircraft, and ships. The usable distance range is between 100 yards and 12 miles. The bulk of target radiation is at the lower wavelengths, but present infrared detectors and optical components do not operate efficiently in the far infrared band.

WARNING

You should avoid being exposed to the beam of an infrared transmitter. Usually, the danger is not great because you will feel the heating effects of the beam and thus be alerted before damage occurs. Unfortunately, however, the eyes can be damaged before the physical heating effect of the radiation provides sufficient warning. Therefore, you should be particularly cautious and not look into a source of intense infrared radiation.



A
PASSIVE INFRARED METHOD



B
ACTIVE INFRARED METHOD

Figure 11-242. Basic Components of an Infrared Facility

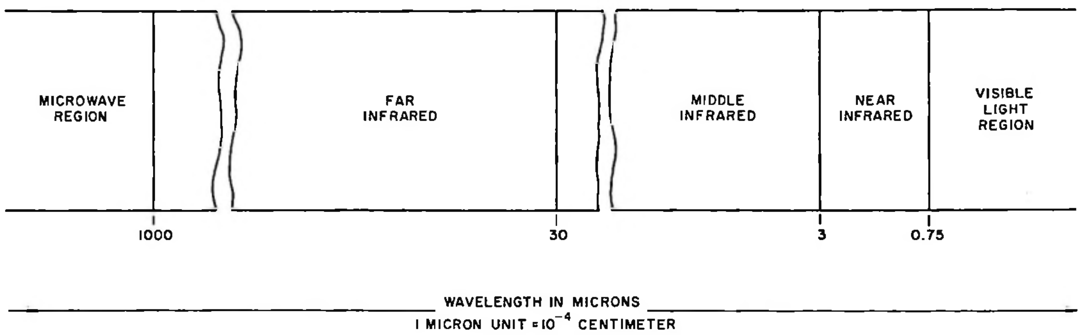


Figure 11-243. Infrared Spectrum

11-1598. DEFINITION OF INFRARED TERMINOLOGY.

11-1599. Many of the terms used in infrared technology are different from those you commonly employ in the field of communications and electronics. Therefore, these terms are listed below for you as a convenient reference:

a. ABSOLUTE ZERO (-273.16°C). This is regarded as the lowest possible temperature.

b. ABSORPTION - The process whereby some or all of the energy of electromagnetic radiations is transferred to the material on which the radiations are incident or through which they traverse.

c. ALBEDO - The reflecting power of an object. Specifically, it is the ratio of the radiation reflected from an object to the total amount of radiation incident upon it.

d. ANGSTROM - A unit of length used in expressing wavelengths of light. It is equal to 0.00000001 centimeter (10^{-10} meter).

e. ATOMIC SPECTRUM - The spectrum of radiation emitted by an atom, caused by changes within the atom, in contrast to the radiation due to changes in the condition of a molecule.

f. ATTENUATION - The reduction in flux density with distance from the source.

g. BLACK BODY - An ideal body which would, if it existed, absorb all and reflect none of the radiation incident upon it. Its reflectivity would be zero, and its absorptivity would be 100 percent.

h. CHOPPER - A device which imparts a pulsating characteristic to a current beam of radiation by a regular and frequent interruption.

i. COLLIMATOR - An optical apparatus for producing parallel beams of radiation.

j. DARK CURRENT - Current that flows in photoemissive and photoconductive detectors in the absence of incident radiation.

k. DETECTOR - Any device that indicates the presence of infrared radiation without necessarily yielding a quantitative measurement.

l. DIFFRACTION - A phenomenon arising from the interruption of a beam of radiation by an opaque obstacle. When radiation, such as visible light, passes by the edge of an opaque body or through a narrow slit, the rays appear to be deflected, producing fringes of parallel light and dark bands.

m. DIFFUSION - The scattering of radiation when it is reflected by a rough surface or transmitted through a translucent material.

n. DISPERSION - The process of separating a beam of radiation into its component parts, in accordance with a characteristic such as wavelength, frequency, or energy.

o. EINSTEIN PHOTOELECTRIC EQUATION - An equation giving the kinetic energy of an electron ejected from a photocathode as a result of its absorbing all the energy of an incident photon (the photoelectric effect).

p. ELECTRON IMAGE TUBE - A cathode-ray tube which produces a visible image from invisible radiation, such as infrared.

q. EMISSIVITY - The radiant energy which comes from the target, divided by the radiant energy which would come from a black body at the same temperature as the target and filling the same field of view.

r. EMITTANCE - The power radiated per unit surface area. This may be expressed in terms of the radiant emittance per unit range in wavelength, the spectral radiant emittance, or the total radiant emittance (which is the integral of spectral radiant emittance for all wavelengths).

s. ENERGY LEVEL - A stationary state of energy of any physical system.

t. F NUMBER - The ratio of the focal length of a lens to the diameter of the aperture through which radiation passes.

u. FOCAL LENGTH - The distance from the optical center of a lens to the point where the light rays converge.

v. GRATING - Generally, any framework, or lattice work, consisting of a regular arrangement of bars, rods, or other long, narrow objects with interstices between them. A diffraction grating consists of rulings upon the surface of a light-transmitting or light-refracting substance, and is used for the production of spectra.

w. HALL EFFECT - The development of an electromotive force between the two edges of a strip of metal in which an electric current is flowing longitudinally, when the plane of the strip is perpendicular to a magnetic field (in which the strip is located). The electromotive force is proportional to the product of the current, the magnetic force, and the sine of the angle between the directions of these quantities.

x. HEAT CAPACITY - The amount of heat necessary to raise the temperature of a unit quantity of a medium (system, entity, or material) 1 degree.

y. INCIDENCE, ANGLE OF - The angle at which radiation strikes a surface, measured from the line of direction of the incident radiation to a line perpendicular to the surface at the point of impact.

z. INTENSITY OF RADIATION - The radiant energy emitted in a specified direction per unit time, per unit area of surface, per unit solid angle.

aa. IRRADIATION - The action or process of being exposed to radiation.

ab. JOHNSON (THERMAL) NOISE - The noise caused by the thermal agitation of charges in a conductor. It is proportional to the absolute temperature and the frequency bandwidth over which the noise is measured.

ac. LINE SPECTRUM - A spectrum produced by radiation in which the energy values of the property being measured (for example, energy, mass, or any related quantity) cluster about one or more discrete values, as contrasted with a continuous spectrum.

ad. LUMINESCENCE - The emission of light due to any cause other than high temperature.

ae. METASCOPE - A small, hand-held, infrared monocular viewing instrument for detecting infrared sources and for observing areas illuminated by near-infrared sources.

af. MICRON - One millionth of a meter.

ag. NOISE EQUIVALENT POWER - A measure of a detector's useful sensitivity in terms of its response to infrared energy as well as its internal noise. It is the infrared power, in watts, which must impinge on a detector to generate a signal equal to its rms noise level over a reference bandwidth.

ah. PHOTOCATHODE - In photoemissive detectors, the electrode which emits electrons upon irradiation.

ai. PHOTOELECTRIC EFFECT - The electrical effect of light or other radiation. These effects can be emission of electrons, generation of voltage, or change in electrical resistance upon exposure to light.

aj. PHOTOELECTRIC THRESHOLD - The quantum of energy, $E_o = h_{eo}$, which is just sufficient to release an electron from a given system (by the photoelectric effect).

ak. PHOTOELECTRON - An electron emitted from a substance as a result of incident radiation.

al. PHOTOEMISSIVE EFFECT - The ejection of electrons from a substance as a result of radiation incident upon the substance.

am. PHOTON - A quantum of electromagnetic energy.

an. PLANCK'S LAW - The basic law of the quantum theory, expressing the concept that energy transfers associated with radiation, such as light, infrared, or X-rays, are made up of definite quanta or increments of energy (E) proportional to the frequency (v) of the radiation. In formula, it states that E = hv, where h is Planck's constant (6.62×10^{-27} erg seconds).

ao. PRISM - In an infrared system, a transparent solid, cut at precise angles, which is used for the refraction or dispersion of radiation.

ap. PYROMETER - A device for measuring high temperatures.

aq. QUANTUM THEORY OF SPECTRA - The theory that an atom or molecule radiates or absorbs energy as it transfers from one energy level to another. The frequency (v) of the radiation associated with such change of energy level is given by the equation:

$$E_1 - E_2 = hv$$

where E_1 and E_2 are the energy levels, and h is Planck's constant.

ar. RADIANCY - The rate of emission of energy per unit area.

as. RADIANT ENERGY - Energy transmitted as electromagnetic radiation.

at. RADIANT FLUX - In infrared systems, the time rate of flow of infrared energy.

au. RADIATION - The emission and propagation of energy through space or through a material medium in the form of waves; for example, the emission and propagation of electromagnetic waves, or of sound and elastic waves.

av. RADIOMETER - An instrument for detecting, and usually also for measuring, infrared radiation.

aw. REFLECTANCE - The ratio of the reflected radiant flux to the incident flux.

ax. REFLECTION - In infrared, the throwing back of radiation by a surface.

ay. REFLECTOR - Any substance, surface, or mechanism that reflects radiation or sound.

az. REFRACTION - The change of direction which radiation undergoes in passing obliquely from one medium to another in which its velocity of propagation is different.

ba. RESISTIVITY (SPECIFIC RESISTANCE) - The reciprocal of conductivity; resistivity is defined by the expression:

$$R = p \frac{1}{A}$$

where R is the resistance of a uniform conductor, 1 is its length, A is its cross-sectional area, and p is its resistivity. Resistivity is usually expressed in ohm-centimeters.

bb. RESPONSE - A quantitative expression of the output of an equipment or device as a function of the input under the conditions stated.

bc. SCATTERING - The random distribution of infrared radiation, caused by atmospheric particles of various sizes.

bd. SPECTRA - Alternative plural of spectrum.

be. SPECTRAL EMISSIVE POWER - The emissive power of a body, at any given wavelength.

bf. SPECTRAL SENSITIVITY - The sensitivity of a detector in relation to the wavelength of the incident radiant energy.

bg. SPECTROSCOPE - An optical instrument for producing and viewing spectra.

bh. SPECTRUM OR SPECTRA - A continuous range of frequencies, within which waves have some specified common characteristic; for example, audio-frequency spectrum, infrared spectrum, etc.

bi. STEFAN-BOLTZMANN LAW OF RADIATION - The energy radiated in a unit time by a black body, as given by the equation:

$$E = K (T^4 - T_o^4)$$

where T is the absolute temperature of the body, T_o is the absolute temperature of the surroundings, and K is a constant.

bj. SUPERCONDUCTIVITY - The abrupt appearance of high electrical conductivity in certain metals when they are cooled through a very low characteristic transition temperature.

bk. THERMAL RADIATION - A form of radiation emitted from a body which is not at absolute zero because of the thermal agitation of its molecules or atoms. This radiation ranges in wavelength from the longest infrared to the shortest ultraviolet rays, its spectral energy distribution depending upon the nature of the body and its temperature.

bl. THRESHOLD FREQUENCY - The lowest frequency of radiation resulting in the emission of electrons from a given surface.

bm. TRANSDUCER - A device, such as a detector, that transmits power from one system to another.

bn. TRANSMITTANCE - The ratio of the radiant power transmitted by a body to the total radiant power entering that body.

bo. VIEWER - An infrared device for seeing in the dark without visible radiations.

bp. WAVE MOTION - A progressive disturbance propagated in a medium by the periodic vibration of the particles of that medium.

bq. WIEN LAWS - Three laws relating to the radiation from a black body. (1) The wavelength of the spectral distribution for which the radiation has greatest intensity is inversely proportional to the absolute temperature of the black body. (2) The emissive power of the black body within the maximum-intensity wavelength interval is proportional to the fifth power of the absolute temperature. (3) The spectral energy distribution of the radiation from the black body at temperature T is expressed by the formula:

$$dE_\lambda = A\lambda^{-5}\epsilon^{-B/\lambda T}d\lambda$$

in which dE_λ is the emissive power within the wavelength interval $d\lambda$, and A and B are constants to be empirically determined.

br. WINDOW - A substance that transmits infrared radiation of wavelengths that depend upon the composition of the window material.

11-1600. The electromagnetic radiation described as infrared exists in a broad frequency range, as illustrated in figure 11-243. Infrared radiation does not consist of heat waves, but is the result of molecular agitation on the surface of a heated object. The greater the heat of an object, the more the radiation emitted. These radiations are converted into thermal energy whenever they strike an object. From this discussion you would assume that the infrared energy emitted from an object is a direct function of the absolute temperature of that object. However, actual radiation depends on the smoothness of the radiation surface, as well as the surface heat. The smoother the object surface finish, the less the radiated energy provided for a given temperature. The emitted infrared energy is attenuated and scattered over some portions of its spectrum, and not attenuated very much nor scattered over other portions of its spectrum. The atmosphere filters some parts of the infrared spectrum and permits other parts to pass freely. Those parts of the infrared spectrum which are not affected by the atmosphere are narrow at sea level, and they become broader as height is obtained until, at approximately 30,000 feet above sea level, only two narrow attenuation bands remain.

11-1601. You will notice that all bodies, either hot or cold (above absolute zero), will radiate some heat. Even ice, at the melting point, is approximately 273 degrees Kelvin above absolute zero, and will radiate a certain amount of heat. Much of this radiation is in the form of infrared, and can be detected by infrared devices. A typical experiment which you can perform in the field to measure the infrared with an ammeter is illustrated in figure 11-244. As shown, a prism is used to separate the beam of light into specific wavelengths. Normally, this separation is accomplished by means of a spectroscope or spectrometer. If you place a camera on the spectroscope or

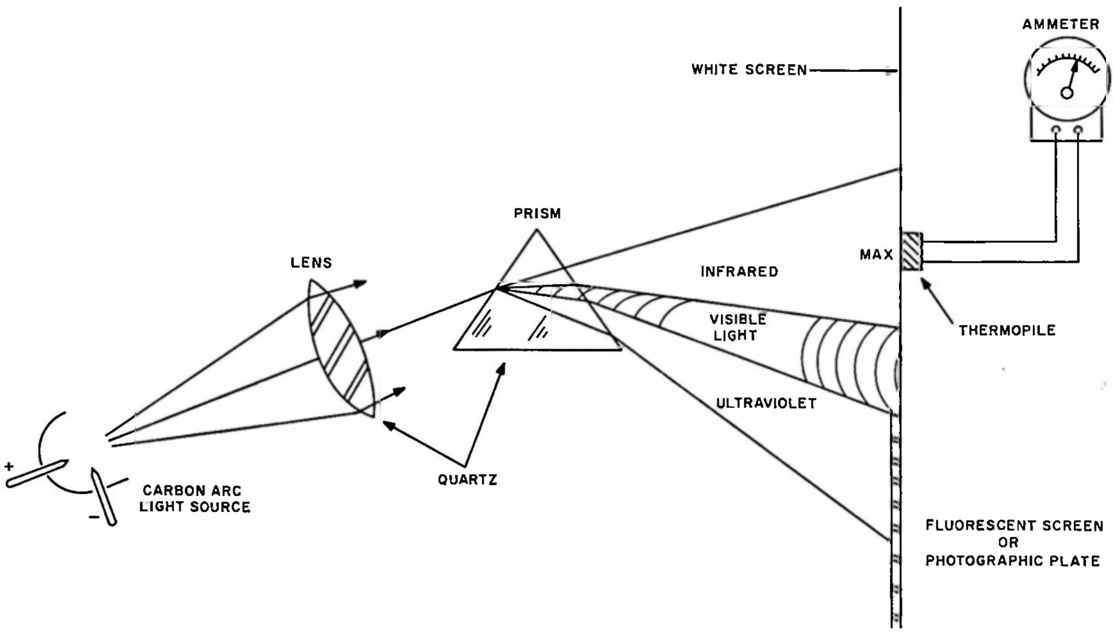


Figure 11-244. Spectrum Analysis Experiment

spectrometer to obtain an image of the spectrum on photographic film, you have an instrument called a spectrograph, and the photographic record is called a spectrogram.

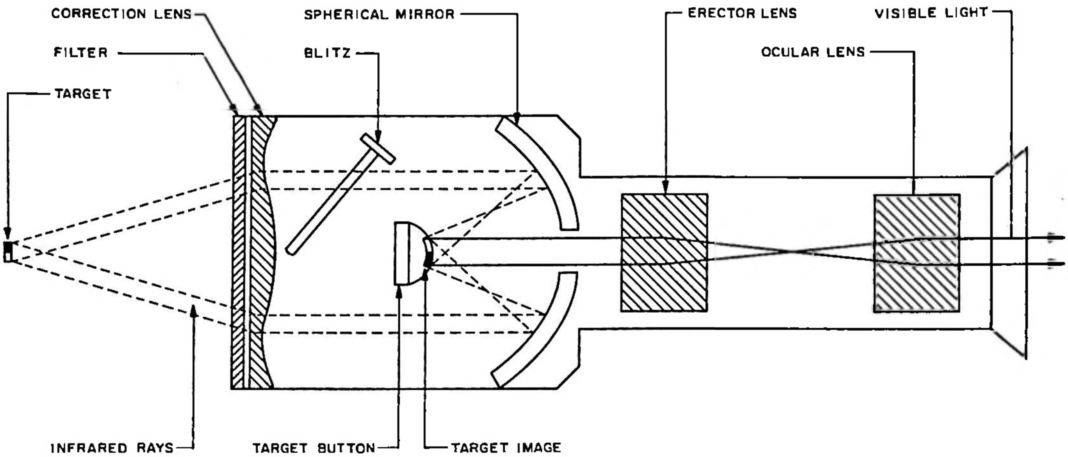
11-1602. INFRARED RECEIVERS.

11-1603. GENERAL.

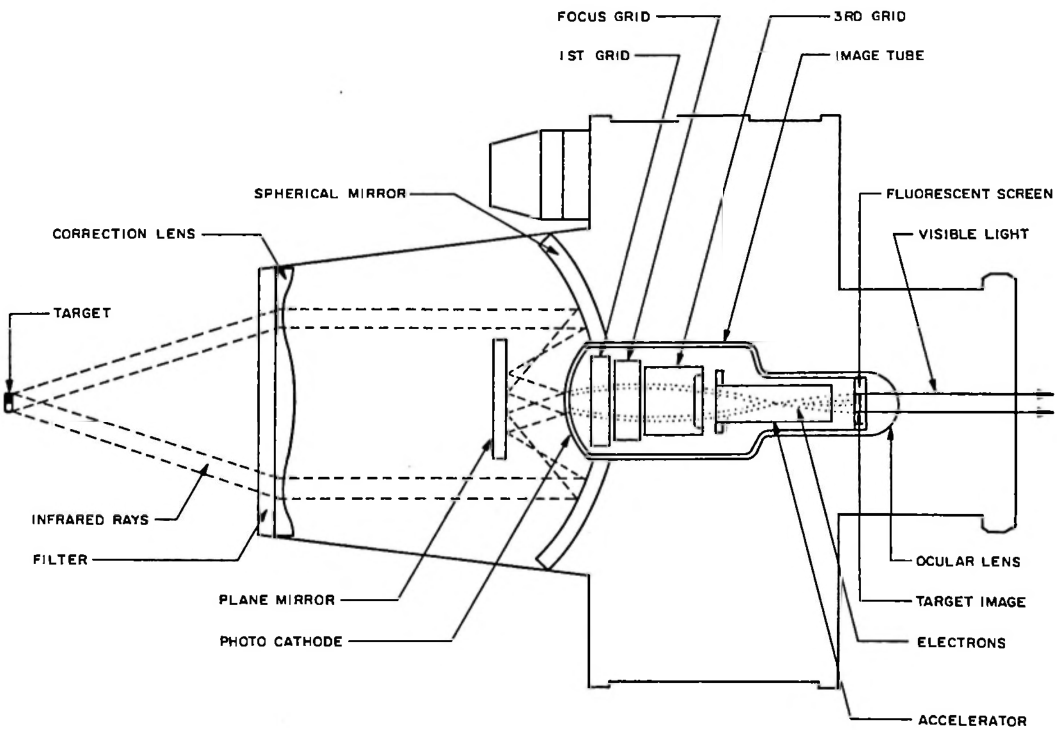
11-1604. Infrared receivers are separated into two distinct groups. The phosphor button type is the simplest, containing a small disk composed of infrared sensitive phosphor, which converts the invisible infrared radiation into visible light. The second group, termed the electronic type, contains an image tube rather than a phosphor button to convert the radiation into light.

11-1605. PHOSPHOR BUTTON RECEIVER. As shown in part A of figure 11-245, infrared rays entering the glass front of the receiver strike a spherical mirror and are reflected back to the phosphor button. The rays, on contact, cause the button to emit visible light, which can be viewed through a hole located in the center of the spherical mirror. That part of the receiver which gathers the infrared rays and directs them to the phosphor button is a reflector-type optical assembly. It includes a corrector plate at the front of the receiver, and a mirror at the center. In addition, an erector lens and an ocular or focusing eyepiece have been physically located between your eye and the phosphor button. The erector lens reverses the upside-down image on the button, for viewing purposes, while the ocular lens magnifies and focuses the image for you.

11-1606. Unfortunately, the phosphor button cannot continue indefinitely to discharge energy (emit light) without some form of re-



A
PHOSPHOR-BUTTON-TYPE INFRARED RECEIVER, CUT-AWAY DIAGRAM



B
ELECTRONIC-TYPE INFRARED RECEIVER, CUT-AWAY DIAGRAM

Figure 11-245. Typical Infrared Receivers charging or storing up of energy for future use. One type of charging method employed consists of exposing the phosphor button to emanations from a disk of gold foil impregnated with radium (Blitz charge).

11-1607. ELECTRONIC RECEIVER. The image-forming infrared receiver is similar in operational principle to the phosphor-button type. However, it uses an electronic image tube for a sensitive element, rather than the phosphor button, to transform invisible infrared radiation into a visible image. As shown in part B of figure 11-245, the electronic receiver consists of a reflector-type optical assembly and an electronic image tube. A high-voltage power supply is required for the tube.

11-1608. The reflector-type optical assembly focuses the image from the infrared source on the photocathode of the image tube. A powderized metal forms the photosensitive inner surface of the large end of the tube. This metal emits free electrons when subjected to either visible light or infrared radiation, and these electrons are attracted to the fluorescent screen by the high-voltage positive anodes inside the tube. The electrons are accelerated by higher and higher voltages as they progress down the tube toward the screen. Their high velocity causes the screen to glow on electron impact. The accelerating anode voltages, in conjunction with the proper deflection voltages, guide the electrons through paths which strike the screen in such a manner as to produce an inverted image. Considering that the image was inverted by the optical setup originally, the image is now right side up again. You can view the image through an optical magnifier.

11-1609. OPTICAL MATERIAL.

11-1610. LENSES AND WINDOWS. Common glass has very limited use in infrared optical equipments, because the transmission of both crown and flint glass in 3-millimeter-thick samples drops below 50 percent at wavelengths of approximately 2.6 microns, as shown in figure 11-246. However, if the glass is in the form of bubbles less than 1 millimeter thick, common glass has a useful transmission out to 5 or 6 microns. In fact, these have been used on photoconductive detector devices. At wavelengths up to 4 microns, special glasses have been developed. However, these glasses have proved very brittle and, therefore, difficult to work with.

11-1611. Sodium chloride (rock salt) is one of the most important materials used in infrared spectroscopy. It is not only economical, easily handled, strong, easy to polish, and readily available, but it is not soluble in common organic solvents. Unfortunately, it has one major drawback, in that it absorbs water. This defect can be partially corrected by using a plastic coating over the rock salt. If you use rock salt formed as a very thin window, it can be used in transmission media up to approximately 16 microns. However, in the 8 through 14-micron region, rock salt lenses are not practicable because of the high rate of dispersion in this frequency range.

11-1612. Silver chloride is a colorless, crystalline solid with the mechanical properties of lead. It is cubic in structure, and does not have a tendency to split (cleavage characteristic). It will not dissolve in water. Its main disadvantage is its photosensitivity, since it will darken and become useless when exposed to ultraviolet radiation, such as that from the sun. In fact, fluorescent lamps will deteriorate the crystal in only a few days. Therefore, silver chloride cannot be used in a practical instrument unless it is coated with some material to absorb ultraviolet light. It can be rolled into various shapes, but must be mounted in glass, plastic, or silver, because it will corrode most other materials very rapidly.

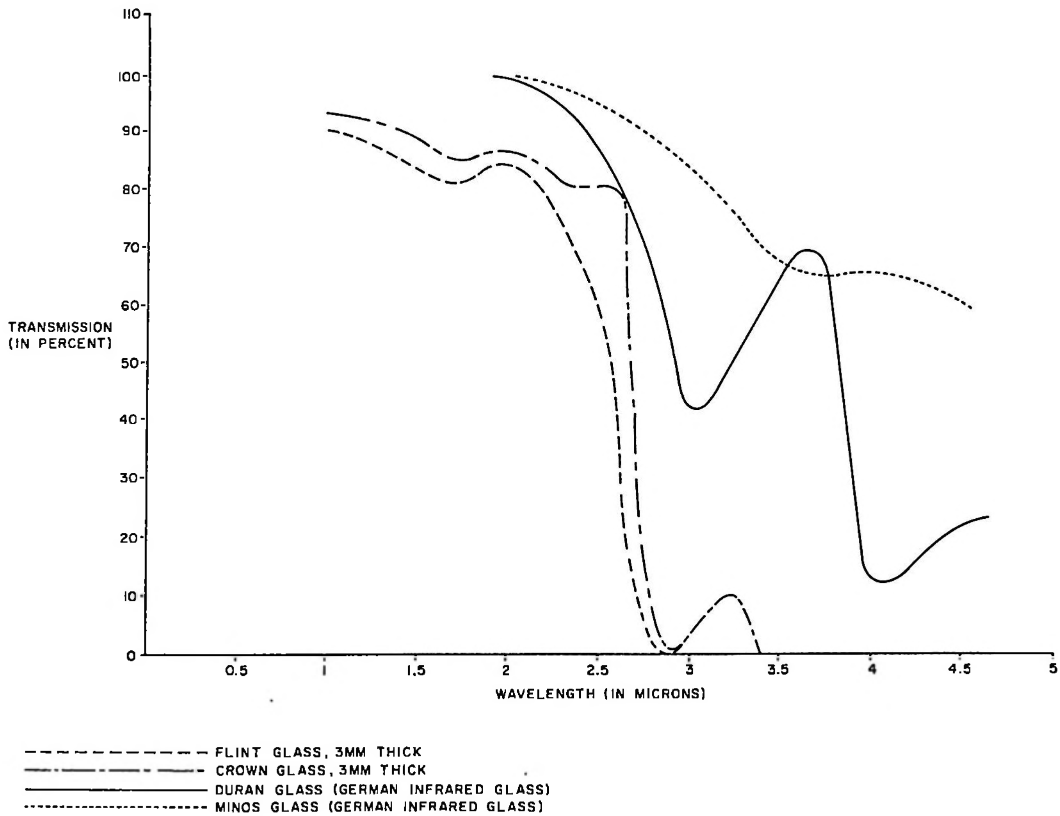


Figure 11-246. Transmission Characteristics of Glass

11-1613. KRS-5 material is a waxy-structure, red-orange cubic solid containing no tendency toward cleavage, and is harder than silver chloride. KRS-5, when used as a thin window, will transmit wavelengths beyond 40 microns. However, considering that the atmosphere is opaque between 25 and 40 microns, the increased transmission characteristic of KRS-5 is not a criterion. Therefore, since the optical properties of coated silver chloride (coated surface of silver sulfide) and KRS-5 are similar, then the criteria determining which material to use can be based on mechanical properties, availability, and cost. KRS-5 does not corrode, and, therefore, can be easily mounted. However, it cannot be fused to the mounting as can silver chloride. Next, KRS-5, unlike silver chloride, has a tendency to flow or lose its shape after a few weeks of mounting, or in even less time under pressure.

11-1614. The following fluorides are good window materials for 50-percent transmission at the listed micron range:

Strontium fluoride (10.7 microns)
Barium fluoride (11.7 microns)
Cadmium fluoride (9.8 microns)
Lead fluoride (11.1 microns)

These fluorides are practically insoluble in water, and lead fluoride is a good prism material.

11-1615. Cesium bromide is comparable to KRS-5 in transmission properties, it is highly soluble in water, but it has only half the tendency to flow or change shape with time, as does KRS-5.

11-1616. Cesium iodide is a good prism material, and has a good transmission characteristic of 80 percent out to 40 microns.

11-1617. Arsenic trisulfide (Servofrax) behaves like glass, it is comparatively hard, and it can withstand temperatures up to 200 degrees centigrade. It is not soluble in water, and has an excellent refraction index. Therefore, this material can be used for both windows and lenses.

11-1618. Germanium has a high reflection loss unless it is coated with selenium. When so coated its refractive index is approximately 4, and it can be used as a lens material.

11-1619. Silicon, in its pure form, has a useful transmission range out to about 22 microns, and, since its refractive index is about 3.5, it can be used as a lens.

11-1620. Selenium can be used as either a window or a lens. This material can be molded easily because it is soft and pliable at 50 degrees centigrade and becomes hard at 15 degrees centigrade.

11-1621. Sapphire is a hard, colorless material and is not soluble in water. It makes ideal windows out to 5.5 microns of transmission range. It can be used in either lead selenide or lead telluride photoconductive detectors, and as a supporting interference filter in the middle-infrared region.

11-1622. Periclase has approximately the same characteristics and properties as sapphire; it has a useful transmission range out to 9.5 microns. However, it has a higher melting point (2800 degrees centigrade).

11-1623. Diamond has a broad absorption band and a useful transmission range of between 6 and 24 microns. It has a high refractive index, but, if coated, can be used as window material.

11-1624. Plastics are useful transmitting materials if they are made very thin (less than 1 millimeter thick). Some plastics, prepared as a thin membrane, are useful out to 38 microns and can be used in the far-infrared region.

11-1625. OPTICAL COMPONENTS. The mirrors, prisms, diffraction gratings, and other infrared components are basically the same as those used for visible radiation.

11-1626. Lenses are normally selected for their index of refraction, transmission characteristics, and low dispersion factor to avoid different focal lengths at different wavelengths. Both glass and quartz are normally used in the near-infrared range (glass up to 2.5 microns, and quartz up to 4 microns). Germanium is good up to 6 microns, and arsenic trisulfide is satisfactory up to 8.5 microns. Beyond the near range, silver chloride and KRS-5 are used from 8 to 14 microns; silicon is normally used to 22 microns, and cesium bromide is used up to 35 microns.

11-1627. Prisms used in spectroscopes are selected for their transmission, dispersion, and resolution characteristics over the required range. The dispersion factor is normally a function of both wavelength and temperature. Lithium fluoride, potassium bromide, sodium chloride, calcium fluoride, and sodium fluoride are materials that are used for constructing prisms for infrared applications.

11-1628. Window material is selected on the basis of transmission characteristics, solubility in water, melting point, coefficient of expansion, and low dispersion factor. Therefore, many materials which are useless as prisms can be used as windows. Some of the materials which may be used are chlorinated rubber, Pliofilm, lacquer, silver chloride, etc.

11-1629. Mirror material usually consists of glass, coated on the reflective side with a layer of silver, aluminum, antimony, or gold. This layer is then coated with silicon monoxide as protection from both temperature and moisture. This type of mirror has a constant focal plane for all wavelengths, but the reflectivity may vary with wavelength. Mirrors are used for wide-band applications because refracting optical devices have a refraction index which varies with wavelength, and is therefore not satisfactory.

11-1630. INFRARED FILTERS. Filters are primarily used to isolate a particular part of the infrared spectrum, and are classified by their spectral characteristics. A band-pass filter will pass a particular band of wavelengths, and may be either a broad or narrow band-pass filter. A wave-pass filter will pass only wavelengths longer than or shorter than a predetermined wavelength. A long-wave-pass filter is commonly referred to as a low-pass filter; conversely, a short-wave-pass filter is referred to as a high-pass filter, because a shorter wave represents a higher transmission frequency. In all cases, a band-pass filter must have a sharp cutoff on either side of its pass band. Actual filtering occurs because of one or more of the following phenomena:

a. SELECTIVE ABSORPTION FILTERS. All materials will absorb radiation selectively. Solids, such as glass and KRS-5, do not have a cutoff selectivity sharp enough to provide good filtering. Dye materials are used primarily as high-pass filters which remove or cut off visible light; they are composed primarily of plastics containing dye, dye-plastic coated glass, dyed cellophane, nylon, polyvinyl, or alcohol films. Semiconductors are also used as high-pass filters to remove visible and near-infrared radiation. A thin slice of a single silicon or germanium crystal, or lead sulphide, lead selenide, lead telluride, bismuth, antimony, or magnesium-oxide films all make good high-pass filters. In fact, even black paper can be used to cut off the low-frequency region.

b. SELECTIVE REFRACTION FILTERS. The refractive index of a lens changes with wavelength, and, consequently, can be used as a filter to remove infrared radiations above or below the desired wavelength.

c. SELECTIVE REFLECTING FILTERS. Many crystalline materials reflect different wavelength selectively. For example, quartz reflects about 75 percent at 9 microns and only 3 percent at less than 8 microns.

d. SCATTERING FILTERS. If you mix small particles of one type of material throughout another material, the transmission effect of the over-all result may be entirely different from that of either material. The transmission characteristics depend on the distributed particle size, the layer thickness and the refractive index of both materials. Maximum transmission will occur when the particle diameter is 1/2 the radiation wavelength. When particles are suspended in a transparent medium, the highest transmission will occur at that wavelength where the refractive indices of both materials coincide. At other wavelengths, the particles embedded in the material scatter the radiation.

e. INTERFERENCE FILTERS. Actually, the interference phenomena existing between two plates is used to make either a reflection- or a transmission-type infrared filter. The simple reflective filter is made by using a dielectric with a semireflective metal. The transmission type could be a thin film of dielectric material coated on both surfaces with a semireflecting metal layer. A reflection type of interference filter is illustrated in figure 11-247, and a transmission type is illustrated in figure 11-248.

f. POLARIZATION FILTERS. Polarization filters are not widely used in infrared applications. This type of filter consists of two polarizers and a birefringent component. The radiation passing through the first polarizer is polarized in one plane, and then passed through the birefringent component; here it is translated into two separate images, each of which is polarized at right angles to the other. These two images are combined as they pass through the second polarizer. The resultant output depends on the phase difference between the two images.

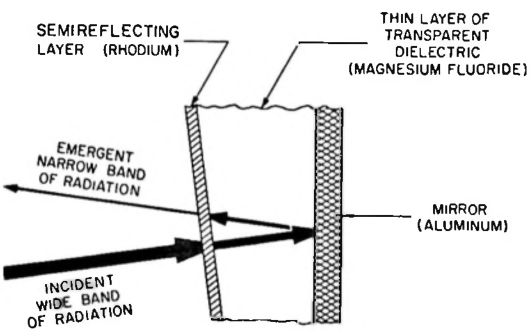


Figure 11-247. Simple Reflection-Type Interference Filter

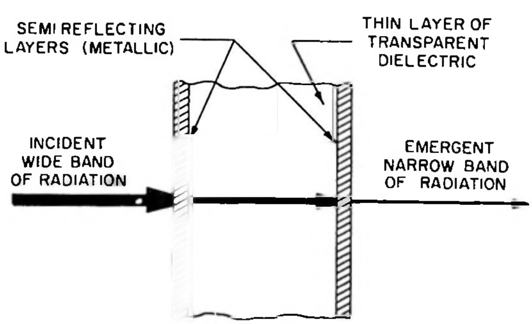


Figure 11-248. Simple Transmission-Type Interference Filter

11-1631. DIFFRACTION GRATINGS. All diffraction gratings are used for spreading the radiation beam into a spectrum. These gratings consist of glass or metal plates containing either ruled lines or grooves, and are classified into two types: transmission type (made of glass) or reflection type (made of metal).

11-1632. The transmission grating generally consists of a transparent plate or film with thousands of lines etched on one surface. The actual lines are opaque to radiation, but the spaces between the lines (slits) transmit radiation. A second type of transmission grating, called echelon grating, consists of many transparent layers arranged in steps so that one part of the radiated beam is retarded by one layer, a second part by two layers, etc. Unfortunately, echelon gratings contain undesirably small separations between successive spectra.

11-1633. The reflection grating consists of a metal plate or film with parallel lines scratched on one surface. The actual lines reflect radiation diffusely (they look relatively black), whereas the spaces between the lines reflect radiation in a normal manner. A second type of reflection grating is the echelette grating. This type of grating is composed of gold-plated copper sheets with grooves in one surface. One side of the groove is vertical, while the other side is tilted at an angle to the gold surface. The echelette grating effectively removes short-wavelength (high-frequency) radiation. Therefore, the grating can be used as a low-pass filter. Good resolution is obtained from this type of grating, and it is generally used with a less sensitive detector than other types of gratings, to detect passive infrared sources.

11-1634. INFRARED DOMES. The infrared dome (irdome) is comparable to the regular well known radar dome (radome) in that its purpose is to shelter infrared equipment. On an aircraft it must not provide over a nominal amount of airframe drag. In any location, it must be transparent or refractive to the infrared radiation from, or to, the equipment being used. The dome must have good optical properties and resistance to wind abrasion; it must also be able to withstand severe temperature changes, and, if used as airborne equipment, it must have good aerodynamic characteristics. Wind resistance on the dome can create heat from friction, and thus permit the equipment detector to detect the irdome. Quartz, with its high resolution, or Kel-F, because it is not brittle, may be used for the dome material. In fact, as shown by table 11-11, many materials may be used for the dome.

11-1635. OPTICAL ASSEMBLY.

11-1636. The infrared optical requirements are similar to those in a camera or telescope, because infrared equipment will collect and focus on a target emitting or reflecting infrared radiations. However, instead of recording it on photographic film or the naked eye, the infrared optical facility focuses the radiation on an infrared detecting cell. The infrared facility must be symmetrical about its central axis and provide enough space for the detecting cell, the cooling equipment, a reference black body, and a chopper. In addition, the tactical requirements of a good optical facility demand long range, a fast scanning rate, a large field of view, high resolution, and the

Table 11-11. Dome Materials for the 3- to 5-Micron Infrared Region

Property	Quartz	Arsenic Trisul-fide	Calcium Aluminate	Synthetic Sapphire	Periclase (MgO)	Calcium Fluoride	Silicon	Germanium
Transmission	Poor	Good	Fair	Good	Good	Good	Good	Good
Low index (low inherent reflective loss)	Good	Good	Good	Good	Good	Good	Poor	Poor
Thermal properties	Good	Poor		Fair	Fair	Fair	Good	Good
Mechanical properties	Good	Poor	Good	Fair	Fair	Good	Poor	Poor
Chemical properties	Good	Fair	Good	Good	Poor	Poor	Good	Good
Availability	Good	Good	Fair	Fair	Poor	Good	Poor	Poor

smallest usable detector for proper performance.

11-1637. EFFICIENCY. The refraction surfaces must be coated with a reducing film, and each reflecting surface with an absorption reducing film, for the most efficient equipment performance. In addition, the refracting elements must be composed of materials having low absorption characteristics.

11-1638. FOCAL LENGTH AND APERTURE DIAMETER. The higher you make the signal-to-noise ratio, the better will be the infrared facility. The signal-to-noise ratio of a photoconductive detector is inversely proportional to the square root of the detector area, and directly proportional to the incident radiant power. Therefore, the detector area should be as small as possible, and the power should be large. This presupposes a short focal length coupled with a large aperture. The focal length and aperture size are extremely important, because the received infrared image deteriorates rather rapidly as the focal length increases or the aperture decreases.

11-1639. ABERRATIONS. The refractive elements (lenses) of your facility are subject to chromatic aberration, which is a result of the changing refractive index of a material with wavelength. This aberration is not related to either aperture diameter or focal length, but is related directly to wavelength. Other aberrations occur with all monochromatic (one-wavelength) radiation, and are tabulated in the following list:

a. Spherical aberration is proportional to the cube of the aperture-to-focal length ratio.

b. Coma aberration is proportional to the square of the aperture-to-focal length ratio, and to the field angle.

c. Astigmatism (on-axis) and field curvature aberration are directly proportional to the aperture-to-focal length ratio, and to the square of the field angle.

d. Astigmatism (off-axis) is proportional to the cube of the field angle.

11-1640. OPTICAL RANGE. The optical range of the equipment you are using will be improved by either increasing the aperture diameter or decreasing the focal length. The range is proportional to the square root of the aperture diameter times the optical efficiency, all divided by the focal length-to-aperture ratio.

11-1641. DETECTORS.

11-1642. GENERAL. The detector is the fundamental sensing device of infrared equipment and, as such, plays a critical role in infrared equipment operation. The detector generally functions as a transducer and converts radiant energy into an electrical signal. You can use this electrical signal in other communications-electronics devices, or you can use it for observation purposes.

11-1643. Detectors may be separated into two primary classes, depending on their working principle of operation. The two basic types are <u>thermal detectors</u> and <u>photo detectors.</u> Thermal detectors are not wavelength-sensitive, but respond to radiation throughout the infrared spectrum. When you think in terms of image-forming detectors, you are thinking in terms of photo detectors, which vary with the frequency or wavelength of the incident radiation.

11-1644. The terminology of infrared detectors which may be new to you is listed below for your ready reference:

a. SENSITIVITY - This term is a measure of the minimum detectable part of the quantity of infrared radiation.

b. RESPONSE - This term is a quantitative expression of output versus input.

c. SPECTRAL SENSITIVITY - This term refers to the operating range of the detector, and is measured in terms of the band of the infrared spectrum to which the detector will respond. It is measured in microns, and is expressed in terms of wavelength.

d. MINIMUM DETECTABLE POWER - The limit of usefulness of a detector is the minimum amount of radiant power, expressed in watts, to which it will respond. This minimum is that point where the detector output voltage is no greater than the internal noise voltage of the detector. Actually, the minimum detectable power is the reciprocal of the sensitivity of a given detector.

e. TIME CONSTANT - This is a measure of the time required for the detector to respond to a given radiation.

11-1645. LIMITATIONS. The limitations imposed on infrared detectors are severe in that the function of a detector is critical. The ideal detector should have a broad spectral response, a short response time, low minimum detectable power, a high output/input ratio, and the ability to discriminate between targets and background noise. Unfortunately, present-day detectors cannot cope with all of these requirements. Thermal detectors have a broad spectral response, but they have a long time constant and low sensitivity. Photo detectors have a short time constant and are approximately a thousand times more sensitive than thermal detectors, but their spectral response is limited to the near-infrared region.

11-1646. Noise constitutes one of the greatest limitations to both detector sensitivity and over-all detector performance, because the noise component restricts the interpretation of target data from a given output signal. Noise is the result of the random time of target emission, background noise from the atmosphere, Johnson noise or current noise within the detector, and various other minor background noises. Johnson noise is a type of thermal shot effect caused by internal temperature fluctuations within the sensitive element. Johnson noise can be reduced by lowering the operating temperature of the detector and by using as high a chopping frequency as possible. Current noise is most prevalent in photoconductive detectors, and is the result of internal resistance fluctuations caused by random contacts between the semiconductor microcrystals and the thermally generated current carriers. Cooling the detector will reduce this type of noise.

11-1647. When discussing noise level and detector sensitivity, the term noise equivalent power (nep) is used. Actually, this factor is comparable to the minimum detectable power, discussed previously. The nep is the infrared power, expressed in watts, which must be received by the detector to generate a signal equal to its rms noise level over a given bandwidth.

11-1648. Infrared detector characteristics are listed in table 11-12.

11-1649. THERMAL DETECTORS. Any temperature change, caused by incident radiation, can be made to produce the following effects in the detector:

a. Generation of an electromotive force (thermocouple).

b. Resistance change (bolometer).

c. Gas volume change (Golay cell pneumatic detector).

Table 11-12. Infrared Detector Characteristics

DETECTOR	TIME CONSTANT (SECONDS)	SPECTRAL SENSITIVITY (MICRONS)	FORM OF OUTPUT
Bolometer	4×10^{-3}	0.1-16	Change in resistance
Thermistor bolometer	3×10^{-3}	1-15	Change in resistance
Cesium	1×10^{-6}	0.1-1.3	External photoeffect
Evaporagraph		0.1-12	Latent image
Germanium	1×10^{-3}	0.3-1.7	
Golay cell	3×10^{-3}	0.5-15	Pneumatic
Lead selenide	1×10^{-6}	0.1-4.5 0.1-6 (cooled)	Internal photoeffect
Lead sulfide	$1\text{-}4 \times 10^{-6}$	0.7-3.0 0.5-4 (cooled)	Internal photoeffect
Lead telluride	1×10^{-6}	1-5.5 (cooled)	Internal photoeffect
Photographic, type 2		0.1-1.2	Latent image
Thermocouple	50×10^{-3}	5-15	Electromotive force
Thermopile	$10\text{-}50 \times 10^{-3}$		Electromotive force

11-1650. The thermal detector is used primarily because of its wide spectral response. It depends solely on the heating effect of the infrared radiation rather than on its wavelength. The choice of which one of the several types of thermal detectors should be used for a particular application is chiefly dependent on the chopping frequency requirements. The thermocouple detector operates best at low chopping frequencies, while the bolometer detector is used at the higher frequencies (over 30 cps). The basic concepts, principles of operation, and illustrative requirements of both thermocouples and bolometers have been extensively covered elsewhere within this publication, and will not be repeated at this point.

11-1651. PNEUMATIC DETECTORS. The pneumatic, or Golay, cell consists of a hermetically sealed cell filled with air or other gas which does not absorb radiation easily, a transmitting window, and an absorbing membrane located in the center of the gas-filled cell. The optical photoelectric components consist of a glass with 100 ruled lines per inch, a meniscus lens, condensing lenses, a photocell-exciting lamp, and a photocell. A typical Golay cell arrangement is illustrated in figure 11-249. The Golay cell detector is actually a type of thermal detector in that its output depends on the heat radiation. However, rather than undergoing a resistive change within the sensitive element, the Golay cell depends on a volume of gas change which is proportional to temperature change. As the Golay cell uses an all-metal film as a radiation absorber, it is not selective over the entire infrared spectrum. The change in indication is the result of a photoelectric circuit which is affected by the physical displacement of one surface of the gas container. This moving surface is the criterion used to determine the limits of sensitivity.

11-1652. The actual operation of the pneumatic, or Golay, cell is simple; the radiation entering the window heats the absorbing membrane, which heats the surrounding gas,

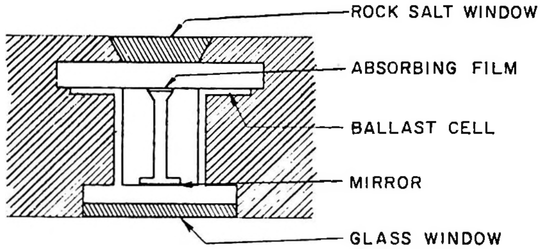


Figure 11-249. Golay Cell Structural Arrangement

which, in turn, distorts the diaphragm. The diaphragm reflects light to the photocell in an amount depending on the diaphragm distortion.

11-1653. The spectrophone is another type of pneumatic detector which operates on the gas-expansion principle. However, sound, rather than an electrical photocell signal to an indicating device, is produced for measurement with a microphone hookup.

11-1654. PHOTO DETECTOR. The photo detector is sensitive to wavelength change rather than heat. It is a selective device with low-frequency cutoff. Typical photo detector devices, which change electrically when exposed to infrared radiation, are photo-emissive, -conductive, or -voltaic. Actually, detectors which depend on luminescent properties are also considered as photoelectric devices, as are photographic photo detectors.

11-1655. In photoemissive type of detector, electrons are emitted from the surface of some substance when you expose that substance to infrared radiation. The photoemissive cell is chiefly composed of a cathode which emits electrons when exposed to infrared radiation, and an anode which receives these electrons; the resultant plate current is sent to an amplifier, as shown in figure 11-250. The fundamental relationship governing the operation is explained by the following photoelectric equation:

$$E = hv - e\phi$$

where:

E = kinetic energy
h = Planck's constant
v = radiation frequency
$e\phi$ = energy required to remove one electron from the surface of an exposed material (work function).

11-1656. The spectral sensitivity of photoemissive detectors extends from the red end of the visible spectrum into the near-infrared range. Therefore, beyond 1.2 microns, the quantum energy level of the radiation is not sufficiently high to cause photoemission. One limitation to detector sensitivity is dark current. Dark current is a current which flows in either the photoemissive or photoconductive detector during total darkness (no applied radi-

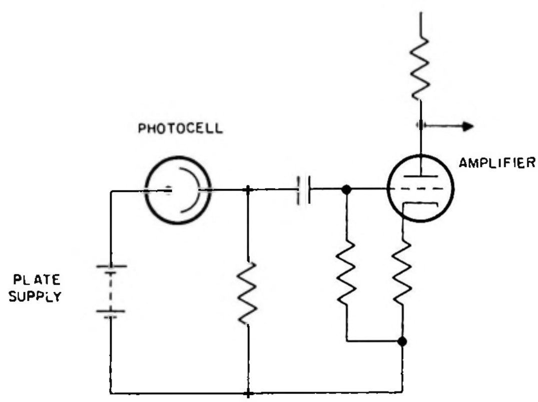


Figure 11-250. Photoemissive Cell, Simplified Schematic Diagram

ant energy), and consists primarily of leakage current and thermionic emission. Since these effects are primarily due to temperature rise, operating the detector at a reduced temperature will tend to reduce both leakage currents and thermionic emission currents.

11-1657. The construction of the photoemissive detector is rather simple; it generally consists of only a cathode of silver-oxygen or antimony-cesium and an anode (plate) both in a vacuum cell. The vacuum cell suffers less from fatigue than does a gas-filled cell. The gas-filled cell permits a multiplication of up to 100 times the original photocurrent because the electrons emitted by the cathode ionize the gas molecules. However, as was mentioned above, the gas-filled cell is fatigued easily, and, it is unstable for low-level measurements; in addition, the presence of positive ions adversely effects the frequency response, which permits a sharp chopping frequency cutoff for frequencies beyond 1000 cps.

11-1658. The photomultiplier type of detector has the advantage of internal radiation current multiplication and high sensitivity to radiation. The current multiplication occurs because electrons emitted from the cathode strike a surface that releases secondary electrons, which, in turn, strike another surface that releases electrons. The electrons continue to build up until they strike the final anode (plate). However, the photomultiplier detector is rarely used for infrared detection.

11-1659. The photoconductive detector obtains the photoelectric effect by virtue of the fact that some materials increase their electrical conductivity as a result of infrared illumination. This effect is not due to heat; it is the result of electron absorption in the conductive bands when exposed to a sufficient quanta of radiation. Since the photoconductor has a definite long-wave limit (it is wavelength-sensitive), only quanta of minimum magnitudes will activate the photoconductor.

11-1660. Semiconducting materials such as lead salts, silicon, and germanium are used in photoconductive detectors, because these materials will reflect or absorb high-energy light rays while they transmit low-energy infrared rays. These materials are normally gold-doped to instill the desired properties. The photoconductive effect is due to the fact that two kinds of charged carriers are contained in the semiconductor: negative electrons and positive holes. When the semiconductive material has been adequately charged, electrons in the valence band will shift over to the conduction band, illustrated in figure 11-251. As the electrons in the conduction band circulate more freely, more current will result. Photoconductive detectors are constructed by either evaporating the semiconducting material on a glass slide or placing it there by chemical processes and vacuum-sealing the glass between two electrodes. The material is then sensitized by introducing oxygen at a high temperature. The combination of elements within the cell will determine its physical size. Normally, photoconductive detectors are operated without special cooling. However, special applications may require that the cell be artifically cooled. The coolant (liquid helium hydrogen) is injected, between the specially constructed glass walls of the cell, by capillary action. The coolant will boil at reduced pressure in its vacuum enclosure, and thus provide adequate cooling.

11-1661. The sensitivity of the photoconductive detector is many times greater than that of a thermal detector. The spectral sensitivity of a photoconductor is normally around 9 microns without artificial cooling, and up to 120 microns (reduced sensitivity) with artificial cooling. Both the temperature and the illumination determine the time con-

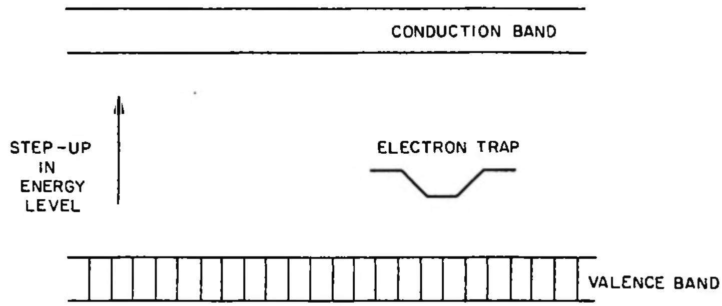


Figure 11-251. Structure of a Semiconductor

stant function. The time constant is not as fast as for photoemissive detectors, but is faster than for thermal detectors. This permits high chopping frequencies (1000 cps) without loss in sensitivity.

11-1662. PHOTOVOLTAIC DETECTORS. The photovoltaic and thermocouple detectors are similar in that both depend on irradiation at the junction of dissimilar metals. However, the photovoltaic detector depends on the photoelectric properties of the material rather than the thermoelectric properties. Three types of photovoltaic cells are mosaic, thallium, and lead sulfide. Mosaic cells are too critical for normal use, and the lead sulfide is not sensitive enough for normal use. The thallium sulfide cells are, therefore, the only cells of interest. These are produced by depositing thallium on a metal base, converting it to sulfide, and covering it with a gold or platinum layer.

11-1663. LUMINESCENT DETECTORS. Some substances become luminescent when stimulated by infrared radiation. The emission of light is not caused by heat or incandescence, because it occurs at low temperatures. Phosphorus is the normal substance used by the Air Force as the luminescent material, because it stores energy when exposed to visible light, and later emits a bright visible light when illuminated by infrared. The Air Force use is normally limited to detecting (receiving) infrared transmissions at night. The phosphorus is exposed to sunlight for energy storage, and is then used as a detector after dark.

11-1664. IMAGE-FORMING DETECTORS. Actually, the image-forming detector is a type of photo detector. It is treated separately because it provides a pictorial display. The image-forming detector requires optical devices to focus the image, but does not require chopping devices as does the non-image-forming detector. The image-forming detector is not very sensitive, and its performance is low. The target must be illuminated with infrared radiation; therefore, because of security considerations, use of the image-forming detector by the Air Force is limited. As shown in figure 11-252, the photoemitter type of image-forming tube employs a photosensitive cathode element which emits electrons when exposed to infrared radiation. These electrons are attracted to an anode (fluorescent screen), to provide a fluorescent display having a point-for-point light-to-radiation correlation. The picture is actually formed by modulating the electron beam sent from the cathode to the anode; the signal can be used in standard video circuits.

11-1665. Many image tubes have been de-

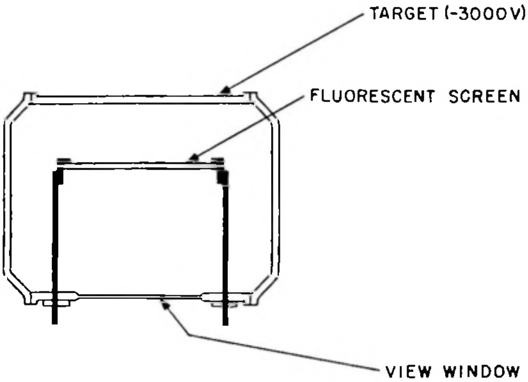


Figure 11-252. Electron Image Converter

veloped. However, the basic types are illustrated in figure 11-253.

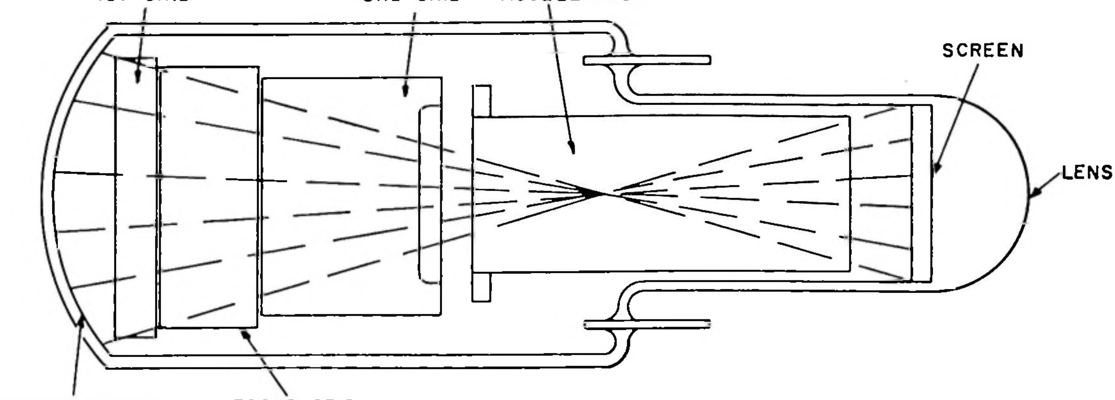
11-1666. TARGETS AND BACKGROUNDS. The target is considered to be whatever object you seek with your infrared receiving or detecting equipments. The background is considered to be the remainder of the picture, which contains the desired object. Actually, the problem is relative, in that the background of one image may be considered as the target under other circumstances.

11-1667. The discrimination between target and background can be difficult; this is especially true with non-image-forming detectors, which reveal only the presence, direction, and intensity of the infrared radiation. With image-forming detectors, the image formed corresponds to the thermal rather than the visual configuration of the target. For example, a truck image would consist of the heated parts of the truck, such as the motor and exhaust pipe, rather than the entire truck.

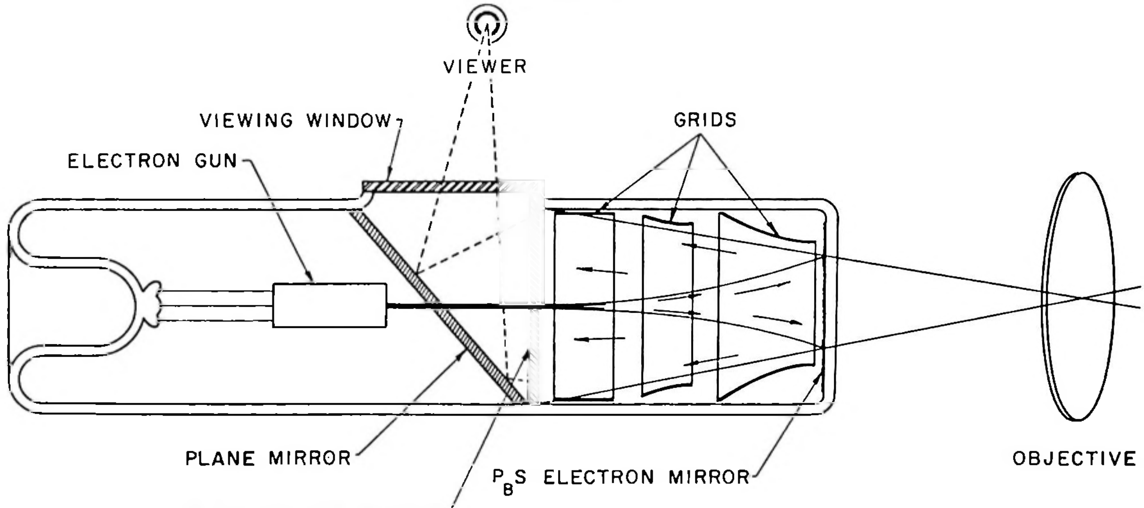
11-1668. Targets may be active (generate infrared energy) or passive (reflect or redirect infrared energy obtained from the sun). An active target can be picked up by a passive infrared device, but a passive target may need an active infrared device to illuminate the target. Obviously, the more heat generated or reflected from a target, the greater the detected energy. Actually, a target such as an airfield may be the background for an aircraft target under other circumstances.

11-1669. Backgrounds are separated into day and night types. A daytime background emits long wavelengths in accordance with the temperatures and emissivities of the objects constituting the background; it also emits short wavelengths because of reflected sunlight. The darker the background, the greater the sunlight heat retention, and, consequently, the more the infrared radiation generated. A nighttime background is due to night sky airglow, the aurora, starlight, etc. Therefore, the background may be difficult to distinguish from the radiated energy from the stars. Actually, if the target is hotter than its background, you can use special optical filters to distinguish one color from another, or the target from its background.

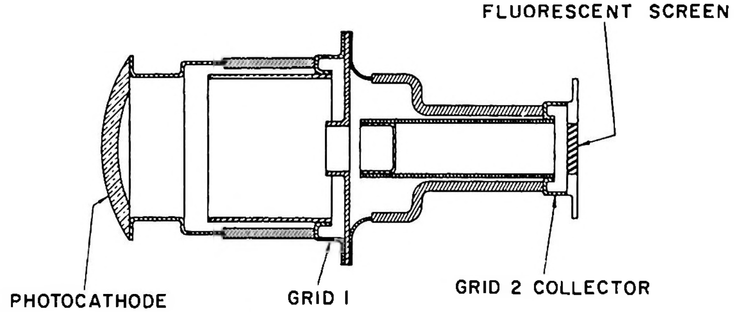
11-1670. A detector may use spectral sensitivity to detect only wavelengths of the specific target, and ignore the longer wavelengths received from the background. Filters may block out wavelengths longer than those emanating from the target, thus effectively providing contrast. Ground targets are primarily distinguished from their backgrounds by the heat differences between the two. Unfortunately, some objects change in temperature with respect to time, and produce "washout" (target not distinguishable from background). For example, when steel and concrete lie close together on a sunlit day, the concrete may be warmer than the steel because of its retained heat from the previous day. However, as the day wears on, the steel heats faster than the concrete;



A
1P25 IMAGE TUBE



B
ELECTRON MIRROR IMAGE TUBE



C
6032 IMAGE TUBE

Figure 11-253. Typical Infrared Image Tubes therefore, at some point in time, both the steel and the concrete may reach the same temperature, and you cannot distinguish between the target and the background. As the day becomes hotter, the steel gradually becomes warmer than the concrete, and you have a target. When evening draws near, the steel cools off more quickly than the concrete, and when they again match in temperature a washout is created. Then the steel becomes cooler and distinguishable from the warm concrete. This effect is illustrated in figure 11-254.

11-1671. INFRARED TRANSMISSION THROUGH ATMOSPHERE. As the atmosphere is the standard medium between the target and the detecting equipment, you should understand its absorption, scattering, and transmission properties. Atmospheric moisture absorbs and scatters infrared rays, and this effect limits the transmission distance of radiation. Water vapor in the air is the primary limitation to transmission or reception distance.

11-1672. INFRARED COUNTERMEASURES. Design and tactics are the two prime infrared countermeasures. Design countermeasures affect the inherent radiation pattern of infrared targets, and tactical countermeasures include the defensive movement of the

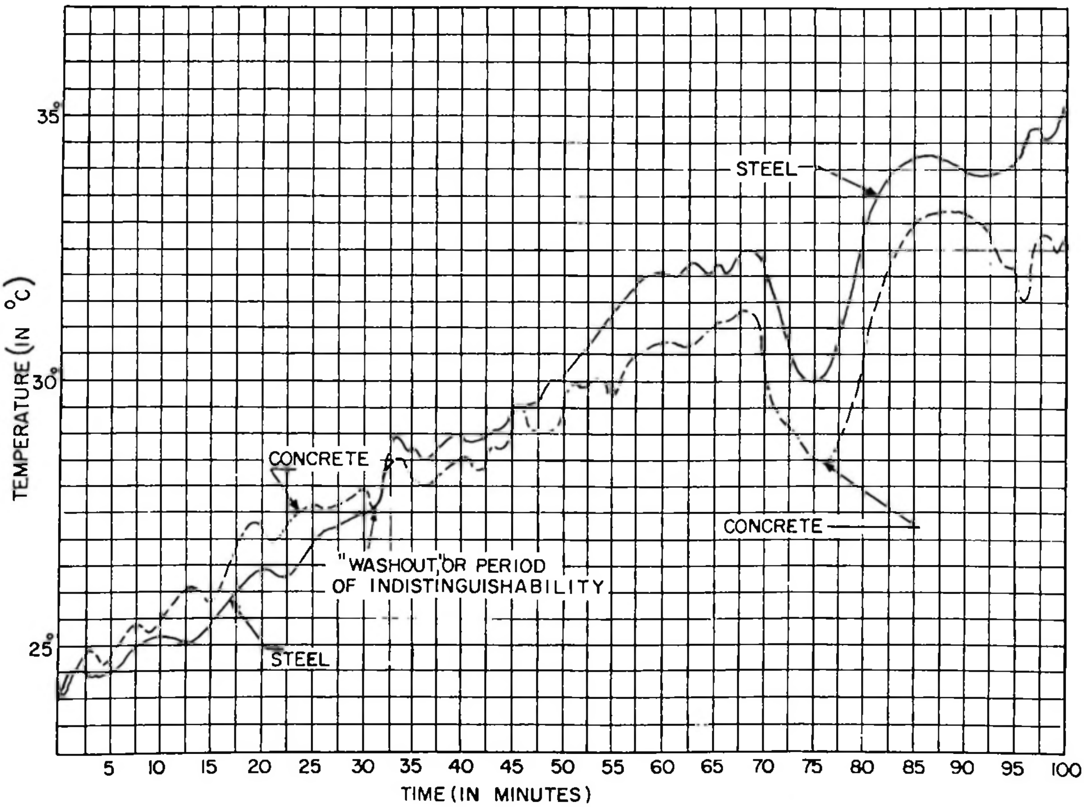


Figure 11-254. Temperature Difference Between Target and Background (Steel and Concrete)

target, uses of screen or decoys, etc. You can design a radiation shield to decrease radiation by decreasing the operational temperature or by using cooling techniques to dissipate the resultant operating equipment heat. In the case of aircraft, you can design a method of mixing chemicals with the jet stream to scatter or absorb the ejected heat gases; you can also eject flares or balls of fire (heated spheres of carbon), or tow a decoy behind the actual aircraft.

11-1673. INFRARED RECEIVERS. Block diagrams of two typical infrared receivers are shown in figure 11-255. These receivers consist of a reflective or refractive optical equipment, an infrared detector cell, an amplifier circuit, and a decoding circuit. The picture received by the optical equipment is focused on a reticle, which modulates the incident energy at a frequency suitable for application to the detector, encodes the signal with target directional information, and discriminates between target and background. The detector receives the picture energy and passes it on to a decoder, which amplifies and decodes the resultant image.

11-1674. AMPLIFIERS.

11-1675. GENERAL. Amplifiers are designed for use with particular types of detectors to prevent incompatible amplifier

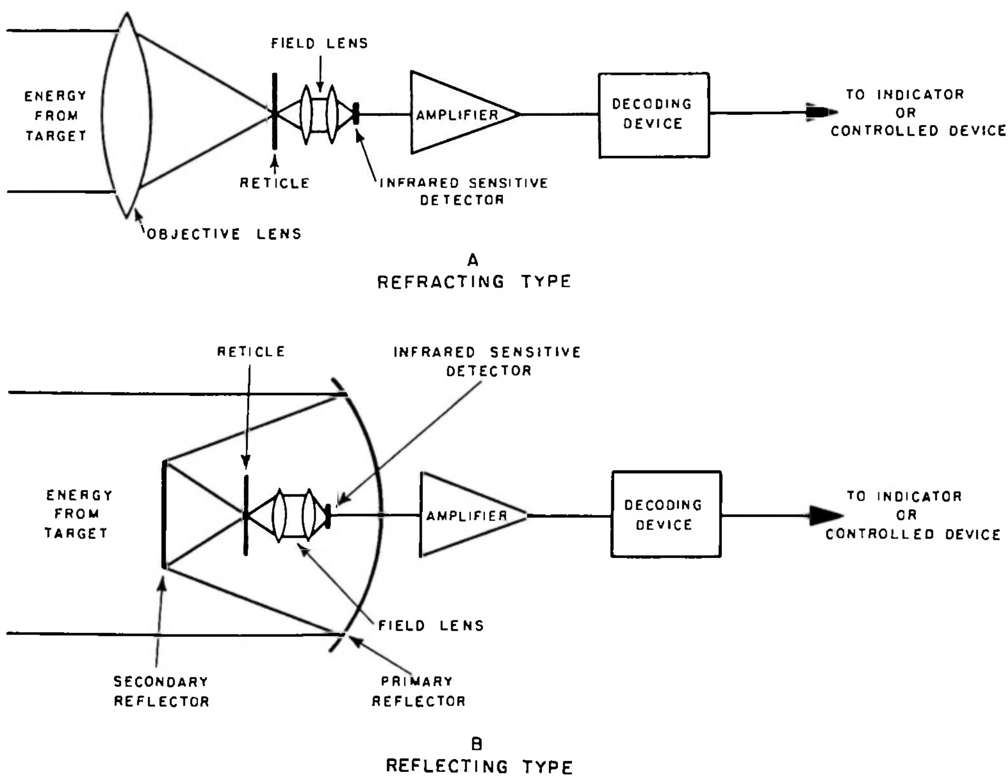


Figure 11-255. Typical Infrared Receivers, Block Diagrams

design from lowering the effectiveness of the detectors. The amplifier must have a lower internal noise level than the detector and a constant gain level, and the chopping frequency should be such that you have the longest possible interval of observation. The primary types of amplifiers are the optical lever, the dc or ac amplifier, and the photomultiplier.

11-1676. GALVANOMETER AMPLIFIER. The galvanometer amplifier functions as an optical lever by permitting a beam of light that is reflected from the galvanometer mirror, to illuminate two photoelectric cells. The output from these cells is sent to another galvanometer or to a current-sensitive device. The galvanometer amplifier gain depends on the light intensity from the first galvanometer mirror, the photoelectric cell sensitivity, and the physical size of the optical lever. This amplifier setup has a normal time of 1 second and a gain of approximately 1 million. Therefore, this type of amplifier cannot be used for a short observation application. Normally, this amplifier is used in conjunction with either a thermocouple or a metal bolometer, and the resistance of the detector matches that of the galvanometer, for maximum sensitivity.

11-1677. DC AMPLIFIERS. These amplifiers are not vibration sensitive as are the galvanometer amplifiers. However, gain stability and drift rate are difficult to maintain. This situation is corrected by interrupting the dc input, amplifying this current in an ac amplifier, and then rectifying the output.

11-1678. AC AMPLIFIERS. The ac amplifier is used in conjunction with low-impedance detectors and thermocouples for low frequencies, with photoconductive detectors, and with superconducting bolometer containing a low and critical impedance. For amplification of low-frequency signals produced by low-impedance detectors, a narrow-band amplifier must be used. Narrow bandwidths can be achieved by electrical networks that have no feedback at the resonant frequency, but high negative feedback at frequencies above or below resonance. One ac amplifier used with photoconductive cells is the homodyne. In this amplifier the output of the detector photoconductive cell is applied to a detector amplifier. The output of a reference radiation source is applied to a second photoconductive cell and reference amplifier. The detector amplifier and reference amplifier are tuned to the same frequency, and the outputs are applied to a comparison rectifier circuit. The output of the comparison circuit is applied to a meter which indicates the strength of the detected radiation. Transformers can be used to match the low impedance of a superconducting-bolometer detector, which may be a fraction of an ohm, to the input impedance of an ac amplifier. With the transformer operated at a low temperature, it is possible to obtain measurements with a sensitivity limited primarily by the inherent noise of the detector.

11-1679. PHOTOMULTIPLIER AMPLIFIER. The cathode of a photomultiplier tube can be used as a photoconductive detector. Electrons emitted from the photocathode are directed by a high-voltage field to an electrode where secondary emission, greater than the initial emission, occurs. The electrons emitted from this electrode are then directed by a high-voltage field to another electrode, where secondary emission is again produced. A photomultiplier tube may have several such electrodes, and the current amplification may be on the order of several hundred thousand times.

11-1680. INFRARED TRANSMITTERS.

11-1681. GENERAL.

11-1682. The infrared transmitter is composed primarily of an energy source and a visible light filter. The radiation source transmits over a wide frequency band, including visible light, and the filter removes the visible light for security reasons. The source could be a tungsten lamp if its response speed were not so slow as to restrict Morse code operations to less than 8 words per minute. However, a high-speed mechanical shutter will increase the signaling speed, and the development of rugged, shockproof tungsten lamps was initiated to eliminate both shock and vibration effects. For some applications, where the source energy must be modulated, a slotted cylinder is rotated to create a chopping or modulation effect. Cesium lamps have been developed for use instead of tungsten lamps because they have less distortion, a higher modulation rate, increased power output (5 times that of tungsten), and narrow-band selectivity.

11-1683. MODULATION.

11-1684. Modulation may be accomplished by passing the radiation through a filament lamp or gas discharge source, at the radiation source. Modulation may also be accomplished with the aid of an electro-optical device (Kerr cell) or some form of mechanical modulation of the outgoing beam. Vibrating mirrors or any beam chopping device may be employed at the output for mechanical modulation purposes. In rf facilities an electrical modulation method using supersonic vibration of an optical element is employed.

11-1685. Electrical modulation is the most efficient method of modulation. However, the mechanical method of modulation, while not as efficient, is lighter in weight. The modulation method to be employed will determine the size, weight, power, efficiency, and security of the radiation source. In addition, it may partially affect the design of the receiving equipment.

11-1686. MEASUREMENTS AND TESTS.

11-1687. CALIBRATION AND ALIGNMENT. Following the repair of infrared equipment, calibration and/or alignment may be necessary. On smaller and less complex equipments, you should perform the alignment procedures by following the instructions contained in the applicable equipment technical manual. You can check the performance standard for equipment operation by calibrating the sensitivity of the amplifying equipment. Replacement of components or units within the electronic portion of an infrared equipment may result in the necessity of readjustment or complete realignment procedures. Random adjustments usually require complete realignment after you have located the trouble. Do not adjust any equipment until you determine that a faulty component is not causing the trouble.

11-1688. TESTS AND EVALUATIONS. Periodic tests or evaluations should be performed during the introductory period of an infrared facility. In single-service testing, a specialized testing agency may perform certain tests for all Air Force services. These tests should provide engineering and service test data. The report on these tests, supplied to all the using services, should leave nothing to the intuition or imagination of the user. They should be complete and comprehensive, supplying the following checklist of data:

a. Complete performance specifications.

b. Technical description and accuracy of the equipment used.

c. Precise details of the layout and necessary procedure to conduct every test necessary to satisfy the performance specification requirements.

d. Complete test data (ie, actual values observed and recorded).

e. Summary.

f. Conclusions.

g. Recommendations.

11-1689. INSPECTION. All repairable infrared equipment should be inspected and screened to determine both the level and extent of the required maintenance. A bench check is necessary to determine the actual damage. Following the installation or repair and reinstallation of infrared equipment, a complete inspection of the facility should be conducted by representatives of the using command, installation agency, engineering agency, and the operating unit.

11-1690. SECURITY. The Air Force Security Service should provide supervision of the maintenance performed on infrared security equipment. Modification, fabrication, and local invention and construction of infrared security equipment are prohibited without proper authorization from the security service.

11-1691. TEST EQUIPMENT. The test sets used in the field will normally be more rugged and compact than those test units used for bench work in depots or other fixed installations. The test equipment used aboard an aircraft is lighter and, consequently, less accurate than depot test equipment. Normally, this lightweight trouble-shooting equipment is not adequate for major calibration or alignment procedures. Depot level test sets may be quite elaborate, and will probably be firmly mounted on workbenches to prevent jarring and vibration. These test sites will contain the maximum possible number of components which are standard in all services, for standardization purposes. This will permit cross-servicing and cross-procurement, as well as emergency supply assistance among the different branches of the Armed Forces.

11-1692. OPTICAL TEST EQUIPMENT.

11-1693. GENERAL. Infrared optical assemblies normally do not require alignment unless you have replaced optical components. The alignment procedures for elaborate optical assemblies on a large infrared equipment may require special training prior to alignment performance.

11-1694. Special test areas may be necessary for properly aligning infrared optical components. Usually a large area, capable of being darkened, is required. A stationary source of infrared radiation, possibly a large incandescent lamp, is also required to perform the alignment procedures.

11-1695. BASIC LENS CHARACTERISTICS. The focal length of a lens is the physical distance from the optical center of a lens to a point where it will bring parallel light rays to a focus. Lenses that are thin have long focal lengths, and are said to be weak. Thick lenses have short focal lengths, and are said to be strong. The terms weak and strong are references to the converging power of a lens, that is, the ability of a lens to refract rays and to bring them to a focus within a certain distance.

11-1696. It is often necessary to measure the convergence, or optical power, of a lens. The unit of measure is the diopter. Weak, long-focus lenses have small diopter values, while strong, short-focus lenses have large diopter values; ie, the lens power is inversely proportional to its focal length. The mathematical definition of a diopter is:

$$\text{Diopter} = \frac{1}{\text{focal length (in meters)}}$$

Thus, a lens with a focal length of 1 meter has a power of 1 diopter, and a lens with a focal length of 1/2 meter has a power of 2 diopters.

11-1697. In any lens assembly, the object distance and the image distance bear a definite relationship to each other. This relationship may be expressed by the simple lens formula which follows:

$$\frac{1}{\text{focal length}} = \frac{1}{\text{object distance}} + \frac{1}{\text{image distance}}$$

11-1698. The optical measurements used to test, align, calibrate, or adjust infrared receivers are similar to those used for telescopes. In addition, tests have been devised to measure the sensitivity of the phosphors or image tubes associated with infrared receivers.

11-1699. COLLIMATION. All infrared receivers have an optical assembly, usually of the reflector type, which collects radiation from the target and projects it in the form of a small, inverted image on the phosphor button of the receiver, or on the cathode of the image tube, in the case of electronic-type instruments. The range of the receiver and the clearness of image resolution are dependent on the accuracy of focus provided by the optical assembly. Because receivers may be used to view objects at several miles distance, they are adjusted for focus at infinity. Under such conditions, the wave front consists of parallel light waves. Therefore, before any infrared receiver can be checked for correctness of focus, it is necessary to duplicate this condition in the test equipment. The process of refracting a wave front into parallel waves is known as collimation, and is accomplished by means of special equipment.

11-1700. The collimating equipment used for testing infrared receivers includes a pattern target of lines or circles illuminated by infrared radiation, and a lens that is larger in diameter than the objective of the receiver to be tested. The lens is mounted on a track to permit its free movement down a graduated scale toward or away from the target. The target is then placed in the principal focal plane of the lens, where the rays coming from the lens are parallel, just as though the target were at infinity. When you move the collimator lens toward the target, the rays become more divergent, and when you move the lens away from the target, the rays become more convergent. Therefore, when you view a target with an infrared receiver through the collimator lens, it is possible to establish whether the receiver is accurately focused for infinity simply by moving the collimator lens and observing the effect of the focus through the receiver. The clearest image should be obtained when the collimator is in its normal position, for only then is the receiver objective in correct focus for infinity. You can readily ascertain any deviation from correct focus by noting the distance the collimator lens is moved to bring the image into focus. Deviations from correct focus can be obtained if the position of the collimator is indicated on a linear scale calibrated in diopters.

11-1701. THE DIOPTOMETER. The image formed on the phosphor button or the screen of the electron image tube of an infrared receiver is too small and too close to the eye to be seen. An eyepiece (or ocular) is provided to produce a vertical image magnified 10 to 15 times, and sufficiently far from the eye of the observer to be clearly seen. In case of phosphor-button receivers, where the image on the phosphor is upside down, the erector lens serves to turn the image right side up.

11-1702. Signals viewed through infrared receivers are usually of low brightness, and can be seen only by an observer who has normal eyesight and is partially or completely dark-adapted. Under these conditions, the virtual image in the receiver eyepiece is most clearly defined and most easily seen when the eyepiece is adjusted to a power of from minus 1 to minus 2 diopters.

11-1703. The power of the eyepiece can be measured by means of a dioptometer. In principle, the dioptometer is a tiny telescope, the eyepiece of which can be adjusted until cross wires located at the principal focus of the eyepiece are sharply defined. When this has been done, light waves through the eyepiece of the dioptometer enter the eye of the observer as parallel waves.

11-1704. In using this instrument, first set up the infrared receiver with the eyepiece to be checked, so that it is illuminated by a collimated signal. Then take the dioptometer, holding the knurled end ring next to the eye. Peer into the eyepiece and turn the knurled ring until the cross-line reticule comes into sharp focus. This operation sets the eyepiece for the individual using the instrument; any other person using the instrument will have to make this same adjustment. Now unscrew the diopter ring to the extreme negative position of -5. Place the dioptometer against the eyepiece of the receiver, and screw the diopter ring in until the image comes into sharp focus. Take the reading from the diopter ring. The reading obtained in this manner is a measure of the eyepiece adjustment of the receiver, and should be -1 or -2 diopters.

11-1705. The dioptometer can also be used to check the collimating lens. When the collimator is in the proper position for checking a receiver, a reading of zero should appear on the diopter scale when light coming through the collimating lens is observed through the dioptometer.

CAUTION

The dioptometer is a precision instrument and must always be kept in a safe place. Do not permit it to remain in direct sunlight or in any heated area.

11-1706. THRESHOLD. The threshold of a receiver is the smallest amount of infrared radiation that can be seen through the receiver by a dark-adapted observer, using for the infrared source a lamp having a specified color temperature, and using a filter of specified spectral transmission. The threshold figure, therefore, is a measure of the sensitivity of an infrared receiver.

11-1707. The unit of measurement used in threshold tests is the nautical-mile-candle (nmc). It is a very small unit (there are approximately 36,000,000 nautical mile candles in 1 foot-candle). One nmc is the illumination received at a distance of 1 nautical mile from a source of 1 candle power. If an instrument can detect a source of 1 candle power at a distance of 1 nautical mile, then the threshold of the instrument is 1 nmc. The numerical value of this unit is inversely proportional to the actual sensitivity of the instrument. In other words, an instrument which requires a source of 4 candle power to operate at a distance of 1 nautical mile is a more sensitive instrument than one rated at 5 nmc. The finest instruments, therefore, are those of low nmc value; the least sensitive instruments are those of high nmc value.

11-1708. A poor threshold figure in a receiver may indicate a defect in the optical assembly, a faulty phosphor, or a defective charging system; it may also indicate a faulty image tube or a defective electrical system in the case of electronic-type instruments.

11-1709. RESOLUTION. The resolution of an infrared receiver is a measure of its ability to distinguish between two point sources separated by a small angle. The unit of measurement used in resolution tests is the angle of separation of the two points, given in minutes (60 minutes in 1 degree) of arc. Unsatisfactory resolution in a receiver usually indicates poor optical alignment or focus. Threshold and resolution tests can be made on a specially constructed infrared optical tester.

11-1710. THE INFRARED OPTICAL TESTER. The infrared optical tester is essentially a point light source which is placed at infinity by means of a collimator lens. This light source consists of two pieces of colorless glass ground on both sides and illuminated by a coiled filament tungsten lamp. The radiation from this source is given the proper spectral characteristics by passage through a standard infrared polaroid filter. The intensity of the light may be varied by known increments by interposing apertures of different sizes over the ground glass, and by moving the tungsten lamp measured distances from the ground glass.

11-1711. Since low light intensities are desired at the receiver for making threshold tests, a lens is placed between the light source and the collimator to reduce the light intensity and permit the use of a short-focal-length collimator. Without this lens, a collimator lens of very long focal length (about 30 feet) would be required to give the same effect.

11-1712. A hole plate in the optical tester contains 10 threshold apertures and 6 resolution patterns. The threshold holes vary in diameter from 0.027 to 0.177 inch to provide threshold values from 0.25 to 10 nautical-mile-candles. The resolution patterns are pairs of holes separated by various distances to correspond to resolution readings of 4, 6, 10, 15, 20, and 25 minutes of arc. (The separation of the holes in tenths of an inch exactly equals their separation in minutes of arc as seen through the optical assembly.) It is important that the threshold holes be kept clear of dust and dirt. The microscope objective and the collimating lens are mounted in threaded sleeves to facilitate adjustment. These are set and locked at the time the tester is calibrated, and should not be moved. The microscope objective can be removed for cleaning, storage, or shipment. When replaced, it should be screwed all the way into the sleeve.

11-1713. The hole plate is connected through gears to a shaft which is rotated by means of the knob on the front panel of the tester. The holes are arranged so that threshold and resolution values increase as the knob is rotated in a clockwise direction. Connected to this shaft is the detent plate which locates the holes as the hole plate is rotated. If it is desired to work between two definite values, as in inspection work, the detent is pulled out to its first slot, rotated, and locked. The detent stops are then screwed to the detent plate at the positions where the desired holes are in place when the stop strikes the detent. The detent can be pulled out further and locked again in such a position that the stops will clear the detent and the shaft can be rotated its full 360 degrees. The values in nautical-mile-candles and minutes of arc are indicated through a hole in the front panel by self-luminous numbers on the detent plate. These numbers can be read in total darkness without the aid of a flashlight or auxiliary lamps.

11-1714. Before threshold readings can be made, you must be completely dark-adapted. This is accomplished in one of two ways. The first method is to remain in total darkness for 30 minutes; the second is to wear approved dark-adaptation goggles for 20 minutes and then spend 5 or 10 minutes in the dark before starting to make the readings. Either method accomplishes the same degree of dark adaptation, the latter permitting you to work under normal lighting conditions. Once dark-adapted, the eyes should by no means be exposed to white light, and only to low levels of red light.

11-1715. After you are dark-adapted, you must learn to use your night eyes. Readings should be made at the dim threshold, which is defined as the lowest level of illumination at which the point source is still continuously visible, but does not tend to appear and disappear as at the absolute threshold. In this text the dim threshold is referred to simply as the threshold.

11-1716. To make the readings, the instrument to be tested is placed on the jig in front of the collimator, and the light value turned up to about 10 times the expected threshold. This will enable you to pick up the spot in the field. Then, while observing the spot, adjust the light value until the threshold is reached. You can now read the nautical-mile-candle rating directly from the scale.

11-1717. ELECTRONIC TEST EQUIPMENT.

11-1718. GENERAL. The replacement of components or units in the electronic portion of infrared equipment may necessitate the resetting of one or more adjustments. Adjustments are always necessary after the replacement of variable components or of fixed components that influence the operation of a variable. Even when exact replacements are used, differences must be accommodated. The differences lie within the manufacturer's tolerance, but in critical circuits they sometimes assume considerable importance. The adjustments made necessary by replacements and repairs may, in some instances, require a complete realignment procedure. After all replacements and repairs have been made, you should check the complete equipment for proper operation, and follow through on all adjustments that are necessary.

11-1719. Many infrared equipments contain a method of calibrating the sensitivity of the amplifying circuits. This check serves as a performance standard for equipment operation, and also as a test to determine whether or not the infrared detector unit is at fault when the equipment fails. This is accomplished by feeding a calibrated signal into the amplifier immediately following the infrared detector unit. The known voltage introduced into the amplifier unit gives an indication of the over-all amplifier response.

11-1720. Electronic-type instruments may be tested on an infrared optical tester. To test a receiver, adjust the infrared source for a circle of 1-inch diameter on the photocathode; then multiply the dial threshold reading by the receiver conversion factor to obtain the true threshold reading. These readings should not be more than 1.5 nmc. Resolution tests may be conducted by projecting the resolution pattern on the photocathode. Set the focus voltage for equal horizontal and vertical resolution in the central areas; the resolution should be at least 8 lines per millimeter in the center of a 0.3-inch-diameter circle from the cathode center, and 2.2 lines per millimeter in the periphery (0.7-inch-diameter circle). The term lines per millimeter means pairs of equal width of both black and white. The number and size of the spots or blemishes within the 0.3-inch-diameter and 0.7-inch-diameter circles should conform to the data provided in table 11-13.

Table 11-13. Infrared Receiver Resolution Data

MAXIMUM NUMBER OF SPOTS		SPOT SIZE
CENTER (0.3-INCH DIAMETER)	TOTAL (0.7-INCH DIAMETER)	(IN MILLI-METERS)
0	0	Over 0.35
0	8	0.25 to 0.35
5	16	0.08 to 0.25
5	20	

11-1721. You should handle electronic instruments with care, both before and during testing. Observe the following precautions carefully: Do not permit light to enter the objective or the eyepiece while taking threshold readings, or from 10 minutes prior to taking threshold readings. Do not take threshold readings in rapid succession with the same instrument because the instrument, retaining a previous glow, will give you a false indication.

11-1722. STANDARD INFRARED DETECTOR TEST EQUIPMENT AND PROCEDURES. The test procedures employed with infrared detectors may be divided into two groups; detector efficiency tests, which require three separate series of measurements with radiation sources, and tests which yield the root power spectrum of the noise. The ratio of the detector efficiency to the noise can be used to indicate the quality of the detector. This ratio is called the detectivity of the detector.

11-1723. To measure the detector efficiency, the signal radiation shall be normally incident on the detector, and the amount of signal radiation will be confined to the range in which the output signal is proportional to the incident power.

11-1724. The calibrated signal generator required for infrared detector tests must produce a voltage sine wave of accurately known rms amplitude in the proper frequency range. Normally, the range will be adjustable from 1 to 100,000 cps, with an approximate amplitude of 1 volt rms. This signal will be passed through a calibrated attenuator for a predetermined amplitude reduction

11-1725. A tunable filter, with a center frequency which can be operated over the required range and with a stable gain (ie, independent of filter tuning) is required for detector efficiency measurements.

11-1726. A multirange voltmeter, to measure both the signal and noise voltages, is required. The voltmeter should indicate true root-mean-square amplitude of the waveform; however, corrective calculations might be required, depending on the waveform of the measured signal.

11-1727. Three sources of radiation are required. A black-body source, which must be frequency-chopped and stable in temperature, should produce an accurately known spectral irradiance uniformly over the responsive surface of the detector area. A monochromatic source of radiation consists not only of radiation, but of a monochromator with a chopper placed between it and the source. The cnopper may be for a single fixed frequency, the same as was used for the black body source. The irradiance of the monochromatic source should also be uniform over the responsive surface of the detector. A variable-frequency source must also be available, with a stable source of radiation and a variable frequency chopper. The source of radiation should be a black-body source, and a filter should be available.

11-1728. The detector-impedance and pulse-time-constant measurements may be performed with normal test equipment, such as that used with other electronic or communication circuits.

11-1729. TYPICAL INFRARED EQUIPMENT APPLICATION.

11-1730. INFRARED SNIPERSCOPE (SET 1, 20,000 VOLTS). The sniperscope (Set 1, 20,000 volts) shown in figure 11-256 is a device used for seeing objects in the dark without visible light illumination. It can be used for signaling or detecting enemy infrared equipment at night. In addition, it can be mounted on a carbine or rocket launcher, as an aid to firing at night, because it has a sight, as shown in figure 11-256.

11-1731. A 30-watt bulb behind a filter which removes the visible light is mounted over a telescope sight, as shown in figure 11-256. The invisible light passing through the filter will illuminate the target at which the light is directed. The reflected infrared rays entering the telescope, as illustrated in figure 11-257, are focused on the electronic tube by the objective and corrector lens assemblies in the telescope. The infrared rays actuate electrons that bombard a fluorescent screen, which forms an image similar in shape to the original target. The image on the screen is magnified by the eyepiece lens assembly. All images appear in various shades of green through the sniperscope.

11-1732. SNIPERSCOPE (MODELS M1 AND M2). Basically, these sniperscope models are the same as Set 1, described in paragraphs 11-1730 and 11-1731, and illustrated in figures 11-256 and 11-257. However, Models M1 and M2 use an operating power

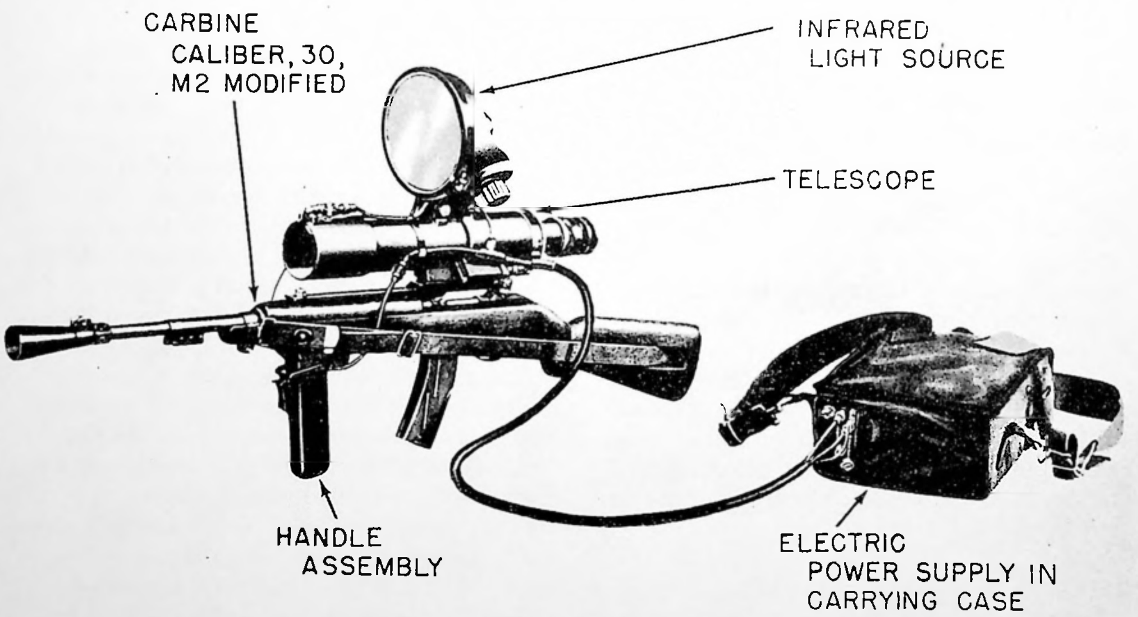


Figure 11-256. Infrared Sniperscope, Set 1, 20,000 Volts

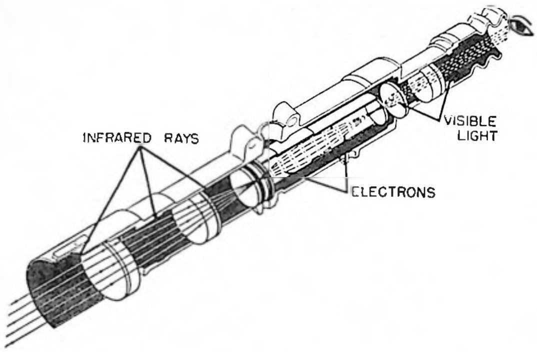


Figure 11-257. Conversion of Infrared to Visible Light Within a Telescope

pack of 4250 volts. Figure 11-258 illustrates the telescope assemblies of both models preparatory to being installed on a rifle or other mounting. Figure 11-259 illustrates the sniperscope on a hand-held mount. On Model M1 and the original issue of Model M2, the infrared lamp source was located below the mount, with a separate grip and trigger from those of the carbine or other mount. However, in later models of the M2 the infrared lamp source is located over rather than under the mount.

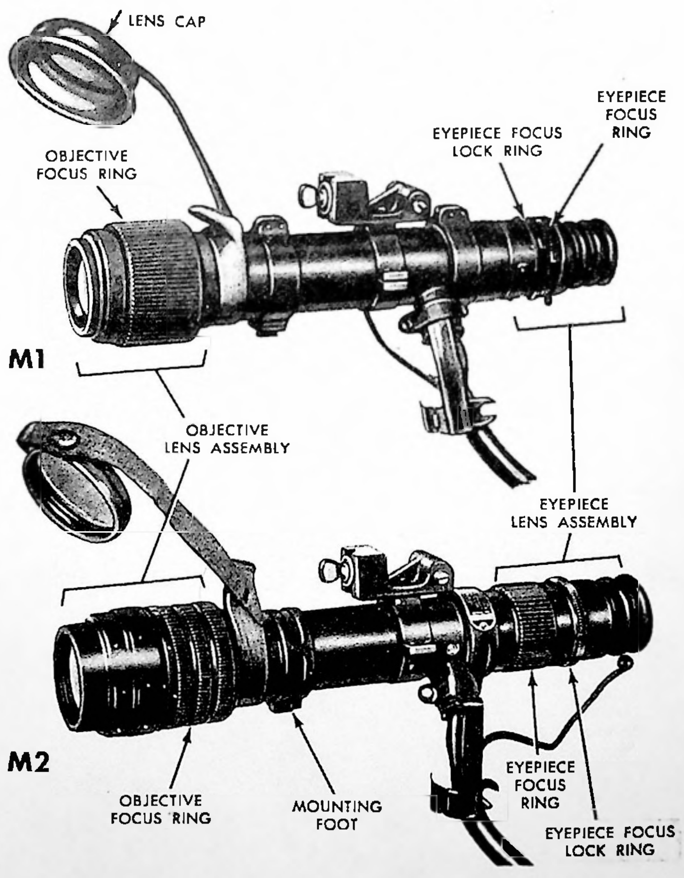


Figure 11-258. Infrared Telescope Assemblies, Models M1 and M2

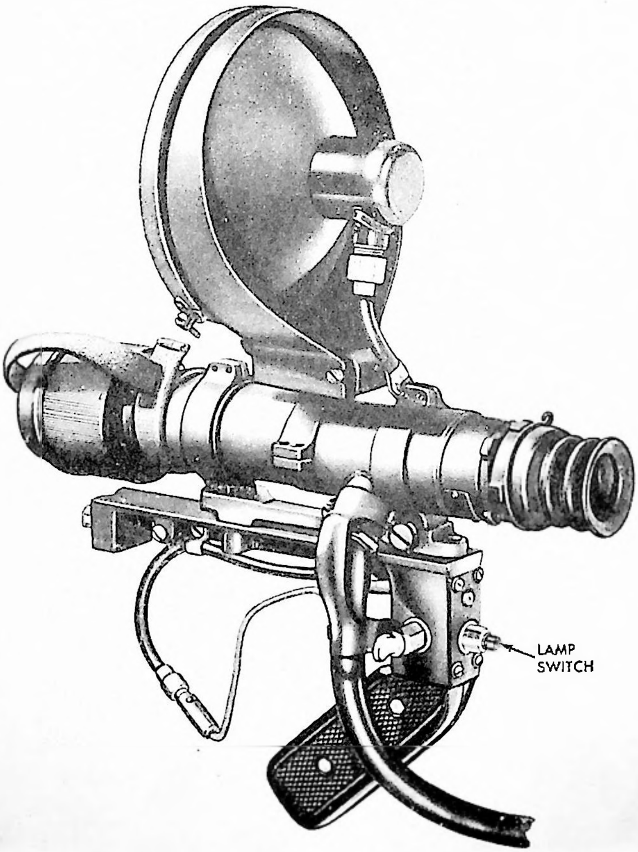


Figure 11-259. Infrared Sniperscope with Hand-Held Mount

11-1732A. STANDARD INFRARED HEAT SOURCE CALIBRATION.

11-1732B. GENERAL.

11-1732C. All matter is made up of molecules which are in a constant state of agitation. There is no molecular motion at the temperature of absolute zero (0 degrees K or −273 degrees C), but the motion occurs and increases in amplitude and frequency as temperature is increased. These molecules of matter comprise atoms consisting of electrical charges which cause a radiation of electromagnetic energy. Thus, the thermal agitation of molecules above the absolute zero temperature causes radiation of electromagnetic energy. It is this electromagnetic energy generated by molecular thermal action in all matter that is defined as infrared radiation. The number of molecules in a given body is extremely large; however, not all of the molecules are excited to the same degree of thermal agitation. If a curve were constructed to show the distribution of numbers of molecules versus molecule excitation at any temperature, it would be seen that there is a peak where most of the molecules have approximately the same degree of thermal agitation. The sloping sides on such a curve would indicate that there are smaller numbers of molecules having higher or lower degrees of agitation. The distribution pattern of this molecular agitation corresponds to the distribution pattern of the electromagnetic energy; therefore, electromagnetic radiation is actually emitted over a wide range of frequencies. In accordance with the manner in which the distribution pattern of molecular agitation evolves in an object as the result of temperature increases, the total energy within the object also increases, and the peak distribution shifts to shorter wavelengths.

11-1732D. It is generally accepted that the infrared radiation spectrum lies between 0.72 micron and 1000 microns, that is, between the borders of visible light at the shorter wavelength end and of microwaves at the longer wavelength end of the infrared spectrum.

11-1732E. CHARACTERISTICS OF RADIATION.

11-1732F. Objects continuously interchange radiation by means of the reciprocal processes of absorption and of radiation. Some objects completely absorb all radiation incident upon them, and conversely, emit all radiation. Other objects only partially absorb and emit infrared radiation. For imperfect absorbers and radiators, the ratio of the energy absorbed by such an object to the energy absorbed by a perfect absorber at the same temperature is the same as the ratio of energy emitted by that object to the energy emitted by a perfect radiator at the same temperature. In developing the techniques of infrared radiation, this ratio development led to the concept of the black body, an opaque object that absorbs all incident radiation and emits more radiation than any other comparable object at the same temperature. Such energy is greater in total energy per unit area than that emitted at or near a specific wavelength which is dependent upon the temperature of the object.

11-1732G. Since no perfectly black material has ever been discovered, a description of the radiation from physical bodies must include a factor which is related to their degree of blackness. This factor is known as emissivity, which is defined as the ratio of the radiation emitted by an imperfect radiator and absorber (which is also called a gray body) to the radiation emitted by a black body at the same temperature. Calculations to determine the characteristics of infrared radiation are based upon three laws; the Stefan-Boltzmann Law, Wien's Displacement Law, and Planck's Equation. Each of these relationships is discussed in preceding paragraphs.

11-1732H. BLACK BODY SIMULATION.

11-1732I. GENERAL. A standard source of radiation of known intensity and spectral distribution is required in photometry, pyrometry, and radiometry. The source need not be a black body; however, experience has shown that it is much simpler to build a standard radiation source with characteristics approaching the black body, than it is to construct any non-black body source of equal precision. A practical source with properties approaching those of the perfect black body is called a black body simulator.

11-1732J. Some of the applications of the black body simulator are: to supply radiation of known intensity; to supply radiation of known spectral distribution; to define points on the International Temperature Scale; to act as a comparison reference for determining the intensity of other sources of radiation; and to act as a comparison reference for determining the emissivity of materials.

11-1732K. The classical black body simulator is a hollow spherical cavity made of an opaque material, and pierced by a small aperture. If the enclosure is insulated from outside thermal influences, all parts of the internal cavity walls will eventually reach the same temperature. When equilibrium is reached, there will be equal energy at all points inside the cavity, regardless of the nature of the walls. The radiation passing through the small aperture will be essentially black body radiation. Almost all black body simulators are constructed in the form of cavities. The most common types of these simulators have been spherical, cylindrical, conical and re-entrant biconical cavities.

11-1732L. Since a black body absorbs all of the radiant energy incident upon it, a practical cavity approaches a black body to the extent that radiation entering through the aperture is absorbed within the cavity. Since all the materials which are used to make these practical cavities have absorptivities less than unity, the performance of a black body simulating cavity is based on multiple internal reflections. If, for example, the inner surface of a black body cavity has an absorptivity of 0.9, an incoming ray will be 90 percent absorbed at the first incidence so that only 10 percent is reflected. If this remaining 10 percent strikes another portion of the internal surface, 90 percent will again be absorbed so that only 1 percent of the original ray appears after the second reflection. The apparent surface represented by the aperture is then effectively a surface with at least 99 percent absorptivity. This apparent surface closely approaches the properties of a black body.

11-1732M. The major problems associated with designing black body simulators are those of devising aperture cavity configurations and the selection of suitable materials so that all of the radiation entering the cavities through the apertures will be absorbed by multiple internal reflection and absorption.

11-1732N. One of the most recent studies on the emissivities of black body simulators shows that the conical cavity, with appropriate baffling, has higher emissivity than either cylindrical or spherical cavities. This condition indicates that emissivity is independent of cavity wall smoothness and minor temperature variations.

11-1732O. TYPICAL BLACK BODY SIMULATOR. A typical black body simulator is contained in a cylindrical aluminum housing which contains a conical radiation core around which a heating element is wound. The typical black body simulator also contains a platinum resistance thermometer sensing element and a front aperture plate. The aluminum housing and the radial I-beam type of mounting are made from a single casting in order to ensure not only accuracy of alignment and ruggedness of construction, but also to provide an excellent thermal sink. The radiation surface, together with its heating and sensing elements, are contained within a very durable and resilient insulating material. An insulation plug, when fastened by its securing screw, compresses the insulating material between the bulkhead disk and the compression plate. This arrangement not only provides a well-insulated core, but it also affords a rigid mount for the core. Two insulating washers should also be mounted between the front aperture plate and the compression plate. As a result of the particular construction used in this type of black body simulator, the temperature of the housing can be kept to a minimum and the temperature of the front aperture plate will remain essentially at ambient temperature. Heat can be supplied to the core assembly by means of the heating element which receives its power from a temperature controller. A continuous indication of temperature can be sent back to the temperature controller by means of the sensing element in close contact with the core casting.

11-1732P. The radiating surface of a typical black body simulator should have a conical cavity with an apex angle of 20 degrees. A study of black body theory and empirical checks will show that a radiator of this geometric design will have an apparent emissivity of over 0.99 for sidewall emissivities of over 0.65. Since the sidewalls of the cavity have an emissivity of approximately 0.9, the total hemispherical emissivity is effectively unity and is independent of any aging effects or minor changes in the properties of the cavity walls.

11-1732Q. INFRARED RADIATION POINT REFERENCE SOURCE. An infrared radiation point reference source is an instrument designed to provide black body infrared radiation which originates essentially from a point source. A single direct-reading dial is used to set the temperature of the radiating surface, and this dial is directly calibrated over a range of from 200 degrees C to 600 degrees C. Source temperature rapidly reaches the selected level and the instrument is ready for use when a front panel meter reaches equilibrium. Special features of the instrument include accuracy, high radiation source emissivity, and ruggedness of design. The radiating surface temperature is precisely maintained as selected and remains virtually unaffected by changes in ambient temperature, line voltage variations and transients, tube aging, and replacement of most of the electronic components. Electrical isolation from line voltage variations and possible interference from other equipment is obtained through the integrated use of both vacuum-tube and magnetic amplifiers. The instrument is suitable for use in any application in which a point source of black body radiation of known characteristics is required. Converging, diverging, or parallel radiation may be obtained through the use of suitable optical equipment of a supplementary character.

11-1732R. TYPICAL STANDARD IR HEAT SOURCE CALIBRATION METHOD.

11-1732S. GENERAL. A standard infrared (ir) heat source calibration method provides a means of calibrating test sources of infrared energy by radiometric comparison with a reference source whose radiation characteristics are accurately known. This method operates on a null principle. It can accommodate test sources, having aperture sizes from 0.0087 inch to 0.50 inch in diameter, which are operating at temperatures from 50 degrees C to 600 degrees C. Sources up to 12 inches in diameter and 50 pounds in weight can be calibrated. The method employs the following major equipments; a controlled ir source, a temperature controller, an ir source comparator, an amplifier, a power and bias supply, a calibrated thermocouple, and a recorder. The power and bias supply, the recorder, and the temperature controller are mounted in a relay rack cabinet.

11-1732T. Calibration of a test source requires the following additional equipment: a source to be tested (with a temperature controller), a potentiometer, a galvanometer, a standard cell, battery and test leads. A block diagram of the standard ir heat source calibration equipment is shown in figure 11-259A.

11-1732U. Three types of detectors may be used with the comparator. A lead sulphide detector with an attached optical filter is used when calibrating a rate-table infrared radiation source. This optical filter is commonly referred to as a Z filter, and it appears to possess a green color when viewed by the human eye. The detector used in the comparator should have a bandpass identical to the bandpass used in calculating infrared power level.

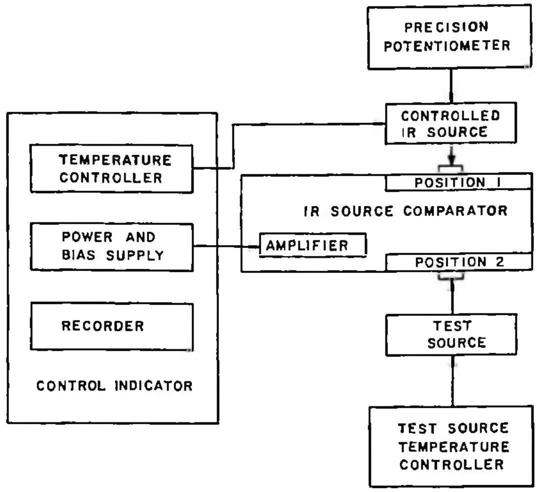


Figure 11-259A. Block Diagram of Typical Standard IR Heat Source Calibration Equipment

11-1732V. The temperature of the simulator's emitting surface is regulated by the temperature controller. By setting the control dial to the desired temperature the radiating surface will approach this desired temperature and, when the voltmeter on the front panel reaches a stable minimum indication of the temperature selected, no further attention is required.

11-1732W. Radiation source temperature is automatically maintained at the desired level by means of a proportional control which is based upon an electronic comparison between the temperature of the emitting surface and the setting of the control dial. As a result of this comparison, the power to the source heating element is adjusted in order to make the temperature of the emitting surface correspond with the temperature indicated by the control dial. Once the selected temperature is reached, the amount of power delivered to the heating element is that amount required to maintain this temperature while making up for heat losses to the surroundings.

11-1732X. The major functional elements of the temperature controller include a bridge network, a vacuum-tube amplifier, and a thyratron with its magnetic amplifier control circuit. The temperature controller forms a high-gain, closed-loop servo arrangement with the simulator. Prominent features of this arrangement are proportional control and essential independence from line voltage variations. Mounted on the front of the temperature controller is a temperature selector dial equipped with a vernier drive giving a 3-to-1 turn ratio between the knob and dial. A front panel meter measures the voltage across the heating element, and the meter reading indicates qualitatively when the radiating surface is at the selected operating temperature.

11-1732Y. The bridge circuit contains a sensing element. The power transformer has a 10-volt winding which applies across this bridge. The bridge output is coupled to a two-stage coupled amplifier with a cathode-follower output stage. Since the resistance of the sensing element changes with temperature, the bridge network output to the amplifier is proportional to the temperature difference between the sensing element and the setting on the variable temperature control resistor. The output of the amplifier is identical to that of the bridge network, except that it is highly amplified.

11-1732Z. The output of the cathode follower is coupled to one winding of a saturable reactor. The reactor acts as a magnetic amplifier. The current flow in both windings is polarized by means of selenium rectifiers. Output from the second winding of the magnetic amplifier is coupled to the grid of a thyratron in series with a 15-volt winding of the power transformer.

11-1732AA. In order for a thyratron to conduct, its plate must be positive with respect to the cathode and its grid voltage must be more positive than the critical bias. Since ac voltage is supplied to the thyratron plate, it can conduct only during the positive half cycle. Only when the temperature of the simulator is below the setting on the control dial is the thyratron grid signal of the proper polarity needed to fire the thyratron. Thyratron plate current flows through the simulator's heating element. The power delivered to the radiation surface is that which can bring the surface most rapidly to the desired temperature and then to maintain this condition.

11-1732AB. OPTICAL BALANCING. The purpose of optical balancing is to ensure that equal amounts of energy are focused onto a lead sulphide detector when radiation traverses the optical path from either aperture 1 (path 1), or aperture 2 (path 2) to the detector when identical radiation sources and apertures are placed at positions 1 and 2. Identical radiation sources must be used when balancing is facilitated by this method because the reference source is physically placed at either positions 1 or 2.

11-1732AC. Identical apertures should be used for balancing purposes and this stipulation is fulfilled by using a field stop with a 0.015 inch diameter and precision apertures with 0.040 inch diameters at positions 1 and 2. The apertures at positions 1 and 2 are large enough such that the field stop is the limiting aperture in the optical arrangement. In effect, this method uses a single aperture for optical balancing to eliminate any balancing errors due to unequal aperture shapes or areas.

Table 11-13A. Optical Balancing Chart

Radiation stop	None
Detector	With Z filter
Field stop	0.015 inch
Apertures 1 and 2 (precision)	0.040 inch
Reference source temperature	290 degrees C
Bandpass	Filter out
Bias supply test	Plus or minus as required
Buckout and test signal	As required
Power	On
Range selector	225 degrees - 600 degrees C
Temperature dial setting	As required
Gain	Maximum obtainable without excessive oscillations*

* Overly high gains cause saturation of the amplifier resulting in improper nulls.

11-1732AD. Equal amounts of radiation should be reflected by the mirrors onto the detector when the reference source is placed at positions 1 and 2. The optical attenuator (a triangular-shaped blade) is placed in path 2 primarily to compensate for this extra reflecting surface. The optical attenuator also compensates for differences in reflectivity of all the mirrors. Adjustment of the optical attenuator to achieve optical balancing completes the balancing phase. Table 11-13A provides a complete listing of dial settings and components used for the optical balancing procedure.

11-1732AE. After obtaining optical balance, null the recorder by blocking apertures 1 and 2. Adjust the buckout signal and balance controls until a null is obtained with the chopper rotating and with the gain control turned to a maximum setting (without excessive oscillations).

11-1732AF. IR SOURCE COMPARATOR. The basic radiation comparison method is provided in the ir source comparator. This instrument is essentially a photometer which alternately images two apertures at a single field image plane. The energy at the field image plane is collected by an ellipsoidal mirror and reimaged down on a detector. The comparator is provided with 3 types of detectors, two sets of precision apertures, one set of field apertures, radiation shields to limit the energy incident on the detector, an adjustable stage to support the test sources, and a fixed stage for the reference source. The comparator also houses the amplifier.

11-1732AG. AMPLIFIER. This unit amplifies the test signal and the output from the detector and the output of the amplifier is applied to a synchronous rectifier assembly mounted on the chopper mirror drive in the comparator.

11-1732AH. POWER AND BIAS SUPPLY. The power and bias supply contains the battery supply for the detector bias and filament and the B supplies for the amplifier. The main operating controls for the calibration equipment are mounted on the front panel.

11-1732AI. CONTROLLED IR SOURCE. The controlled ir source unit provides a precisely known and controllable level of radiation flux.

11-1732AJ. TEMPERATURE CONTROLLER. The temperature controller provides controlled power to heat the ir source. The desired temperature can be set on a front panel dial, and the controller will provide power to heat the source and to maintain it at the desired temperature.

11-1732AK. RECORDER. The synchronously rectified output from the comparator is fed to the strip chart recorder, which serves as an indicator of the balance between the radiation levels from the test and the reference sources.

11-1732AL. CALIBRATED THERMOCOUPLE. The actual temperature of the reference source is measured by the precision thermocouple, which serves as an accurate means for determining operating temperature.

11-1732AM. Several modes of operation are possible with the standard ir heat source calibration method providing various ways of calibrating a particular type of source. The primary mode of operation which will provide the most accurate method of calibration is simultaneous comparison of the radiant flux from the test and reference sources by alternately imaging the radiation from their respective apertures on a single detector. This alternate imaging process is referred to as chopping.

11-1732AN. In the optical diagram shown in figure 11-259B, notice that in one position of the semicircular chopper mirror, the toroid mirror will form an image of aperture 1 at the field stop. Reflective mirrors 1, 2 and 3 have plane surfaces and are used to redirect the beam. When the chopper mirror is in the radiation beam, it blocks energy from aperture 1 and permits the toroid to image aperture 2 at the field plane. The ellipsoid mirror images the field plane onto the detector reducing the field image size by a ratio of 6:1.

11-1732AO. The output from the detector will be a direct function of the radiant flux from the source being imaged upon it. If the total energy being imaged on the detector is equal for each source, an unvarying output level will be obtained. If there is a difference between the sources, the detector output will vary, being higher for the

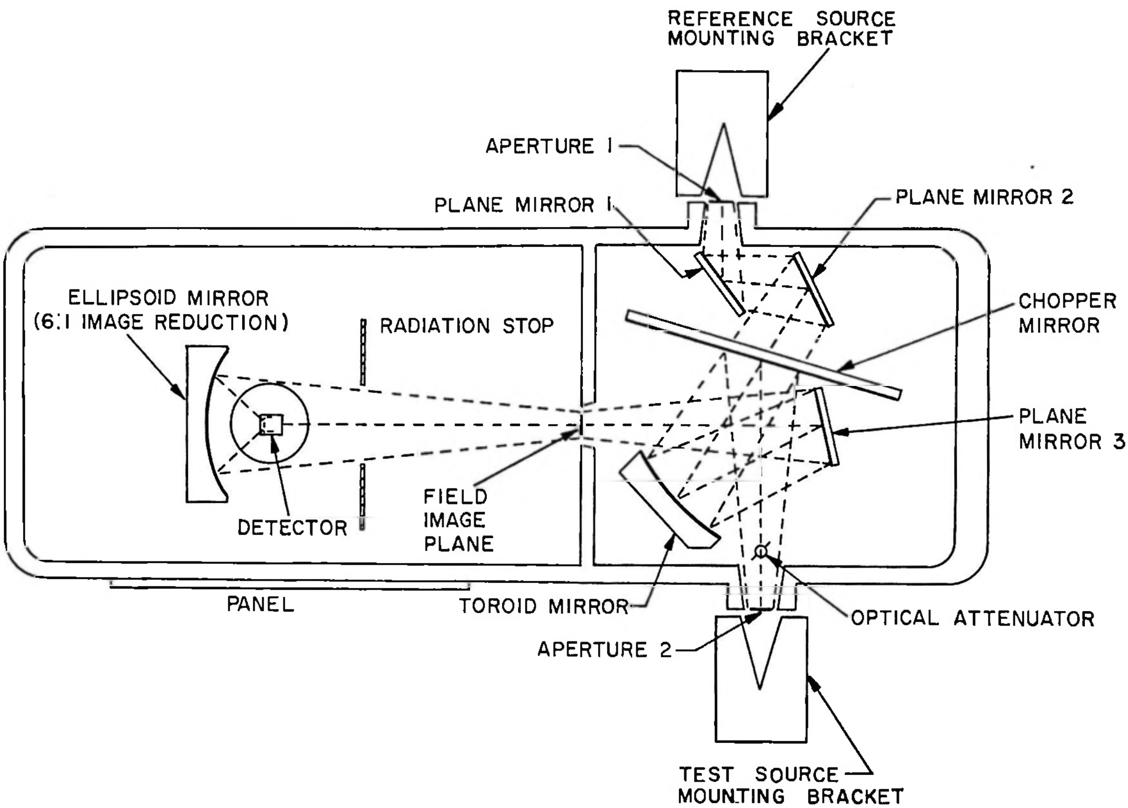


Figure 11-259B. IR Source Comparator Optical Diagram

source of greater radiant output and lower for the source having less output. This variation in output occurs at the rate of alternation between the sources, and is referred to as the chopping frequency. In this equipment it is 13 cycles per second, as determined by the rate of rotation of the semicircular plane mirror 3.

11-1732AP. The detector output has both a dc and an ac component. The dc component is a function of the level of radiant energy incident on the detector, while the ac component will be a function of the difference in level of energy from the two sources. Only the ac component is useful, since the purpose of this method is to provide a means for reducing the difference in radiant output between two sources to zero. This is achieved by adjusting the temperature level of the test source until the ac component is either zero or null.

11-1732AQ. While the instrument is primarily intended for simultaneous comparison of sources having equal apertures operating at equal temperatures, it is actually possible to calibrate either simultaneously or nonsimultaneously over reasonably large limits, sources with different aperture areas and/or operating temperatures. This is possible with the use of a test or buckout signal which electrically simulates a radiation signal from one of the sources. The buckout signal adds to or subtracts from the test signal, as desired.

11-1732AR. The buckout signal can be preset to represent a given reference source radiant power level by obtaining a null between it and the reference source. The reference source can then be removed, and the test sources calibrated against the buckout signal by adjusting their radiant output to obtain a null. This method of nonsimultaneous comparison is not so accurate as simultaneous comparison with a reference standard.

11-1732AS. An ideal comparator method for calibrating a test source against a reference standard is the single aperture simultaneous comparison method which employs a single aperture for the two sources to eliminate any effects due to unequal aperture shape or area and which uses behind-aperture chopping to prevent any self-radiance from the aperture adding to the chopped signal. This condition can be achieved with the comparator by locating the precision aperture stop at the field image plane. Either no stops are used, or stops of larger size would be mounted at the source apertures 1 and 2 as shown in the figure. The test and reference source cavities will alternately be imaged at the field aperture. Chopping occurs behind this aperture, so the only measurable contribution to the chopped signal will come from the source cavities.

11-1732AT. Some sources will have apertures attached as an integral part of the source. In this case the separate aperture simultaneous comparison method is used, which determines the over-all effect on radiation at the detector resulting from both source temperature and its aperture surface. This comparison is made with a precision aperture mounted on the comparator as an entrance stop for the reference source at position 1. The test source with its integral aperture of the same nominal size will be mounted for comparison at position 2. A field stop slightly larger than the aperture stop must be installed at the field image plane in the comparator.

11-1732AU. After a null has been obtained by adjusting the temperature of the test source, you are assured that the radiation power levels of the two sources are the same, although their temperatures may be different to compensate for differences in emissivity of the cores and the aperture areas. In this mode of operation, the aperture is chopped, and a hot aperture plate contributes part of the ac signal affecting the null balance.

11-1732AV. The purpose of the field stop is to limit the amount of aperture plate seen by the detector by restricting the field of view. To determine the significance of aperture plate radiation, the field stop should be removed and the balance between the reference and the test sources should be checked again. If the null is maintained without the field stop, it is a good indication that there is no significant heating of the aperture.

11-1732AW. The radiant power level of a source is primarily a function of the aperture area and the source temperature. Varying either or both of these parameters will change the radiated power level. The buckout and test signal may be employed to permit calibration of a source operating at a power level differing from that of the reference source. This is an electrical simulation of the detector output due to a radiation signal. It can be added to or subtracted from the test source signal to achieve a null. It will be necessary to calibrate the amount of buckout signal supplied to ensure that it represents the proper radiant power level. Calibration at a power level different from the reference source may be desirable when a number of similar sources are to be checked rapidly and where several power levels are of interest. In this instance it is desirable to perform a rapid calibration check without having to change the reference source aperture and/or temperature each time that a different power level is checked. This is accomplished as follows:

a. With the reference source at position 1, and operating at the proper temperature and aperture size for power level 1, calibrate a test source of nominal level 1 power by adjusting the temperature to obtain a null.

b. Replace aperture 1 with the appropriate aperture size for level 2 power and set the reference source to the proper temperature.

c. After the source has stabilized, null out the difference between the level 2 power reference source and the level 1 power test source by reversing the polarity of the bias supply switch and adding a buckout signal.

d. Record the position of the coarse and fine settings for the buckout signal. The signal will be removed by switching the coarse control to off.

e. To check a level 2 power source, properly adjust the reference source for a direct comparison, leaving the coarse control in the off position.

f. To check a level 1 power source, switch the buckout coarse control to bring the two levels into approximate balance, and adjust the test source temperature to obtain final calibration.

11-1732AX. The buckout signal may be used to represent the entire radiation signal for non-simultaneous comparison of a test source with the power level of a reference source. As in the preceding modes, the power levels of two sources are being compared and the entire reference level is being simulated by the buckout voltage. This is accomplished as follows:

a. Place the reference source in position 1 and set the reference source to the desired power level with aperture 2 masked.

b. Null the output with the buckout signal.

c. Remove the reference source, install the test source in position 1, and balance against the buckout voltage.

11-1732AY. The recorder chart may be calibrated for several power levels, with one representing zero deflection (null) and others representing higher or lower power levels as evidenced by deflection to the right or to the left of the chart center. However, none of these methods of calibration is as precise as the simultaneous comparison of two sources operating at the same power level.

11-1732AZ. The primary purpose of the buckout and test signal is to provide a convenient means for centering the recorder pen and for providing a means of calibrating the equipment gain and sensitivity by obtaining the chart deflection induced by a test signal input voltage.

11-1732BA. The signal is generated in the power and bias supply, where a fixed resistor and a potentiometer connected between the B supply and ground provide a dc voltage variable from 15 to 120 volts. This is the fine control. A second potentiometer, constituting the balance control, provides for very small changes of buckout voltage. The voltage is applied to a resistance divider (the course control) whose switched output is further reduced by a resistance divider in the ir source comparator. Now at a millivolt level, the voltage is changed from dc to ac by the center set of contacts on the synchronous rectifier assembly which shorts the signal to ground during one half of the cycle. The resultant rectangular waveform is further shaped with an integrator to approximate the detector output waveform before it is fed into the amplifier. It is then mixed with detector signals, amplified, synchronously rectified and applied to the recorder.

11-1732BB. Since the buckout voltage is of one polarity, between zero and some positive voltage, it is necessary to provide a means of phasing it with respect to the chopped radiation signal so that it can be made to add to or subtract from the detector output. This is accomplished by providing a bias reversal switch for the detector called the bias supply test switch.

11-1732BC. When the buckout signal is being employed in the comparison of different power levels or in non-simultaneous calibrations, it may appear that the signal is adding to the unbalance rather than canceling it out. This indicates that detector bias polarity should be reversed to obtain proper sensing, which can be accomplished by switching the bias supply test control from minus to plus, or vice versa.

11-1732BD. OPERATING CONTROLS AND INDICATORS. Table 11-13B provides a list of the typical operating controls and indicators which may be provided with the equipment of a typical secondary standard ir calibration method.

Table 11-13B. Typical Operating Control Chart for IR Test Equipment

EQUIPMENT	CONTROL OR INDICATOR	FUNCTION
Power and Bias Supply	Power Switch	Supplies power to power supply and ir source comparator
	Bandpass Switch	Selects a number of frequencies or disconnects filter
	Buckout and Test Adjustment	Provides coarse, fine and balance adjustments. The coarse adjustment provides an off position and scales of control in the microvolt and millivolt range
	Output Terminals	Provides the filtered and rectified outputs from the amplifier

Table 11-13B. Typical Operating Control Chart for IR Test Equipment (Cont)

EQUIPMENT	CONTROL OR INDICATOR	FUNCTION
	Bias Supply Switch	Provides off, operate, and test positions. The operate and test positions have both a minus (–) and plus (+) position for each
	Bias Supply Meter	Indicates output of battery supply voltage
IR Source Comparator	Gain Selector	Provides non-linear gain for amplifier in increments to 100
	Signal Test Points	Provides test points for measuring signal immediately after preamplifier stage of amplifier
Temperature Controller	Power Switch	Provides power to controller
	Temperature Set Dial	Provides a direct reading dial to permit setting the operating temperature of the source with the aid of a calibration chart of dial divisions versus temperature, e.g., dial setting of 359 increments equals 241 degrees C
	Range Selector Switch	Provides selection of ranges which adjusts the load to permit linear operation of the thyratron over the whole temperature scale
	Output Voltage Meter	Indicates level of power being applied to the ir source
	Warning Light	Indicates, when lit, that maximum power is applied to the source

Table 11-13B. Typical Operating Control Chart for IR Test Equipment (Cont)

EQUIPMENT	CONTROL OR INDICATOR	FUNCTION
Recorder	Power Switch	Energizes chart drive
	Zero Set Adjustment	Provides zero adjustment for recorder pen with screwdriver adjustment

11-1732BE. APERTURE AND FIELD STOPS. When the equipment is set up to operate with a precision aperture at the field image plane, larger apertures should be installed at positions 1 and 2 to serve as limits and to prevent flooding the comparator with excessive radiation. Recommended values are listed in table 11-13C.

Table 11-13C. Recommended Apertures for Positions 1 and 2 On IR Source Comparator

PRECISION APERTURE AT FIELD IMAGE	APERTURES 1 AND 2
0.0087 in.	0.030 in.
0.015 in.	0.040 in.
0.030 in.	0.090 in.
0.040 in.	0.090 in.
0.090 in.	0.200 in.
0.200 in.	0.394 in.
0.394 in.	0.500 in.
0.500 in.	None

11-1732BF. When using the equipment with precision apertures at position 1 and/or 2, a field stop slightly larger than the aperture should be used. Recommended combinations are listed in table 11-13D.

Table 11-13D. Recommended IR Test Instrument Apertures

TEST INSTRUMENT APERTURE	PRECISION APERTURE AT FIELD IMAGE
0.0087 in.	0.018 in.
0.015 in.	0.030 in.
0.030 in.	0.040 in.
0.040 in.	0.052 in.
0.090 in.	0.099 in.
0.200 in.	0.213 in.
0.394 in.	0.422 in.
0.500 in.	0.531 in.

11-1732BG. The optical equipment is aligned to permit use of the smaller field stops for the small-diameter apertures. However, the alignment of sources with integral apertures is extremely critical when small field stops are used. The larger sizes are recommended, and the smaller ones may be used when aperture plate radiance is severe.

11-1732BH. RADIATION STOPS. The sensitive infrared detectors employed in this type of equipment are susceptible to damage if excessive radiation is focused on them during operation. In particular, the thermistor bolometer should not have more than 10 milliwatts imaged on it. Radiation stops should be used to limit radiation energy. One type of radiation stop has a keyhole-shaped center obscuration to limit the field for test of the point sources used in certain missile check-outs. Another stop has six holes to limit the total energy that can fall on the detector and is intended for use with large apertures and/or high operating temperatures. An approximate relationship to determine maximum permissible aperture for a particular operating temperature may be found in the following formula:

$$\text{Aperture Diameter (in.)} = 0.9\sqrt{\frac{1}{W_T/\text{in}^2}}$$

Table 11-13E. Maximum Recommended Operating Temperatures for IR Heat Source

APERTURE DIAMETER	MAXIMUM TEMPERATURE
.200 in.	550 degrees C
.394 in. (1 cm)	300 degrees C
.500 in.	250 degrees C

where W_T/in^2 is the total radiation into a hemisphere for an emitting area at the operating temperature in degrees K. Table 11-13E shows the maximum recommended operating temperatures for various apertures. Operation at higher temperatures should be with the radiation stop.

11-1732BI. An attenuating mask should be used when operating at power levels above 10 mw or the detector will be damaged.

11-1732BJ. REFERENCE SOURCE AND CONTROLLER. The controlled ir source and temperature controller operate independently from the rest of the comparator equipment and may be used for other applications. The source and controller together form a closed loop control arrangement permitting an operator to set a desired control temperature on a temperature set dial. The control circuit will then heat the source and maintain it at the set point.

11-1732BK. The operation of the source-controller combination is relatively simple. A platinum resistance thermometer is mounted in the source core in close proximity to the heater winding. This element has a linear temperature-resistance characteristic that is precisely known. As the source is heated, the resistance of the platinum element increases in a predictable and repeatable manner. The resistance thermometer is connected to form one arm of an ac bridge circuit in the temperature controller. Two other arms of this bridge are fixed precision resistances. The fourth arm includes a ten-turn precision potentiometer which is attached to a temperature set dial.

11-1732BL. In operation, the desired temperature is used to obtain a dial setting from a temperature-versus-dial division calibration chart similar to figure 11-259C. The dial is set for the desired temperature, thus determining the value of the bridge resistance arm. The bridge will be initially unbalanced, since the source and resistance thermometer are cool. An ac voltage results from the bridge unbalance, and this forms an input to a magnetic thyratron control circuit (mtc) consisting of a transistorized preamplifier and a magnetic amplifier in a hermetically sealed unit. This unit controls the phase of a trigger pulse that fires the thyratron. Large inputs cause firing of the thyratron early in each cycle, and a maximum amount of power is fed into the load.

11-1732BM. The load in this case is a resistance element wound around the source

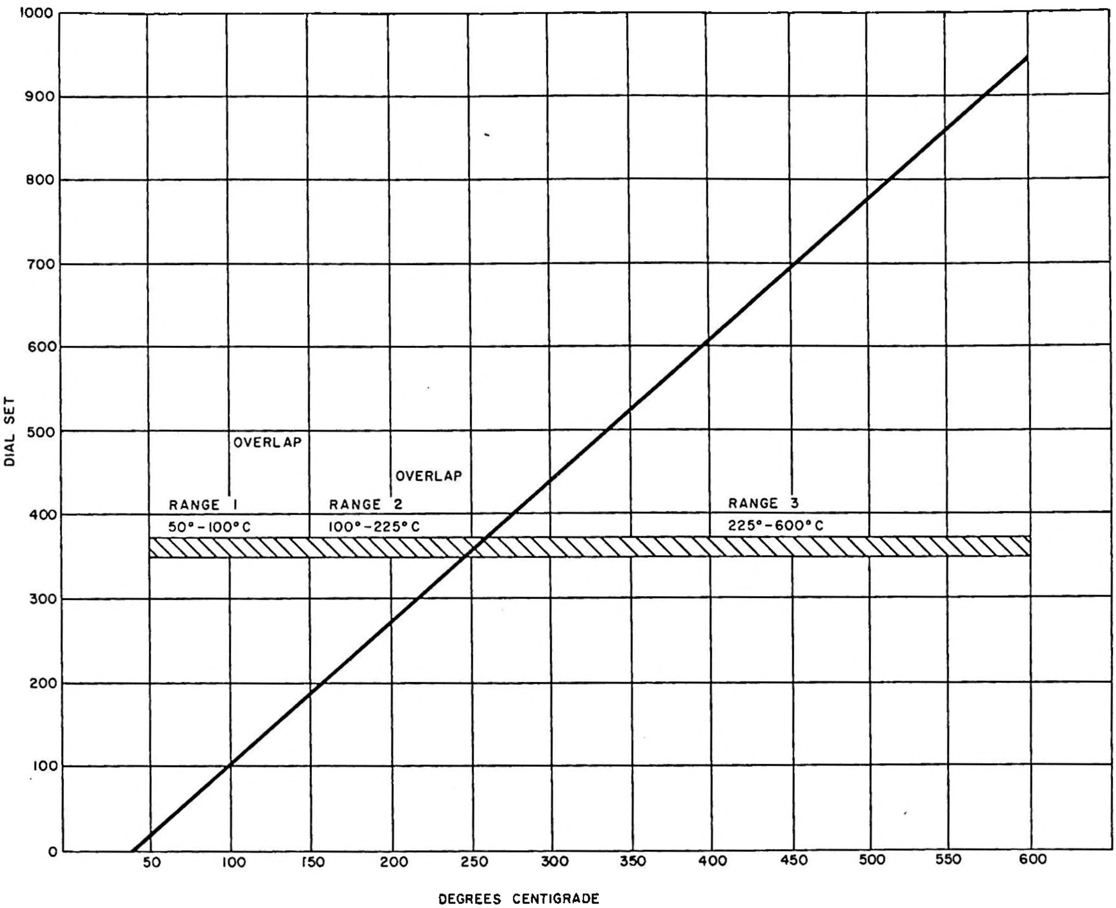


Figure 11-259C. Typical Temperature Set Control Calibration Chart

core. As the source temperature approaches the desired level, the value of the resistance thermometer approaches that of the control arm of the ac resistance bridge, and the approaching balance condition reduces the ac unbalance, thus lowering the input voltage to the unit. As the input is lowered, the phase of the trigger voltage shifts to a later portion of the cycle, the thyratron conducts for a shorter period, and less power is fed to the load. When the core reaches the set temperature, the bridge will be balanced, and the power fed into the load will be the minimum amount required to maintain the source temperature and make up for radiated and conducted losses.

11-1732BN. The extended range of operation from 50 degrees C to 600 degrees C, as shown in figure 259C, results in a large dynamic range of quiescent operating power required by the source. Auxiliary loading can be provided for three temperature ranges of operation in order to keep the thyratron operating on a linear portion of its characteristic. The range selector switch setting accomplishes this load matching. It does not change the actual operating range of the source, which is established by the range of adjustment of the temperature set and range selector will result in unstable operation of the source, with poor temperature control.

11-1732BO. The output voltage panel meter indicates the level of power being supplied to the load. During the source heating cycle, the meter will read approximately 45 volts. At low temperatures, during initial setting, the source will overshoot and the meter will return to zero, indicating that the power is cut off while the source temperature stabilizes.

11-1732BP. Under normal operation, the output voltage meter will stabilize at some level between approximately 8 and 35 volts (depending on the temperature setting) when the source reaches operating temperature. Radiation measurements should not be made with the source until it has been stabilized for about one minute.

11-1732BQ. A typical ir source is designed to operate from 50 degrees C to 600 degrees C. The radiating element is normally a pure silver core with a 15-degree conical cavity. Silver has the highest thermal conductivity of any material, assuring uniform temperature along the walls of the cavity. The core surface is normally coated with a refractory material to increase its emissivity to approximately unity.

11-1732BR. The conical cavity opening should be 0.75 inch for a recommended maximum aperture diameter of 0.5 inch. Radiation distribution will normally be within 2 percent of the ideal cosine characteristic over a 20-degree field for the 0.5 inch aperture.

11-1732BS. OPTICAL EQUIPMENT ALIGNMENT. The optical equipment is divided into two sections. Alignment of the first section involves adjustment of the various mirrors, including the chopper assembly, in the photometer section to provide a proper in-focus image of aperture 1 and aperture 2 within the field stop. Optical alignment of the detector section involves adjustment of the ellipsoid mirror to focus the field image on the detector.

11-1732BT. Alignment of the photometer section requires special equipment consisting of a small 50-X microscope having a prism or mirror mounted on the objective to fold the optical axis by 90 degrees, and with provisions for fine focusing of the microscope. This equipment should be supported on a suitable stand and positioned vertically to observe the field image in the comparator from the detector side. Perform the alignment check as follows:

a. Install properly matched precision apertures and field stops. The field stop should be approximately 0.010 inch larger than the aperture.

b. Set up the microscope to view the field image from the detector side.

c. Rotate the mirror chopper by hand to observe apertures 1 and 2 alternately. They should be centered in the field stop, and they should be well focused.

11-1732BU. The toroid mirror is the image-forming element in the photometer section. Toroids will not make perfect images, and the resulting sharpness of the image of apertures 1 and 2 will be slightly less than that of the field stop. Positioning of the toroid mirror is very critical for proper imaging with minimum distortion.

11-1732BV. When housing the photometer, make sure that the images of aperture 1 and aperture 2 are sharply focused. A slight flare or smearing of the image in one direction may be reduced by masking a portion of the toroid mirror. Focus should be adjusted with any flare minimized in this way.

11-1732BW. The first step is to adjust the focus of aperture 1 as it appears in the field image plane. This is accomplished by adjusting the position of the toroid mirror shown in figure 11-259B. This mirror adjustment can be performed as follows:

a. Remove the ir source comparator base plate.

b. Illuminate aperture 1.

c. Rotate the chopper mirror out of the optical path.

d. Sufficiently loosen the toroid mirror mount to permit lateral motion of the mount.

e. Adjust the lateral position of the toroid mirror mount to obtain optimum focus of aperture 1 in the field plane.

f. Tighten the toroid mirror mount.

g. Check the image focus to make sure that the tightening of the mount did not shift the focus.

h. Illuminate aperture 2 and rotate the chopper to observe this aperture at the field image plane.

i. Adjust the position of the chopper mirror on the shaft to optimize the focus of aperture 2 at the field image plane.

11-1732BX. Adjustments of the chopper mirror for focus should be done carefully to prevent rotational motion about the shaft which will result in improper phasing.

11-1732BY. The images of aperture 1 and aperture 2 can be centered in the field stop as follows:

a. With the chopper mirror in the optical path, adjust the plane mirror for tilt and rotation to center the image of aperture 2. The mirror tilt should be adjustable with a set screw at the top of the mirror mount. Rotation of the mirror is adjustable by loosening a set screw on the front of the mount just below the mirror assembly.

b. With the aperture 2 image centered, rotate the chopper mirror out of the optical path.

c. Adjust mirrors RM1 and/or the chopper mirror for tilt and swing to center the image of aperture 1 in the field image plane.

d. Tighten all mirror mounts.

e. Rotate the chopper mirror to check the superposition of aperture 1 and aperture 2 images.

11-1732BZ. Alignment of the detector section including the ellipsoid mirror requires no special optical equipment. The procedure is as follows:

a. With the largest aperture and field stops in place, mount a test instrument at position 1 (do not exceed 250 degrees C source temperature).

b. Adjust the temperature of the test instrument and the gain setting of the comparator to provide an on-scale deflection of approximately five divisions on the recorder pen.

c. Loosen the four set screws locking the mirror shaft on the ellipsoid mirror.

d. Adjust the knurled knob at the rear of the mount to change the focus of the ellipsoid. This will be indicated by a maximum deflection of the recorder pen. Adjust the mirror focus position and the gain selector on the comparator, as required, to obtain maximum on-scale deflection.

e. Lock the four set screws.

f. Adjust the position of the ellipsoid mirror with four socket head screws at the rear of the ellipsoid mirror mount. These should be adjusted to obtain maximum on-scale deflection of the recorder and that should be reasonably tight to prevent loosening or changing of position by vibration.

11-1732CA. The prevision apertures normally supplied with the equipment are carefully assembled to assure concentricity of the apertures with respect to its mounting block. The apertures are aligned on a pilot diameter of the mounting block and should be seated against the face of the mounting block to provide an in-focus image. The precise location requirements have established tight mechanical tolerances with the result that close fits are necessary between the apertures and the mounting block. Extreme care should be exercised in mounting and in changing these apertures to prevent damage to the small aperture disk.

11-1732CB. CALIBRATION PROCEDURE. The following paragraphs discuss a typical procedure for the calibration of an ir heat source. This procedure is typical and the equipment used is typical; therefore, all of the equipments, controls, and components mentioned are assumed to be available to you.

11-1732CC. After performing all of the initial procedures outlined in the individual equipment manuals, set up the equipment for calibration of an ir heat source. The following procedure is provided with the assumption that the equipment is connected and in good operating condition. Proceed as follows:

a. Mount the controlled ir reference source on the reference source mounting bracket of the comparator shown in figure 11-259D so that it is centered, but not touching aperture 1. Use the vertical and transverse adjustment controls if necessary.

b. Turn on the reference source temperature controller and adjust the temperature set control on the temperature controller so that the reference source temperature will be close to the desired temperature of the test source.

c. While allowing the reference source to warm up, self check the power and bias supply by adjusting the mechanical zero.

d. Turn the bias supply switch to the plus (+) position. Indication should be within tolerance of the bias supply specification.

e. Turn the bias supply switch to the minus (−) position. Indication should be within tolerance of the bias supply specification.

f. Turn the bias supply switch to the off position.

g. After warming up the reference source for the specified time, set the power

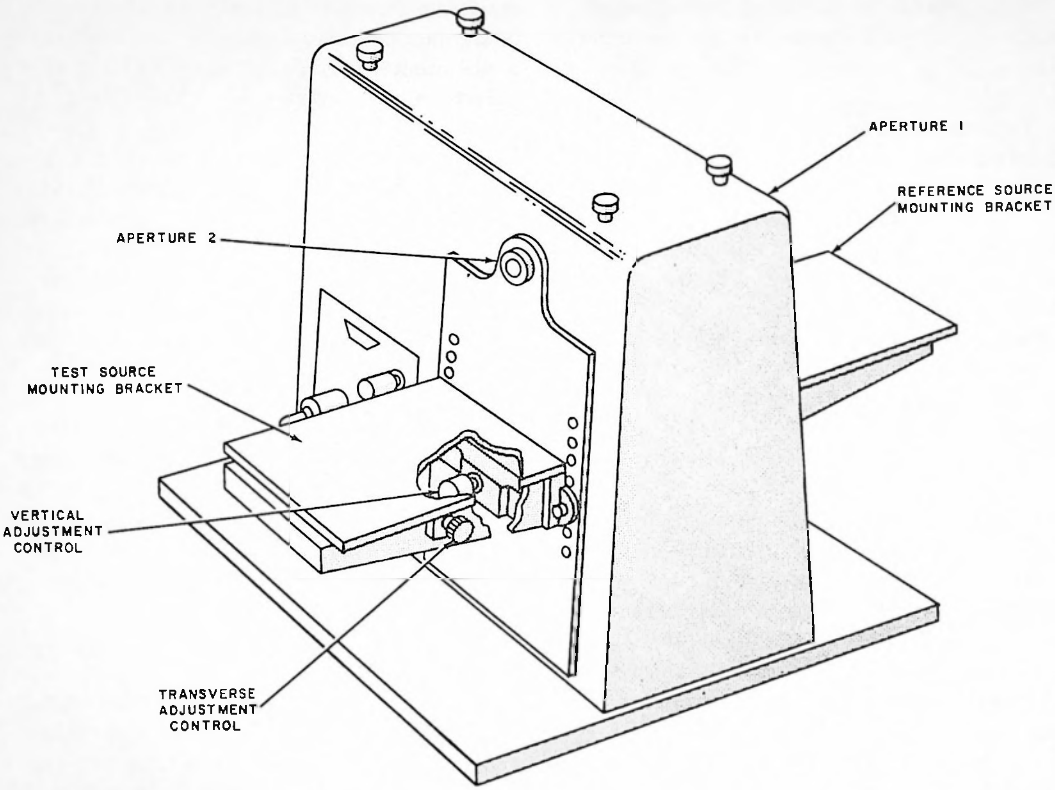


Figure 11-259D. IR Source Comparator

and bias supply to an appropriate position and check that the buckout and test signal coarse control is in the off position.

h. Adjust the recorder stylus on the graph to zero by adjusting the zero set control.

i. Set the power and bias supply bandpass control to the desired position.

j. Set the power and bias supply switches to the operate position.

k. Advance the ir source comparator gain selector control to obtain a maximum on-scale indication of the recorder stylus.

l. Readjust the coarse, the fine, and the balance controls of the buckout and test signal to return the stylus to the zero indication on the graph.

m. Advance the gain selector control one clockwise position and record.

11-1732CD. Balance the optical paths as follows:

a. Repeat step c of the preceding paragraph.

b. Return the gain selector to position 1.

c. Return the power switch to the standby position.

d. Remove the mask from ir source comparator aperture 2.

e. Remove the reference source from mounting bracket 1 and place it on mounting bracket 2 and adjust the vertical and transverse adjustment controls so that they are flush, but not touching and centered about aperture 2.

f. Adjust the vertical and the transverse controls so that the stylus on the graph indicates maximum peak deflection to the left.

g. Mask aperture 1 with a suitable opaque material.

h. Set the power and bias supply switches to an appropriate position.

i. Advance the gain selector control of the ir source comparator so that the stylus indicates the reading that you recorded in step m of the preceding paragraph and check that the recorder stylus remains at zero.

j. Return the power switch to the standby position.

11-1732CE. Perform an optical balance check of the ir source comparator as follows:

a. Remove the mask from aperture 1 of the ir source comparator.

b. Remove the reference source from the adjustable mounting bracket and reinstall it on the reference source mounting bracket adjusting the vertical and transverse adjustment controls so that it is flush, but not touching and centered about aperture 1.

c. Mask aperture 2 with a suitable opaque material.

d. Set the power and bias supply switches to the operate position.

e. Advance the ir source comparator gain selector control to obtain the recorded reading in step m of paragraph 11-1732CC and check that the recorder stylus remains on zero. If the recorder stylus does not remain on zero, adjust the coarse and balance controls of the buckout and test signal until no stylus deflection results and repeat all steps in paragraph 11-1732CD. Next, repeat steps a through e of paragraph 11-1732CE.

f. Return the gain selector control to position 1.

g. Set the power and bias supply buckout and test signal coarse control to off; the bias supply switch to off; and the power switch to standby.

h. Turn off the recorder and remove the mask from aperture 2.

11-1732CF. Table 11-13F provides a typical temperature-to-millivolt conversion chart for an iron vs constantan thermocouple. Use the applicable conversion chart for the type of thermocouple you use, for example, iron, constantan, copper, etc. To determine the temperature of the standard reference source, proceed as follows:

a. Insert the hot junction of the thermocouple fully into the receptacle at the rear of the reference source. Most reference sources will have a lock nut on the rear. If the reference source that you are using has a lock nut, lock the thermocouple into the rear of the reference source.

b. With the reference (cold) junction of the thermocouple placed in an ice bath, connect the thermocouple leads to a potentiometer and verify that the potentiometer is standardized with respect to the standard cell.

c. Operate the potentiometer millivolt dial setting to approximately the indication that you would have if the reference source temperature was the desired temperature. This setting can be found by referring to table 11-13F. For instance, if you want the reference source to be 360 degrees C, the millivolt indication on the potentiometer should be 19.64 mv. Therefore, operate the millivolt dials to a setting of 19.64 mv.

d. Depress the sensitivity key on the potentiometer. If the galvanometer deflection is to the left-hand side of the scale, the temperature of the reference source must be increased; right-hand deflection indicates that the temperature must be reduced.

e. Vary the temperature set control of the reference source controller (allowing time for temperature stabilization) until no deflection on the galvanometer occurs when the sensitivity key is depressed.

f. If the galvanometer indicates no deflection, and the millivolt dial setting is 19.64 mv, the temperature of the reference source is 360 degrees C.

Table 11-13F. Iron Versus Constantan Temperature Conversion Chart

DEGREES C	MILLIVOLTS WITH REFERENCE JUNCTION OF ZERO DEGREES C									
-190	-7.66	-7.69	-7.71	-7.73	-7.76	-7.78				
-180	-7.40	-7.43	-7.46	-7.49	-7.51	-7.54	-7.56	-7.59	-7.61	-7.64
-170	-7.12	-7.15	-7.18	-7.21	-7.24	-7.27	-7.30	-7.32	-7.35	-7.38
-160	-6.82	-6.85	-6.88	-6.91	-6.94	-6.97	-7.00	-7.03	-7.06	-7.09
-150	-6.50	-6.53	-6.56	-6.60	-6.63	-6.66	-6.69	-6.72	-6.76	-6.79
-140	-6.16	-6.19	-6.22	-6.26	-6.29	-6.33	-6.36	-6.40	-6.43	-6.46
-130	-5.80	-5.84	-5.87	-5.91	-5.94	-5.98	-6.01	-6.05	-6.08	-6.12
-120	-5.42	-5.46	-5.50	-5.54	-5.58	-5.61	-5.65	-5.69	-5.72	-5.76
-110	-5.03	-5.07	-5.11	-5.15	-5.19	-5.23	-5.27	-5.31	-5.35	-5.38
-100	-4.63	-4.67	-4.71	-4.75	-4.79	-4.83	-4.87	-4.91	-4.95	-4.99
-90	-4.21	-4.25	-4.30	-4.34	-4.38	-4.42	-4.46	-4.50	-4.55	-4.59
-80	-3.78	-3.82	-3.87	-3.91	-3.96	-4.00	-4.04	-4.08	-4.13	-4.17
-70	-3.34	-3.38	-3.43	-3.47	-3.52	-3.56	-3.60	-3.65	-3.69	-3.74
-60	-2.89	-2.94	-2.98	-3.03	-3.07	-3.12	-3.16	-3.21	-3.25	-3.30

Table 11-13F. Iron Versus Constantan Temperature Conversion Chart (Cont)

DEGREES C	MILLIVOLTS WITH REFERENCE JUNCTION OF ZERO DEGREES C									
-50	-2.43	-2.48	-2.52	-2.57	-2.62	-2.66	-2.71	-2.75	-2.80	-2.84
-40	-1.96	-2.01	-2.06	-2.10	-2.15	-2.20	-2.24	-2.29	-2.34	-2.38
-30	-1.48	-1.53	-1.58	-1.63	-1.67	-1.72	-1.77	-1.82	-1.87	-1.91
-20	-1.00	-1.04	-1.09	-1.14	-1.19	-1.24	-1.29	-1.34	-1.39	-1.43
-10	-0.50	-0.55	-0.60	-0.65	-0.70	-0.75	-0.80	-0.85	-0.90	-0.95
(-)0	0.00	-0.05	-0.10	-0.15	-0.20	-0.25	-0.30	-0.35	-0.40	-0.45
(+)0	0.00	0.05	0.10	0.15	0.20	0.25	0.30	0.35	0.40	0.45
10	0.50	0.56	0.61	0.66	0.71	0.76	0.81	0.86	0.91	0.97
20	1.02	1.07	1.12	1.17	1.22	1.28	1.33	1.38	1.43	1.48
30	1.54	1.59	1.64	1.69	1.74	1.80	1.85	1.90	1.95	2.00
40	2.06	2.11	2.16	2.22	2.27	2.32	2.37	2.42	2.48	2.53
50	2.58	2.64	2.69	2.74	2.80	2.85	2.90	2.96	3.01	3.06
60	3.11	3.17	3.22	3.27	3.33	3.38	3.43	3.49	3.54	3.60
70	3.65	3.70	3.76	3.81	3.86	3.92	3.97	4.02	4.08	4.13
80	4.19	4.24	4.29	4.35	4.40	4.46	4.51	4.56	4.62	4.67
90	4.73	4.78	4.83	4.89	4.94	5.00	5.05	5.10	5.16	5.21
100	5.27	5.32	5.38	5.43	5.48	5.54	5.59	5.65	5.70	5.76
110	5.81	5.86	5.92	5.97	6.03	6.08	6.14	6.19	6.25	6.30
120	6.36	6.41	6.47	6.52	6.58	6.63	6.68	6.74	6.79	6.85
130	6.90	6.96	7.01	7.07	7.12	7.18	7.23	7.29	7.34	7.40
140	7.45	7.51	7.56	7.62	7.67	7.73	7.78	7.84	7.89	7.95
150	8.00	8.06	8.12	8.17	8.23	8.28	8.34	8.39	8.45	8.50
160	8.56	8.61	8.67	8.72	8.78	8.84	8.89	8.95	9.00	9.06
170	9.11	9.17	9.22	9.28	9.33	9.39	9.44	9.50	9.56	9.61
180	9.67	9.72	9.78	9.83	9.89	9.95	10.00	10.06	10.11	10.17
190	10.22	10.28	10.34	10.39	10.45	10.50	10.56	10.61	10.67	10.72
200	10.78	10.84	10.89	10.95	11.00	11.06	11.12	11.17	11.23	11.28
210	11.34	11.39	11.45	11.50	11.56	11.62	11.67	11.73	11.78	11.84
220	11.89	11.95	12.00	12.06	12.12	12.17	12.23	12.28	12.34	12.39
230	12.45	12.50	12.56	12.62	12.67	12.73	12.78	12.84	12.89	12.95
240	13.01	13.06	13.12	13.17	13.23	13.28	13.34	13.40	13.45	13.51

Table 11-13F. Iron Versus Constantan Temperature Conversion Chart (Cont)

DEGREES C	MILLIVOLTS WITH REFERENCE JUNCTION OF ZERO DEGREES C									
250	13.56	13.62	13.67	13.73	13.78	13.84	13.89	13.95	14.00	14.06
260	14.12	14.17	14.23	14.28	14.34	14.39	14.45	14.50	14.56	14.61
270	14.67	14.72	14.78	14.83	14.89	14.94	15.00	15.06	15.11	15.17
280	15.22	15.28	15.33	15.39	15.44	15.50	15.55	15.61	15.66	15.72
290	15.77	15.83	15.88	15.94	16.00	16.05	16.11	16.16	16.22	16.27
300	16.33	16.38	16.44	16.49	16.55	16.60	16.66	16.71	16.77	16.82
310	16.88	16.93	16.99	17.04	17.10	17.15	17.21	17.26	17.32	17.37
320	17.43	17.48	17.54	17.60	17.65	17.71	17.76	17.82	17.87	17.93
330	17.98	18.04	18.09	18.15	18.20	18.26	18.32	18.37	18.43	18.48
340	18.54	18.59	18.65	18.70	18.76	18.81	18.87	18.92	18.98	19.03
350	19.09	19.14	19.20	19.26	19.31	19.37	19.42	19.48	19.53	19.59
360	19.64	19.70	19.75	19.81	19.86	19.92	19.97	20.03	20.08	20.14
370	20.20	20.25	20.31	20.36	20.42	20.47	20.53	20.58	20.64	20.69
380	20.75	20.80	20.86	20.91	20.97	21.02	21.08	21.13	21.19	21.24
390	21.30	21.35	21.41	21.46	21.52	21.57	21.63	21.68	21.74	21.79
400	21.85	21.90	21.96	22.02	22.07	22.13	22.18	22.24	22.29	22.35
410	22.40	22.46	22.51	22.57	22.62	22.68	22.73	22.79	22.84	22.90
420	22.95	23.01	23.06	23.12	23.17	23.23	23.28	23.34	23.39	23.45
430	23.50	23.56	23.61	23.67	23.72	23.78	23.83	23.89	23.94	24.00
440	24.06	24.11	24.17	24.22	24.28	24.33	24.39	24.44	24.50	24.55
450	24.61	24.66	24.72	24.77	24.83	24.88	24.94	25.00	25.05	25.11
460	25.16	25.22	25.27	25.33	25.38	25.44	25.49	25.55	25.60	25.66
470	25.72	25.77	25.83	25.88	25.94	25.99	26.05	26.10	26.16	26.22
480	26.27	26.33	26.38	26.44	26.49	26.55	26.61	26.66	26.72	26.77
490	26.83	26.89	26.94	27.00	27.05	27.11	27.17	27.22	27.28	27.33
500	27.39	27.45	27.50	27.56	27.61	27.67	27.73	27.78	27.84	27.90
510	27.95	28.01	28.07	28.12	28.18	28.23	28.29	28.35	28.40	28.46
520	28.52	28.57	28.63	28.69	28.74	28.80	28.86	28.91	28.97	29.02
530	29.08	29.14	29.20	29.25	29.31	29.37	29.42	29.48	29.54	29.59
540	29.65	29.71	29.76	29.82	29.88	29.94	29.99	30.05	30.11	30.16
550	30.22	30.28	30.34	30.39	30.45	30.51	30.57	30.62	30.68	30.74
560	30.80	30.85	30.91	30.97	31.02	31.08	31.14	31.20	31.26	31.31
570	31.37	31.43	31.49	31.54	31.60	31.66	31.72	31.78	31.83	31.89
580	31.95	32.01	32.06	32.12	32.18	32.24	32.30	32.36	32.41	32.47
590	32.53	32.59	32.65	32.71	32.76	32.82	32.88	32.94	33.00	33.06

Table 11-13F. Iron Versus Constantan Temperature Conversion Chart (Cont)

DEGREES C	MILLIVOLTS WITH REFERENCE JUNCTION OF ZERO DEGREES C									
600	33.11	33.17	33.23	33.29	33.35	33.41	33.46	33.52	33.58	33.64
610	33.70	33.76	33.82	33.88	33.94	33.99	34.05	34.11	34.17	34.23
620	34.29	34.35	34.41	34.47	34.53	34.58	34.64	34.70	34.76	34.82
630	34.88	34.94	35.00	35.06	35.12	35.18	35.24	35.30	35.36	35.42
640	35.48	35.54	35.60	35.66	35.72	35.78	35.84	35.90	35.96	36.02
650	36.08	36.14	36.20	36.26	36.32	36.38	36.44	36.50	36.56	36.62
660	36.69	36.75	36.81	36.87	36.93	36.99	37.05	37.11	37.18	37.24
670	37.30	37.36	37.42	37.48	37.54	37.60	37.66	37.73	37.79	37.85
680	37.91	37.97	38.04	38.10	38.16	38.22	38.28	38.34	38.41	38.47
690	38.53	38.59	38.66	38.72	38.78	38.84	38.90	38.97	39.03	39.09
700	39.15	39.22	39.28	39.34	39.40	39.47	39.53	39.59	39.65	39.72
710	39.78	39.84	39.91	39.97	40.03	40.10	40.16	40.22	40.28	40.35
720	40.41	40.48	40.54	40.60	40.68	40.73	40.79	40.86	40.92	40.98
730	41.05	41.11	41.17	41.24	41.30	41.36	41.43	41.49	41.56	41.62
740	41.68	41.75	41.81	41.87	41.94	42.00	42.07	42.13	42.19	42.26
750	42.32	42.38	42.45	42.51	42.58	42.64	42.70	42.77	42.83	42.90
760	42.96	43.02	43.09	43.15	43.22	43.28	43.35	43.41	43.48	43.54
770	43.60	43.67	43.73	43.80	43.86	43.92	43.99	44.05	44.12	44.18
780	44.25	44.31	44.38	44.44	44.50	44.57	44.63	44.70	44.76	44.82
790	44.89	44.95	45.02	45.08	45.15	45.21	45.28	45.34	45.40	45.47
800	45.53	45.60	45.66	45.72	45.79	45.85	45.92	45.98	46.05	46.11
810	46.18	46.24	46.30	46.37	46.43	46.50	46.56	46.62	46.69	46.75
820	46.82	46.88	46.94	47.01	47.07	47.14	47.20	47.27	47.33	47.39
830	47.46	47.52	47.58	47.65	47.71	47.78	47.84	47.90	47.97	48.03
840	48.09	48.16	48.22	48.28	48.35	48.41	48.48	48.54	48.60	48.66
850	48.73	48.79	48.85	48.92	48.98	49.04	49.10	49.17	49.23	49.29
860	49.36	49.42	49.48	49.54	49.61	49.67	49.73	49.79	49.86	49.92
870	49.98	50.04								

11-1732CG. The previous paragraphs have provided procedures which, when completed, have set up the equipment for the calibration of one or more ir heat sources. To calibrate an ir heat source, proceed as follows:

a. Install the proper precision aperture on the ir source comparator at the field image plane, in the center part of the comparator shown in figure 11-259D.

b. Remove the reference source from the reference source mounting bracket.

c. Mount the proper radiation-limiting apertures at ir source comparator apertures 1 and 2. These apertures should be of the same size. Table 11-13G provides a list of apertures to be used.

d. Adjust the buckout signal as follows:

1. Mask both ir source comparator apertures with an opaque material.
2. Turn on the recorder.
3. Set the power switch to operate, the bias supply switch to off, and the buckout and test signal coarse control to 1-10 μv.
4. Advance the ir source comparator gain selector control to the maximum position.
5. Adjust the buckout and test signal fine and balance controls until the recorder stylus indicates zero. Do not change these settings until the buckout signal is again adjusted.
6. Return the power and bias supply power switch to the standby position.
7. Turn off the recorder.
8. Return the ir source comparator gain selector control to position 1.

e. Replace the reference source on the reference mounting bracket and adjust the vertical and transverse adjustment controls so that it is centered but not touching aperture 1.

f. Connect the test ir source to the test ir source temperature controller, mount the test source on the adjustable mounting bracket and adjust the vertical and transverse adjustment controls so that the source is centered but not touching aperture 2.

g. Turn on the test source temperature controller and adjust the temperature set control to the desired temperature and permit the source to warm up for the specified time.

h. Verify that the potentiometer is standardized with respect to the standard cell.

i. Depress the sensitivity key on the potentiometer. If the galvanometer deflection is to the left-hand side of the scale, the temperature of the test ir source must be increased; right-hand deflection indicates that the temperature must be reduced.

j. Vary the temperature set control of the test source controller (allowing time for temperature stabilization) until no deflection on the galvanometer occurs when the sensitivity key is depressed.

k. Turn on the recorder.

l. Adjust the power and bias supply switches to operate.

m. Advance the ir source comparator gain selector control until the recorder stylus deflects from zero.

n. Adjust the vertical and transverse adjustment controls on the adjustable mount-

Table 11-13G. Radiation Limiting Apertures

PRECISION APERTURE AT FIELD IMAGE	APERTURES 1 AND 2
0.0087 in.	0.030 in.
0.015 in.	0.040 in.
0.030 in.	0.090 in.
0.040 in.	0.090 in.
0.090 in.	0.200 in.
0.200 in.	0.394 in.
0.394 in.	0.500 in.
0.500 in.	None

ing bracket to minimize the recorder stylus deflection.

o. If adjustment of the controls returns the stylus to zero, repeat steps m and n until the stylus deflection cannot be further reduced by adjustment of the controls.

p. Adjust the test source operating temperature to reduce the stylus deflection and allow the source temperature to stabilize.

q. Repeat steps m and p until the gain selector control is at maximum gain setting and the stylus remains on zero with minimum fluctuations. The ir source under test is now calibrated to the operating temperature of the standard reference source.

r. Return the gain selector control to position 1.

s. Turn off the bias supply switch and return the power switch to standby.

t. Turn off and remove the test ir source and controller. At this time additional ir sources may be tested. If you want to calibrate another ir source, repeat steps f through i for each ir source to be calibrated.

11-1732CH. INFRARED FIBER OPTICS.

11-1732CI. GENERAL.

11-1732CJ. The most extensive use of fiber optics is for the transmission of visible radiation. However, the glass materials of visible light fibers transmit radiation only in the frequency range of the near infrared (ir) and the materials are not suitable for wavelengths beyond 2.5 microns. Techniques have been developed for the processing of fiber materials which possess the ability to transmit a frequency range higher than the near-infrared region.

11-1732CK. CHARACTERISTICS OF FIBER OPTICS. The optical characteristics of the fibers are related to the physical properties of the fiber material used and to its index of refraction. The physical properties of greatest interest are usually the numerical aperture, the spectral response, and the transmission. Together with the feasibility of producing fibers with the proper coating, these properties determine which of the infrared fiber-optics material is to be used for a specific application.

11-1732CL. The numerical aperture (na) is a measure of the light-gathering ability of an optical device. The range of numerical apertures available in fiber optics is limited only by the materials from which the fibers can be made. Fiber optics bundles made of selected pairs of glasses may use any value of numerical aperture to 1.2. Infrared materials such as arsenic trisulfide glass have higher refractive indices, and it is therefore possible to obtain large numerical apertures for the same coating-to-core refractive index ratio.

11-1732CM. Energy travels from the entrance end of the fiber to the exit end by means of total internal reflection. If the fibers were perfect, the entrance cone would be equal to the exit cone. However, since the fiber walls are not perfectly parallel and smooth, the output radiation pattern exceeds the pattern of the entrance cone.

11-1732CN. Factors which affect the overall transmission of an infrared fiber-optics bundle are internal transmittance, dead areas caused by packing density, broken fibers, the type of cladding material used, the efficiency of the antireflection coating used, and the length of the bundle. Infrared-transmitting fiber-optics bundles can be either rigid or flexible. One of the most valuable applications for rigid bundles is its use in connection with image conversion. For example, it is often desirable to use a slit pattern in the primary image plane and to use a circular pattern on a detector cell in the secondary image plane. This application is easily achieved using a fiber-optics converter. Another application is found in converting from a line to a raster for a television display. This enables a wide-angle, high-resolution image to be broken into segments so that the detecting device will not limit the overall resolution. Flexible fiber bundles can be used in infrared equipment by transmitting the energy from a gimbaled optical arrangement to a stationary detector and a cooling arrangement. Other uses might be found in the form of infrared fiber scopes for medical diagnosis where internal temperature differences can be detected and imaged. In any application, one of the most important considerations is overall transmission. However, other parameters such as cooling of the bundle and methods of efficient coupling of the fiber bundle to the detector or imaging device must be determined in order to use an infrared bundle in the best manner.

11-1732CO. PROPERTIES OF IR FIBER OPTICS. For all practical purposes, the infrared region covers a broad spectral span from approximately 0.8 micron to 100 microns. As a result, most infrared materials transmit selectively in the infrared and do not pass radiation with equal efficiency at all wavelengths over the complete infrared spectrum. Some materials are difficult to fabricate into optical elements and some are even toxic, requiring special treatment and careful handling. Another difficulty is that many optically-desirable intervals are not easily drawn into a fiber. The material that has thus far proven most suitable and most easily drawn into a fiber is arsenic trisulfide glass. Short lengths of this material, when clad with a modified arsenic sulfide, have exhibited spectral response to a wavelength of 12 microns.

11-1732CP. The fiber is fabricated from a melt of arsenic trisulfide glass. The liquid glass, along with the cladding material, is drawn through a die. Extreme caution must be used when heating the arsenic trisulfide glass since it emits toxic fumes at a critical temperature of 300 degrees F.

11-1732CQ. The method used for fabricating the fibers into a coherent bundle involves a process which draws the fibers around a drum in complete loops. These loops are then fixed in place by an epoxy before cutting and polishing to keep the fibers at each end in the same relative position, and to maintain coherency between the input and output ends of the finished bundle. Since arsenic trisulfide does not transmit in the visible region, locating individual fibers to check bundle coherency is a problem. Minimum bundle fiber lengths are limited by drum size. The fabrication of the infrared fibers into useful bundles invariably results in mechanical defects such as broken fibers, nonuniformity of fibers in the cross-sectional dimension along the fiber length, gaps between the fibers, and fiber displacement. Broken fibers reduce the total energy transmission, lower the resolution capabilities of the bundle in these dead areas, and contribute to the background noise of a transmitted image.

11-1732CR. Breakage of fibers up to 1 or 2 percent of the total number of fibers in a bundle is common in bundles used in the visible spectrum. Existing state-of-the-art infrared fibers can experience up to 5 percent breakage. However, flexible infrared bundles have been made and flexed as many as 1.5 million cycles without their sustaining any apparent damage. Arsenic trisulfide fibers have been fabricated with a cladding thickness that is about 10 percent of the core diameter. In addition to a variation of fiber diameter between fibers of a bundle, a variation occurs in the cross-sectional area along the length of the fiber. This variation contributes to a spreading of the transmitted energy at the output end. Gaps between fibers and fiber displacement result in image-resolution deterioration. The existing state-of-the-art for infrared fiber optics indicates a fiber displacement tolerance of $\pm 1/2$ of the fiber diameter. Packing density can be increased by fusing the ends of fibers together since distortion of the circular cross-section does not significantly affect optical properties of the fibers.

11-1732CS. IR DETECTOR WITH LIGHT PIPE.

11-1732CT. The transmission of radiation in fibers by multiple internal reflections is based on a simple physical principle: the total reflection of radiant energy occurring at a boundary if the index of refraction of the medium in which the radiation exists is greater than the index of refraction of the medium on the other side of the boundary (that which the radiation would normally pass into if it were not reflected back internally). Thus, if the radiation is in a medium of index (fiber-core material), and the index of the material on the other side of the boundary is a fiber-cladding material, the criterion for internal reflection is such that any entering radiation incident on the internal fiber walls at an angle greater than the critical angle is continually reflected. The fiber thus acts as a smooth, transparent cylinder, transmitting the radiation by multiple internal reflections. The diameter of the fiber core must be larger than the longest wavelength to be transmitted and no substantial change will occur in the behavior of this phenomenon until the diameter of the fiber core becomes comparable to the wavelength of light, or about 5 microns.

11-1732CU. If many such fibers are gathered together into an orderly array, an image placed at one end will be transmitted to the other end, with the fibers breaking up the image and transmitting each component separately. Transmission is improved by cladding each fiber with a coating that has an index of refraction which is lower than that of the fiber material. The cladding prevents leakage or cross-talk from one fiber to another.

11-1732CV. A method which solves the problem of an obstructed target in an rf field uses an infrared detector and a fiber-optic light pipe to measure the transistor-header temperature during die-bonding operations. An infrared detector is used as the sensing element because of its advantages over optical pyrometers, thermocouples, and resistance thermometers. Optical pyrometers have a low useful limit of approximately 650 degrees C and are inherently slow. An infrared detector, however, has a lower limit, slightly above room temperature, and has a rapid response time. Thermocouples and resistance thermometers are affected by rf pickup while an infrared detector is not.

11-1732CW. Figure 11-259E shows the arrangement of equipment for measuring the transistor-header temperature. A 0.050-inch diameter sapphire rod with a 0.025-inch diameter die end is used for mechanical considerations and the rod acts as both a hold-down probe and a light pipe. The reduction of the rod diameter at the die end has no effect on its operation as a light pipe. The header in this typical arrangement is gold-plated and the germanium die is approximately 0.016 inch x 0.016 inch x 0.004 inch in size. The sapphire rod is placed in contact with the die, with the field of view taking in the entire germanium surface and a part of the header in about a 1:1 ratio.

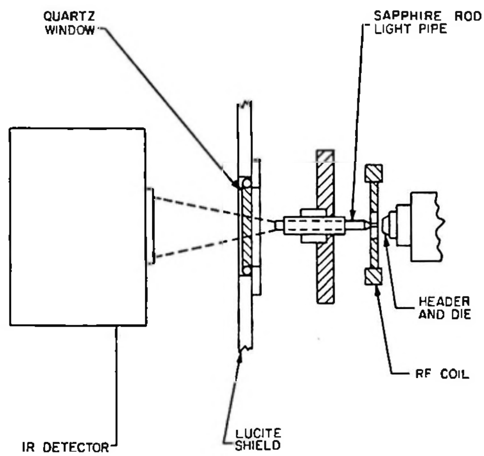


Figure 11-259E. IR Detector and Light-Pipe Measurement Method

11-1732CX. Calibration of the ir detector and sapphire light-pipe combination may be accomplished with an iron-constantan thermocouple. By placing the thermocouple wires in contact with the header close to the germanium die, the thermocouple temperature is found to be the average of the two temperatures at the contact points. This method of measuring surface temperature with an open thermocouple is more effective than the spot welding of the junction to the header.

11-1732CY. A heater which controls both the temperature and the heating rates is

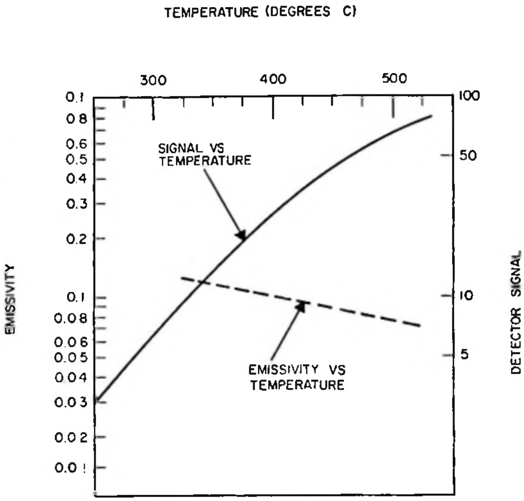


Figure 11-259F. Typical Light-Pulse Calibration Curves for IR Detector and Light-Pipe Arrangement

used in the measurement of header surface temperature, and two recorders are connected for simultaneous recording of thermocouple and detector outputs. Typical calibration curves, shown in figure 11-259F, illustrate the detector signal versus header temperature, and the effective emissivity versus temperature. The transmission factor of the light pipe is not necessary unless the target is to be viewed both with and without the light pipe.

11-1732CZ. In the die-bonding operation, a lucite shield is used to keep an inert atmosphere around the header, and, since lucite has high infrared absorption, a quartz window is used between the detector and the light pipe. The calibration curve is corrected for a 0.3 degree C temperature drop through the quartz window. To find the field of view of the light pipe, which is essential in setting up the light pipe and the detector, you can direct the beam of a small flashlight into the detector eyepiece to illuminate the light pipe. The light emerging from the entrance face onto the target will indicate the area seen by the light pipe.

11-1732DA. INFRARED INSTRUMENTS AND APPLICATIONS.

11-1732DB. GENERAL.

11-1732DC. Control of radiation patterns offers the greatest opportunity for improvement in effectiveness of infrared applications. Significant factors to be considered in the evaluation of an infrared application include efficiency of emitting and directing the energy. The efficiency of generation of infrared energy from the total energy input is the efficiency of the element or infrared generator. The efficiency of emitting the infrared radiation generated from the fixture is the efficiency of the fixture. The housing, reflector, and other parts of the fixture absorb some infrared energy produced by the element, and convert it to heat to be convected away. The efficiency of directing the infrared energy emitted from the fixture into the desired pattern of radiation is the efficiency of control of the pattern.

11-1732DD. EFFICIENCY OF GENERATION. The temperature of an element has a dominant influence of the efficiency of generation. The nature and conformation of the element and many of its environmental factors affect this efficiency materially.

11-1732DE. Quartz lamps operate at temperatures of about 4,000 degrees F, use 100 watts input per inch of heated length, and radiate energy at a reported efficiency of 86 percent based on input wattage. The radiant energy is of short wavelength, and peaks at about 1.17 microns. This radiant energy includes much visible light (about 7 1/2 lumens per watt input), and is termed near-infrared because the energy is in or near the visible spectrum (0.4 to 0.7 microns). Because of the visible component, when quartz lamps are operated at the voltage for which designed, a measurement of the foot-candles delivered to the surface reveals the amount of incident radiation arriving at the surface. Dividing the foot-candle reading by 7.5 provides the calculation necessary to determine watts per square foot of radiation delivered.

11-1732DF. Other high-intensity infrared elements (metal sheath, quartz tube, vycor, open wire, ribbon elements, etc.) are often operated at temperatures between 1,200 and 1,800 degrees F. These elements have had various efficiencies attributed to them, with values between 55 and 63 percent generally accepted. Radiant energy with wavelengths of 3.0 microns or more is termed far-infrared. Since most of the energy emitted from elements below 2,000 degrees F temperature is of longer wavelength than 3.0 microns, they are termed far-infrared elements. There is some visible radiation when the emission temperature is greater than 1,000 degrees F. It is a dull red glow, visible only in darkness at the lower temperatures, increasing to bright orange at about 1,800 degrees F.

11-1732DG. EFFICIENCY OF EMITTING. A less efficient fixture, with respect to radiation emission, will result in a more effective application. The losses in emission are the energy absorbed by the housing, the reflector, and other parts of the fixture. A rule of the thumb is: the greater the control of the radiation pattern exercised by the fixture, the greater the area of the reflector, housing and other parts, and the greater the losses of radiant energy emitted.

11-1732DH. EFFICIENCY OF DIRECTING. Although the nature of the concentrated beam from various fixtures is often identified, the efficiency of producing that concentrated beam is seldom realized. The radiation outside the concentrated beam is often referred to as spill or scatter, or as stray or lost radiation. Redirecting the spill radiation into the desired pattern offers an opportunity for greatly increasing the effectiveness of the fixture. However, means of accomplishing this redirection invariably reduces the efficiency of the fixture.

11-1732DI. IR RADIOMETERS AND PYROMETERS. The amount of radiation from an object is dependent upon the temperature of the object and the emissivity of its surface. Radiometers and pyrometers measure radiation and they must be calibrated with respect to a standard black-body source at a known temperature in order to determine the temperatures corresponding to the meter readings. Radiometers and pyrometers use various types of detectors and optical arrangements in their construction. Pyrometers use a refractive lens method and radiometers use a reflective lens method.

11-1732DJ. RADIATION PYROMETER. A typical radiation pyrometer used for automatic measurement and control purposes is shown in figure 11-259G. The pyrometer head contains two detector cells. The ir radiation from the source being monitored is focused on one cell by a meniscus lens. The lens is made of arsenic trisulfide glass and can be focused from a range of 2 ft. to infinity. Radiation from an internal reference heat source, which can usually be adjusted by an amplifier reference dial, falls on the other cell. The two cells form a bridge which may be balanced by adjusting the internal heat source until the radiation output is equal to that of the external source.

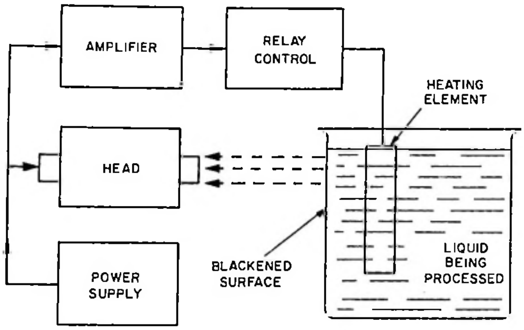


Figure 11-259G. A Typical Radiation Pyrometer Arrangement

11-1732DK. This procedure may be repeated at the tolerance levels required for manual measurement and control. One example would be the measurement and control of a heat-cured product such as an armature; the pyrometer could also be preset at an optimum reading. The difference between the source monitored and the internal reference source is amplified to drive an automatic control to balance the incoming radiation with respect to the preset internal reference source.

11-1732DL. A wide range of detectors and optical windows for very small fields of view and representing various spectral wavelength regions are currently available. The pyrometer discussed is sensitive enough to detect the heat from an area no larger than the finger of a man at a range of 200 yards.

11-1732DM. RADIOMETER. The typical radiometer shown in figure 11-259H uses a high-gain reflective optical arrangement. The ir radiation from the target source being monitored is collected and focused on the detector by a pseudo-Cassegrainian optical arrangement composed of a primary mirror with a clear center area in conjunction with a secondary mirror. A mechanical chopper, placed in front of the detector and electrically driven, modulates the incident radiation. Radiation from an internal

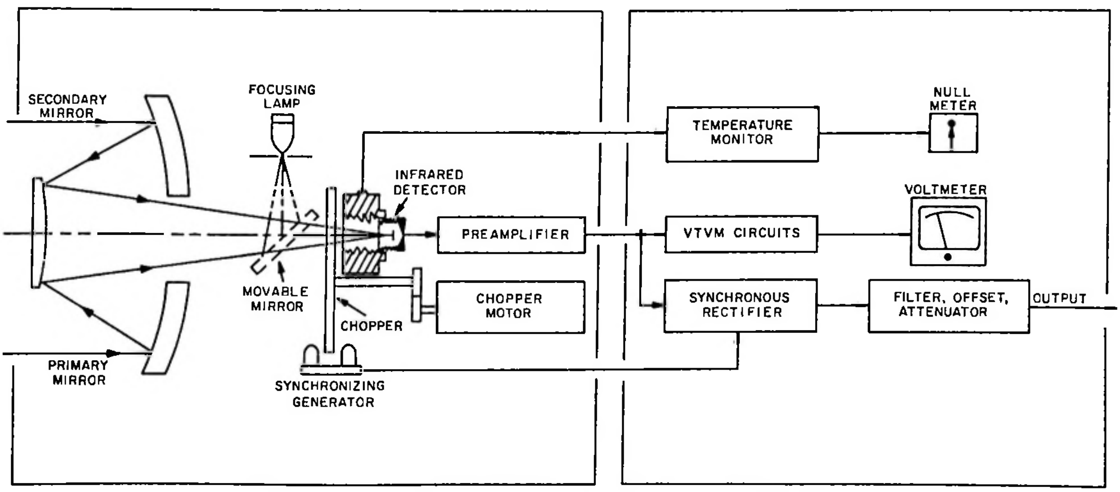


Figure 11-259H. Typical Radiometer Arrangement reference black-body source with variable temperature control is focused by the auxiliary mirror, which has a clear center area, on to the chopper, which has opaque sectors that are silvered on the inside surface. The reference-source radiation is then reflected onto the detector surface.

11-1732DN. The detector, therefore, receives target-source radiation transmitted through the clear sectors of the chopper, and reference-source radiation reflected by the opaque sectors of the chopper, thus modulating both sources in a similar fashion. Calibration can be achieved by adjusting the reference-source temperature to null the incoming radiation. This typical radiometer is used extensively in far ir photography and thermography. As with the radiation pyrometer, a variety of windows and detectors may be used.

11-1732DO. INFRARED CAMERA.

11-1732DP. GENERAL. Several types of infrared cameras operating over various spectral ranges have been developed. These cameras employ a variety of detectors and optical methods and generally employ some kind of scanning mechanism to scan the relatively narrow field of view of the optical arrangement over a wider area. A typical example of such a camera is shown in figure 11-259I.

11-1732DQ. TYPICAL FAR-IR CAMERA. The far-ir camera is an excellent application of the principles described in previous paragraphs. It consists of a standard 8-in radiometer with a flake thermistor detector, a scanning attachment, and a small camera using standard photographic film. A land camera is employed in this application to

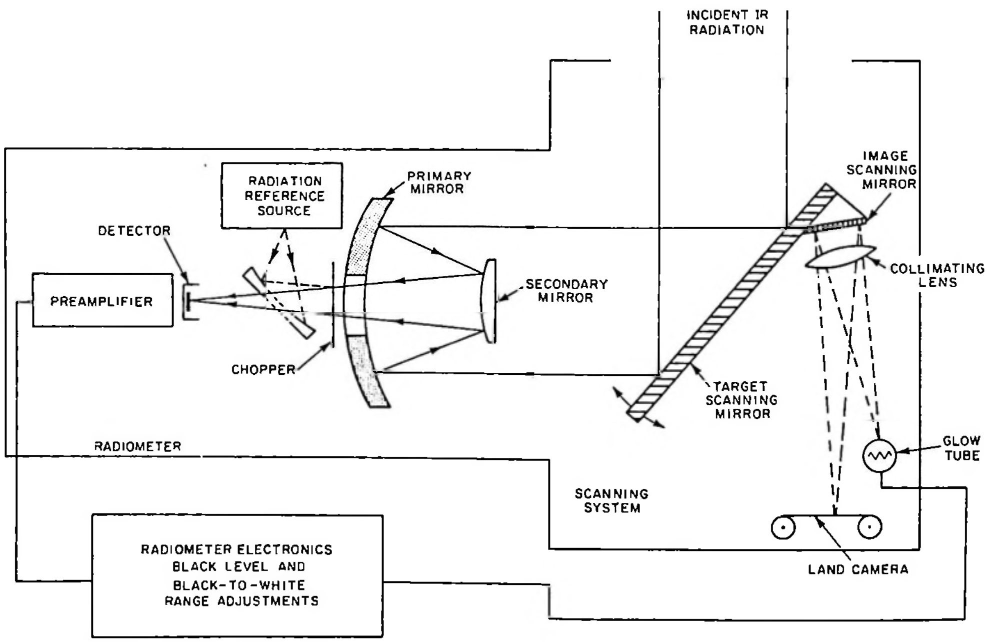


Figure 11-259I. Typical Far IR Camera

provide rapid results. The scan mirror consists of a large plane mirror actuated by a conventional gear cam arrangement in order to scan the small area opposite to the ir detector over the object plane. The detector output is amplified and used to modulate the intensity of a glow tube. Light from the glow tube is focused by a collimating lens and is reflected by a small recorder mirror which is rigidly attached to the back of the scan mirror and moving synchronously with it, across the photographic film. The scanning mirror covers the object plane in a rapid horizontal scan, with a small vertical step deflection after each horizontal sweep. A blanking circuit cuts off the glow-tube output during each rapid horizontal return to avoid retrace lines on the film.

11-1732DR. A calibrated gray scale is automatically superimposed on the film by the camera mechanism. Calibration for a given target is achieved by adjusting the electronic gain so that the black area in the scale corresponds to a reference source at a known temperature, and the black-to-white range in the scale corresponds to a known temperature difference. For a fast visual analysis of a thermal photograph the gray scale is divided into eight black-to-white sections, each section representing a precisely known temperature area. Thus, the temperature variations in the object under investigation can be determined directly from the photograph.

11-1732DS. IR MONOCHROMATORS.

11-1732DT. Monochromators are used in ir applications and in spectroscopy to provide single-wavelength or very narrow waveband radiation of high spectral purity. This is achieved by the use of either a diffraction grating or a prism to provide wide dispersion and high resolution of the ir radiation.

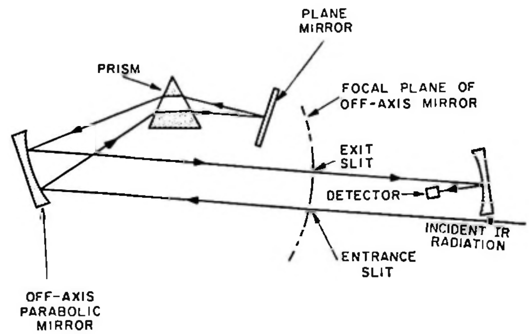


Figure 11-259J. Typical Single-Pass Prism Monochromator

11-1732DU. SINGLE-PASS MONOCHROMATOR. A typical single-pass prism monochromator is shown schematically in figure 11-259J.

11-1732DV. As shown in figure 11-259J, the ir radiation from the object under test passes through the entrance slit to an off-axis paraboloid mirror which sends a collimated beam to the prism. The beam is dispersed by the prism, reflected by the plane mirror back to the prism where a second dispersion occurs, and the beam is then focused by the paraboloid through the exit slit. The exit and entrance slits are placed as close together as possible for minimum aberration. The narrow exit slit permits the passage of a narrow wavelength band of radiation from the dispersed beam to the detector. By rotating the prism, the entire spectral region under investigation may be scanned. The type of monochromator shown has a limited resolution, determined by the finite size and quality of the prism employed. Radiation scattered by the various optical components causes an impure output spectrum. Great improvement in both resolution and purity of spectrum can be achieved by using either a double monochromator or by converting the single-pass instrument to a double-pass monochromator.

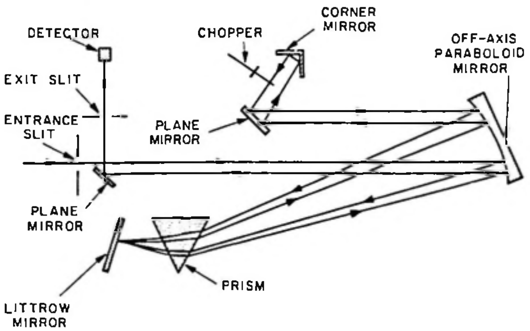


Figure 11-259K. Typical Double-Pass Monochromator

11-1732DW. DOUBLE-PASS MONOCHROMATOR. A typical double-pass monochromator is shown in figure 11-259K.

11-1732DX. Incident ir radiation passing through the entrance slit is collimated by the off-axis paraboloid mirror on a prism where refraction occurs. The beam is reflected by a mirror back to the prism where further dispersion occurs, and the beam is then focused by the paraboloid onto the corner mirror. The corner mirror slightly displaces the beam and returns it through a chopper back through the arrangement for further dispersion in the prism. The emerging beam, now highly dispersed, is focused by the paraboloid on a narrow slit through which the beam passes to the detector. The signals of first-pass radiation of wavelength λ and second-pass radiation of wavelength $\lambda + \Delta\lambda$ emerging through the exit slit are separated by the chopper. The bilateral entrance and exit slits may be controlled equally and simultaneously by a micrometer screw graduated in microns. The angle setting of the mirror may also be controlled by a micrometer.

11-1732DY. To ensure wavelength stability, the monochromator is maintained at a constant temperature by strip heaters, and bimetallic temperature-compensating strips are attached to the mirror. Mirrors are used for high reflection efficiency and low radiation absorption. Excellent resolution and higher energy for a given slit width are obtained in this type of monochromator. Greater purity of the spectrum due to the suppression of scattered radiation is achieved by means of the longer path length traversed by the radiation and the improved dispersion obtained in this instrument.

11-1732DZ. IR GRATING SPECTROGRAPHS.

11-1732EA. Infrared grating spectrographs are used for spectroscopic studies of molecular structure, where rapid automatic recording and extremely high resolution are required. When used with multiple reflection cells, studies can be made in the long path-lengths of gases and liquids. Fast recording can be achieved by using highly sensitive lead sulfide, lead telluride, lead selenide, or indium antimonide detector-cells with rapid response times. The off-axis grating spectrograph is simple in design, uses on-axis paraboloid and plane mirrors, and provides superior images.

11-1732EB. The majority of modern spectrographs are either evacuated or sealed to prevent unwanted absorption bands due to atmospheric gases. All controls to mirrors, slits, and prisms are operated through vacuum seals. The accuracy of required wavelength determines the optical, the geometry and the number of lines per ruled inch on the grating. The angular displacement required of the grating arc may be a fraction of a second of arc for the accomplishment of wavelength measurements to an accuracy of a few hundredths of an angstrom unit.

11-1732EC. An efficient spectrograph requires a wide range of grating driving speeds and extremely uniform motion while minimizing friction. Vibration and heating can be minimized by mounting the drive motor outside the spectrograph and transmitting the required motion through selsyns. Gear reductions of 100,000 to 1 between the drive motors and the grating table will minimize irregularities during slow-speed motion. Motion can be transmitted to the grating table by using an accurate tangent screw which operates against a rigid arm. A typical off-axis-type double-pass ir grating spectrograph is shown in figure 11-259L.

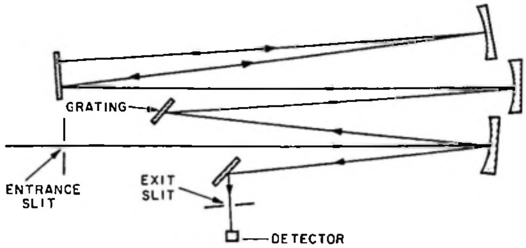


Figure 11-259L. A Typical Double-Pass Off-Axis Grating Spectrograph

11-1732ED. The grating spectrograph in the figure has a focal length of 10 meters, achieved by mounting the grating in Littrow fashion to ensure double passage of the radiation beam through the optical equipment. A resolving power of 120,000 to 140,000 lines per inch in the 1.3 micron to 1.7 micron region, with an increase in resolving power of about one order of magnitude at longer wavelengths, is capable of being obtained with a perfected grating.

11-1732EE. A typical on-axis type direct-recording spectrograph is shown in figure 11-259M. Incident ir radiation focused by a collimating lens on an entrance slit and modulated by a chopper, passes through the central hole of a plane mirror, M1 and is reflected by a paraboloidal mirror, M2. The radiation emerges as a parallel beam which is reflected by the M1 mirror to the grating. The diffracted beam, reflected by

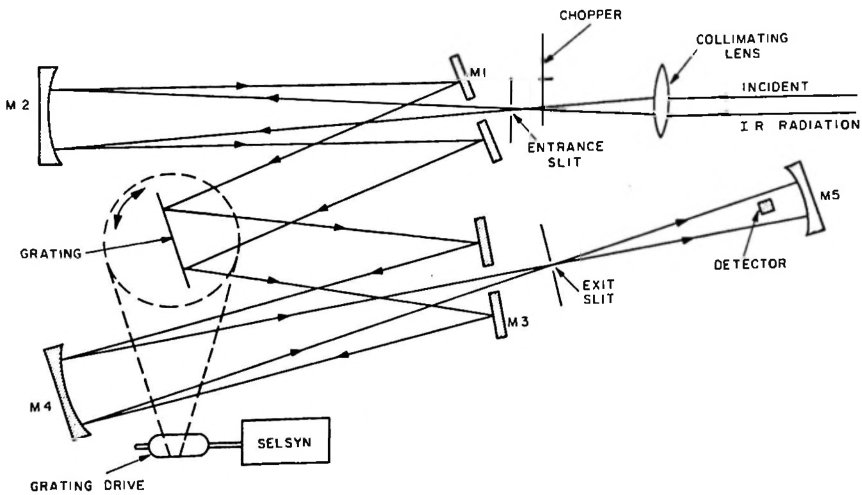


Figure 11-259M. Pfund-type Grating Spectrograph

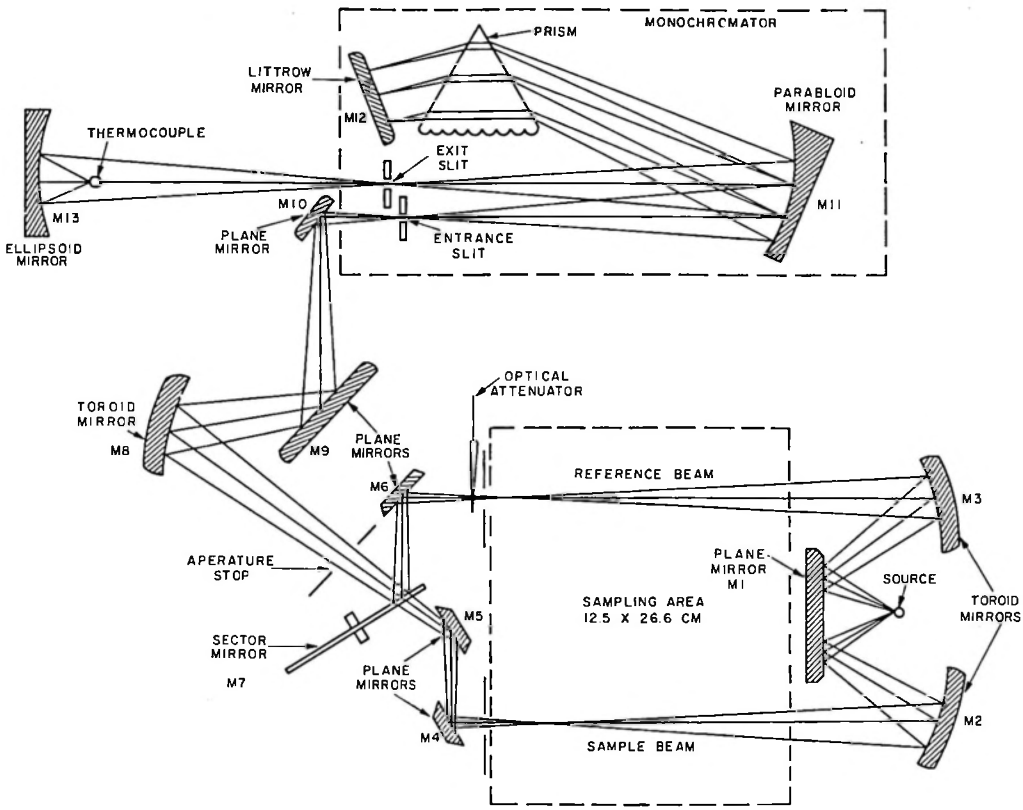


Figure 11-259N. Typical Infrared Spectrophotometer

another plane mirror M3, is focused by a second paraboloid mirror M4 through the central hole of M3 to the exit slit. The beam will emerge onto the ellipsoidal mirror, M4, which will focus the beam on the detector. This type of on-axis arrangement produces a more efficient spectral image and superior resolution.

11-1732EF. IR SPECTROMETERS.

11-1732EG. Modern ir spectrometers and spectrophotometers are used for the rapid analysis of complex organic compounds and the routine analysis of gaseous, liquid, or solid samples. The rapid identification of complex structures, which would require many hours of routine chemical analysis can be achieved by automatically recording absorption spectra.

11-1732EH. INFRACORD SPECTROPHOTOMETER. Figure 11-259N shows an optical schematic of an infrared spectrophotometer. The plane mirror, M1, and two toroid mirrors M2 and M3 split radiant energy from a 1200 degree C heated ceramic-tube source of ir energy in the 2.5 micron to 15 micron region into sample and reference beams. Toroid mirror M2 focuses the sample beam through the sampling area onto a wave filter (comb). Toroid mirror

M3 focuses the reference beam onto a highly-accurate optical wedge or attenuator. After passing through the sampling area, both the sample and reference beams are directed by plane mirrors M4, M5, and M6 onto a semicircular-sector mirror, M7, rotating at 13 revolutions per second. M7 alternately reflects the reference beam and transmits the sample beam through an aperture stop. This ensures that both the reference and the sample beams will be the same size and follow identical paths through the remainder of the optical equipment.

11-1732EI. The signal beam leaving the sector mirror, M7, is composed of alternate pulses of sample and reference radiation, chopped 180 degrees out of phase for comparison purposes. The signal beam is then focused by toroid mirror M8 and reflected by plane mirrors M9 and M10 through the entrance window and to the entrance slit of the monochromator.

11-1732EJ. The beam next diverges to an off-axis paraboloid mirror M11, which reflects it as a collimated beam (parallel rays) onto a prism constructed with a 70 degree apex angle. This prism has a serrated base, forming light traps to reduce scattered light. The 70 degree apex angle provides a high dispersion rate and about 50 percent more energy than does the conventional 60 degree prisms. The prism disperses the component wavelengths into ir wavelengths which are then reflected by the Littrow mirror M12 back into the prism for further dispersion.

11-1732EK. The dispersed radiation is then focused by the paraboloid mirror M11 onto the exit slit as a band of individual wavelengths falling across the slit. The rotational position of the Littrow mirror M12 determines the particular wavelength of the beam which emerges from the exit slit. The beam is then focused by an ellipsoid mirror M13 onto the thermocouple detector. The Littrow mirror M12 is mechanically linked with a recorder-drum shaft so that the wavelength scale on the drum exactly corresponds to the reflected wavelength. True correlation between the drum-wavelength scale and the transmitted wavelength, despite temperature variations, is automatically maintained by a bimetallic strip on the Littrow mount. The monochromator is sealed and dessicated to protect the prism from moisture and dust.

11-1732EL. A high-speed, high-sensitivity thermocouple is enclosed in a steel casing, sealed by a potassium bromide ir window to eliminate pickup, and evacuated to increase the signal-to-noise ratio. When the energy in both sample and reference beams is equal, the thermocouple produces a dc voltage. When radiation at characteristic wavelengths is absorbed by the sample, the intensity of the sample beam is reduced. An unequal signal is then produced at the detector by the pulses of energy from the sample and reference beams. This signal is then converted to an alternating voltage and amplified by a 13-Hz amplifier. This amplified signal is used to drive a servo motor which moves the optical wedge in or out of the reference beam to equalize or null the beam intensities. The wedge is mechanically coupled to a recorder pen with no backlash. Attenuation of the optical wedge is linear, so that the pen records directly in transmittance. This type of recording spectrophotometer is used for both quantitative and qualitative measurements in the fundamental infrared region.

11-1732EM. IR SPECTROMETER WITH PRISM-GRATING DOUBLE MONOCHROMATOR. An ir spectrometer with a prism-grating double monochromator provides the widest possible spectral range. The prism monochromator has four interchangeable prisms and is ganged with a grating monochromator having two interchangeable gratings, by cams linear in wave number and driven by a common shaft. The instrument may be used either as a prism-grating double monochromator, or as a prism spectrometer by blanking the grating monochromator. Gratings, prisms, and cams can be automatically interchanged by means of push buttons. Magnetically operated slits, programmed by a tapped potentiometer, will provide a constant energy background.

11-1732EN. FLAME-TEMPERATURE SPECTROMETER. A typical example of the adaption of the techniques described in the previous paragraph is for measuring large-area flame temperatures. Thermocouples and pyrometers require either the introduction of a probe into the flame, or a knowledge of the emissivity of the object to be measured. With the advent of turbojets, ramjets, afterburners, and rockets, a method is required for measuring the temperature of flames up to 5 ft in diameter, at temperatures of several thousands of degrees centigrade and moving at supersonic speeds.

11-1732EO. One method has been developed that is rapid, accurate, and requires no calibration or attenuation of the gas stream. This method uses an ir spectrometer. The ir radiant energy from a source of known emission passes through a hot gas stream where absorption bands occur due to the water vapor and carbon dioxide present in the gas stream as products of combustion. The ir beam is focused on the entrance slit of a monochromator. At a given wavelength, with the shutter out of the beam, you can measure the radiant energy.

11-1732EP. IR MICROSCOPE.

11-1732EQ. The application of ir spectrometry to minute, nonhomogeneous samples, requires an ir microscope. The elements of a typical ir microscope are shown in figure 11-2590. The optical arrangement

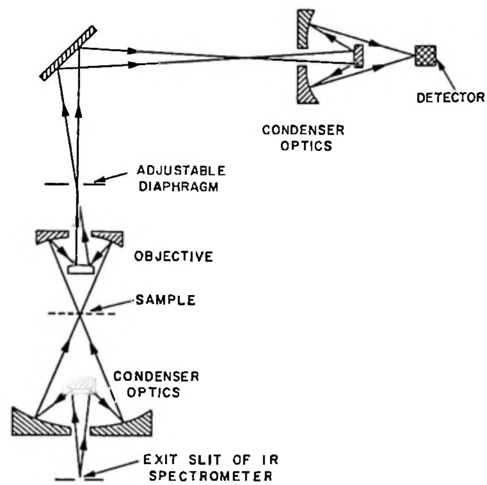


Figure 11-2590. Elements of a Typical IR Microscope

concentrates ir energy from the entire useful exit slit length of the monochromator of an ir spectrometer on a microsample, magnifies the sample image, and focuses the sample image on a detector. Magnifications on the order of 200X of a microsample a few microns thick can be obtained. Reflective optics are used throughout the arrangement to avoid chromatic aberration. The Pfund-type on-axis optics previously discussed are used to minimize distortion.

11-1732ER. IR TELESCOPES.

11-1732ES. The optical arrangement of an ir telescope is always designed for a particular purpose. The telescope itself is used with various types of detecting, chopping, and scanning mechanisms. As an example, the following description pertains to a telescope designed for use with a grating spectrometer and a cooled lead sulfide detector. This instrument was designed for investigation of the ir solar spectrum in the 1 micron to 3.6 micron region, using the solar-tower telescope shown in figure

11-259P. with an equatorial drive.

11-1732ET. A parallel beam of solar radiation is maintained in the field of view by a coelostat controlled by an equatorial-drive arrangement which compensates for the motion of the earth. The long-focal-length solar telescope uses reflective optics throughout to eliminate chromatic aberration. Radiation emerging from the telescope is focused on the entrance slit of a double monochromator, which also uses reflective optics throughout, with the exception of a calcium fluoride prism which has high transmission and dispersion in this spectral region. Radiation leaving the double monochromator, chopped at 1,080 Hz by a sectored disk driven by a synchronous motor, is focused on the entrance slit of a long-focus all-reflecting Pfund-type spectrometer employing a plane reflection grating. A grating with 600 lines per millimeter is used for the 1 micron to 2.5 micron band; a similar reflection grating with 200 lines per millimeter is used for the 2.5 micron to 3.6 micron band.

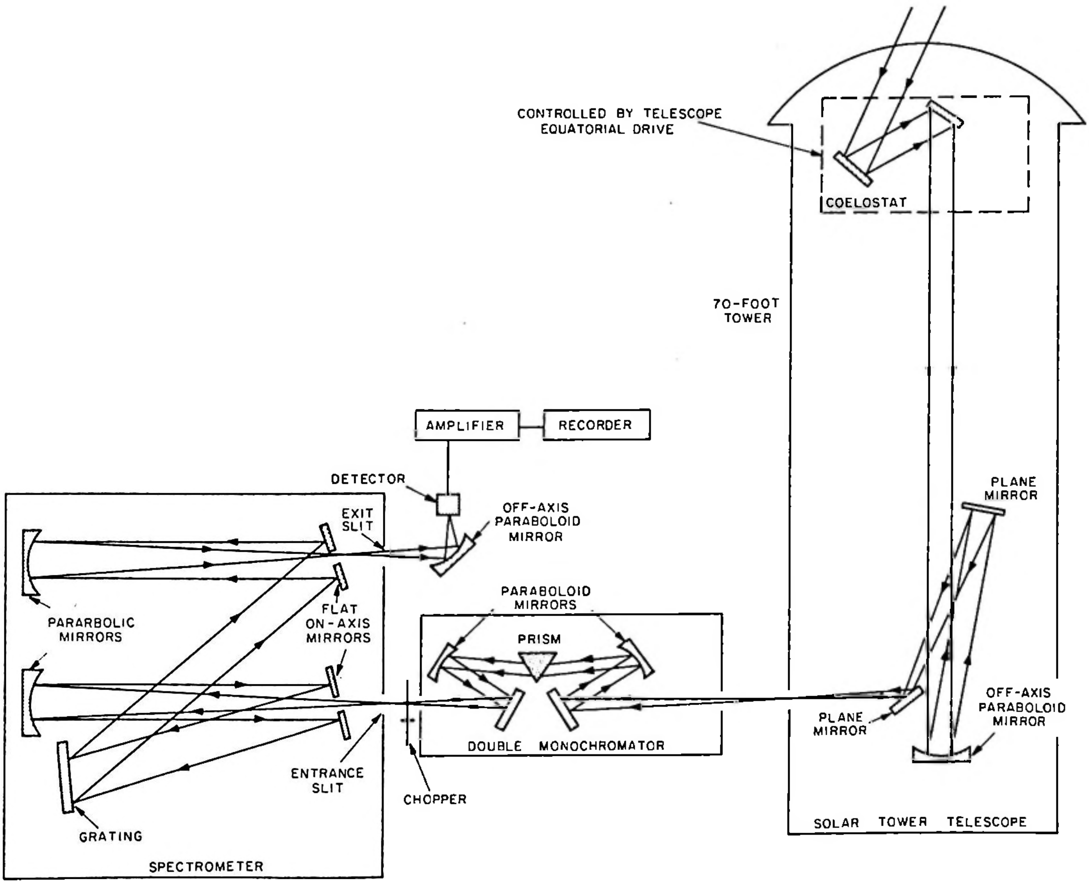


Figure 11-259P. Infrared Solar Spectrometer

11-1732EU. Radiation emerging from the exit slit of the spectrometer is focused by an off-axis paraboloid mirror on a lead sulfide detector cell cooled with solid carbon dioxide and acetone. The cell output signal, modulated at 1,080 Hz by the chopper, is amplified in a variable bandpass amplifier designed to operate at this frequency and match the cell characteristics. The overall time constant of this arrangement is 2 sec.

11-1732EV. IR CONTINUOUS-PROCESS ANALYZER.

11-1732EW. An ir process-stream analyzer is used for the continuous measurement and analysis of gaseous and liquid process streams. This analyzer operates on the principle that the presence of particular chemical substances or compounds in the process stream are detected by their absorption of ir radiation at wavelengths corresponding to the characteristic vibration frequencies of their molecules.

11-1732EX. The present type of ir analyzer operates approximately in the spectral region between 2 microns and 15 microns. In this wavelength region a great many chemical compounds have strong absorption bands and can, therefore, be detected in complex mixtures. The wavelength region mentioned is limited only by the optical windows, materials, and detectors used. The development of new materials and detectors with improved responsivity at longer wavelengths will expand the capabilities of ir analyzers.

11-1732EY. Chemical compounds exhibit one or many relatively narrow absorption bands. These absorption bands are unique for a particular substance. Detectors used in analyzers may be nonselective or wideband detectors, such as thermocouples and thermopiles, or wavelength-selective, such as a capacitor-microphone detector filled with a specific IR-absorbing gas. There are two basic types of analyzers, distinguished by the method they employ for wavelength selection. Nondispersive analyzers use an optical filtering method. Actually, they are self-filtering, since analyzers of this type use a sealed container filled with a gaseous or liquid sample of the component to be detected as a filter. Nondispersive analyzers may employ either nonselective or selective detectors. They are simple in construction and are stable over long periods of time. Dispersive analyzers select the desired wavelengths by means of a dispersive element such as a grating or prism. Movement of a narrow slit-shaped aperture in the ir spectrum isolates the particular wavelength desired. These analyzers have superior sensitivity and background-elimination qualities. However, they are more difficult to stabilize in certain environments than nondispersive analyzers.

11-1732EZ. TYPICAL NONDISPERSIVE IR ANALYZER. An excellent example of a nondispersive ir analyzer is shown in figure 11-259Q. This analyzer is used in the field of process measurement and control for quality control, and for the measurement of toxic and combustible gases or vapors. It can accomplish almost any gas or liquid analysis which can be made by ir spectrometers.

11-1732FA. Beams of radiation from two similar sources of ir radiation, in the form of nichrome filaments, travel through stainless steel cells. One beam traverses the sample cell; the other beam traverses the comparison cell. Radiation emergent from the two cells converges in the beam combiner into a single condenser microphone detector. Radiation is absorbed in the detector where it causes the gas temperature and pressure to be increased which causes in turn, a movement of the membrane. This

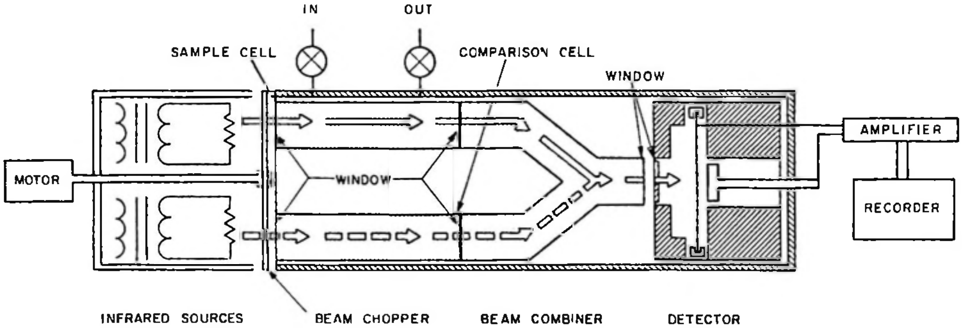


Figure 11-259Q. Nondispersive IR Analyzer

movement is converted and electrically amplified to produce an output signal.

11-1732FB. A metal slide positioned between the sources and the cells alternately blocks the radiation to the sample cell and to the comparison cell. The amplifier is tuned to produce an output signal only when a variation in light intensity occurs at the alternating frequency. When the beams are equal in intensity an equal amount of radiation enters the detector from each beam and the output becomes zero.

11-1732FC. When the gas or fluid to be analyzed is introduced into the sample cell, radiation reaching the detector from the sample beam is reduced by absorption. The beams now become unequal, and the radiation entering the detector flickers as the beams are alternated, causing a corresponding expansion or contraction of the detector gas. The resulting movement of the membrane varies the capacity of the condenser microphone, which generates an electrical signal proportional to the difference between the two radiation beams. This signal is then amplified and applied to a recorder or indicating meter.

11-1732FD. DISPERSIVE IR-ANALYZER. A dispersive analyzer employs an optical null principle to continuously measure the ratio of radiant power at two different wavelengths. This has the advantage of largely eliminating the effects of changes in sample-cell transmission and changes in source temperature. The IR radiation from a source is chopped as shown in figure 11-259R, and directed through the sample cell into a monochromator, where any two preselected wavelengths are isolated by a split Littrow mirror. The two preselected wavelengths are selected so that one is at a strong absorption band of the material under investigation, and the other is not absorbed at all, or only slightly. The source and monochromator are separately enclosed and pressurized. This type of dispersive analyzer, unlike the nondispersive type previously discussed, does not have to store a gaseous sample of the material to be detected in the detector compartment. It can, therefore, analyze nonvolatile and unstable materials.

11-1732FE. EMISSIVITY-MEASURING INSTRUMENT.

11-1732FF. An instrument used for the rapid and accurate measurement of the emissivity of various materials which may be in liquid, paste, powder, or solid form is shown in figure 11-259S.

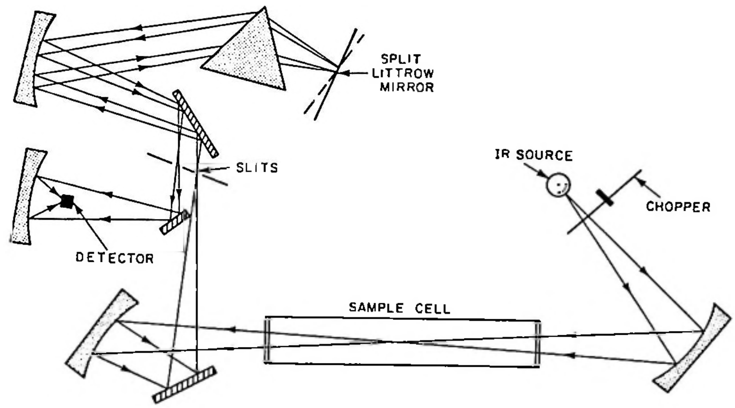


Figure 11-259R. Dispersive IR Analyzer

11-1732FG. The device in the figure, is composed of a standard radiometer and an emissivity-measuring device coupled to the radiometer by an alignment collar. A collimating optical arrangement composed of primary and secondary front-surfaced mirrors forming a Cassegrainian telescope similar to that used in the radiometer, collects the ir energy reflected from a selector mirror placed near the focal point and directs it into the radiometer.

11-1732FH. In one position the selector mirror directs ir radiation from a black-body standard-radiation source through the optical arrangement to the detector. In the

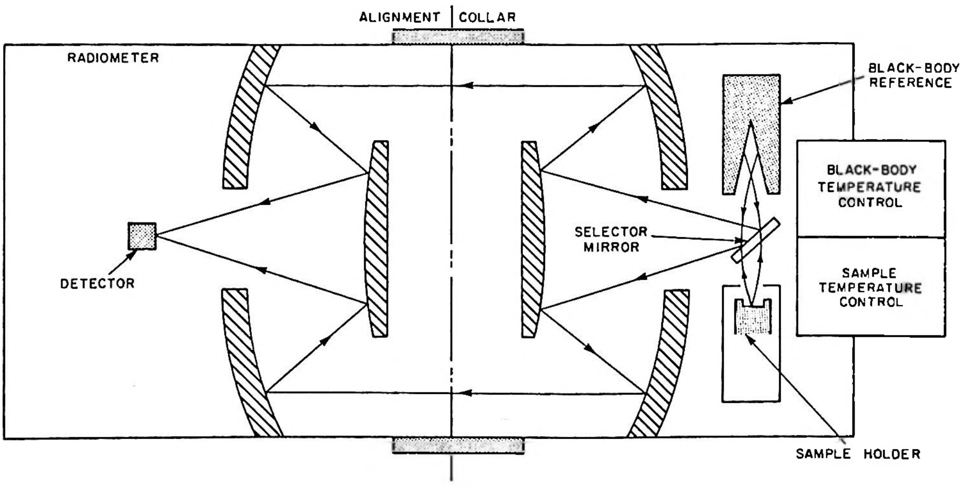


Figure 11-259S. An Emissivity-Measuring Device other position, the selector mirror directs ir radiation from the sample to the detector. The temperatures of both the standard source and the sample are independently and accurately regulated by means of temperature controllers set at a desired temperature. Radiation from the standard black-body at the same temperature as the sample is measured as a constant voltage output. The ratio of these two measurements provides the emissivity. To eliminate inaccuracies due to stray radiation and temperature-gradient effects, these measurements should be repeated at another temperature.

11-1732FI. IR THERAPEUTIC LAMPS.

11-1732FJ. Infrared therapeutic lamps have been found to be beneficial in heat-treating human beings for certain ailments. It has also been found that the infrared radiation which penetrates most deeply into human flesh lies between the wavelengths of 7,000 and 14,000 anstroms. The lamps used for this purpose are special and operate at a somewhat higher filament temperature than the standard lamp of the same size. For use where the light would be objectionable, the lamp is made with a black-glass bulb which still transmits the infrared energy.

11-1732FK. Infrared drying and heating lamps are gas-filled, and have a tungsten-filament. This type of construction is beneficial for heat rather than light and for long life. It has been found that 65 to 95 percent of the watts input to a lamp is radiated as light and heat. However, only 6 to 12 percent of this amount represents the invisible infrared radiant heat which accompanies the visible light. All this energy, both visible and invisible infrared, creates heat when absorbed by the object the energy strikes. Such radiant energy does not heat the air during passage. Wavelengths longer than 50,000 anstroms are absorbed by the glass bulb. Open-type heaters, like bathroom radiators, are of relatively low temperature, generating the longer wavelengths which are largely absorbed by the surrounding air. That is why a lamp with a filament temperature of some 4000 degrees F is such an efficient generator of radiant heat. A drying lamp operates at about 6 lumens per watt and at a color temperature of about 2500 degrees F. Heat lamps are a modification of the drying lamp technique and operate at approximately 1 lumen per watt and at a color temperature of approximately 2000 degrees K. Better control in high-wattage lamps, for concentration of energy on the work is obtained by using special filament forms.

11-1732FL. IR SIGNAL GENERATOR.

11-1732FM. An infrared signal generator has been developed which provides an infrared source of variable wavelength and power which can be readily adjusted for general experimental work, production testing, and pre-flight calibration. The output is available at your discretion as either a point source or a collimated beam as shown in figure 11-259T. The source of power is a tungsten filament contained in a housing with a special window. This hot body provides a broadband radiated noise spectrum in the wavelength range 1 to 14 microns.

11-1732FN. The radiated power is focused through a potassium bromide lens on the entrance aperture of a four-pass monochromator, with a set of cylindrical stops for prefiltering. The spectral profile of the instrument due to the field stops alone provides the effect of a double monochromator in reducing stray light. The spectral profile of the output is determined by the entrance and exit apertures of the monochromator, which are circular and programmed to provide the resolution specified.

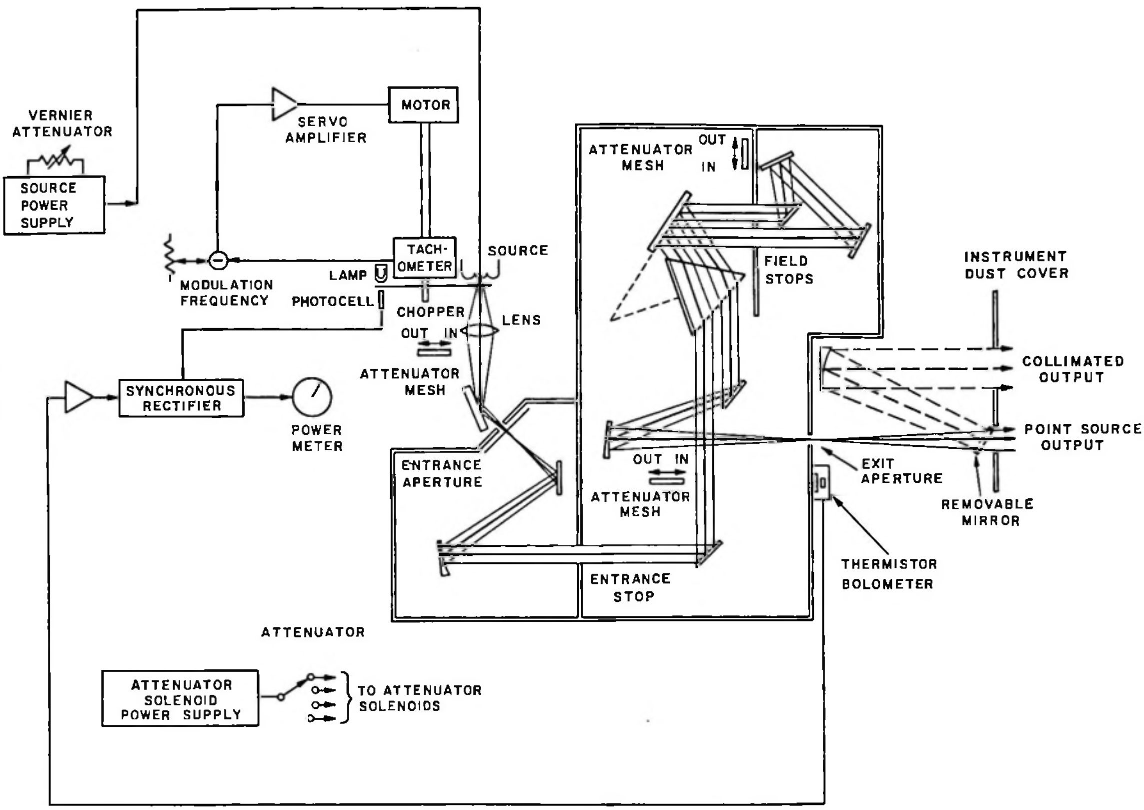


Figure 11-259T. Typical IR Signal Generator

11-1732FO. Since heat oriented objects, by virtue of their temperature, radiate infrared, the case and the interior components of a test instrument will radiate a broadband signal. When you set the attenuator settings very high, the output signal from the test instrument may well be equal to or less than the normal background radiation. By connecting an arrangement of ir equipment to view the output from the test instrument, the output signal plus this background radiation will become apparent. For this reason a rotating device termed a chopper is placed at the input to the monochromator and the chopping frequency is adjusted by means of a calibrated servo motor drive. Two ranges, representing a 2 to 70 Hz and a 70 to 2600 Hz signal are provided. The blade of the chopper is located between the light source and the lens assembly which focuses the radiation on the monochromator and attenuator. Between the light source and the lens assembly the signal level is independent of the attenuator setting and the ratio of signal-to-chopper radiation is favorable to the point that the resultant modulation can be considered perfect. Since the background radiation is not modulated, it can be eliminated as a variable. With a chopper, it is possible to perform extremely low level signal measurements such as are required for measuring noise, figures, and detectivity of equipments and components.

11-1732FP. DETECTOR SELECTION.

11-1732FQ. GENERAL. The selection of ir detectors can be facilitated by dividing detectors into two classes. These are: point and area detectors, or extended detectors. Point detectors have a sensitive surface that integrates all incident radiant energy into a single expression. Point detectors include such types as radiation thermocouples, pneumatic-type detectors, and bolometer-type detectors. Area, or extended detectors provide a picture of the radiation pattern. They include image tubes, vidicons, and evaporographs. The following paragraphs contain data pertaining to the advantages and disadvantages of processing techniques associated with these two classes of detectors as applied to fault isolation techniques.

11-1732FR. POINT DETECTORS. The point detector can be used in at least three ways. In one method, a probe can be placed on each component that is being monitored. A temperature differential indicator will provide readings (operating temperature minus ambient temperature) that can be compared with a standard list of readings for the particular unit being tested. From changes in readings (initial, positive, and negative) the fault can theoretically be isolated. This is an entirely manual operation. This bench-type method bypasses all optical and cooling constraints.

11-1732FS. In another method, specially matched probes can be permanently mounted directly on all components being monitored. Probes, such as thermistors, are small and weigh but a few milligrams, thus permitting placement in cramped quarters. The components themselves need not be exposed to view. A sequencer can sample each probe, one at a time, constantly scanning the circuit. Minute temperature changes can be indicated easily, accurately, and rapidly in any of several ways: a controller can stop the sequencer at the point where unbalance occurs; an out-of-tolerance light could glow indicating unbalance position; or a computer could look for changes and print out the probe location or perhaps the name of the component that has changed and the degree of change. Such controls can operate to a precision of 0.0005 degree C. This method also bypasses the optical and cooling constraints of other detection techniques. It is adaptable to either semiautomatic or completely automatic techniques and may be used on the bench or in flight.

11-1732FT. Point detectors can also be placed in a matrix to form an area-type detector. The matrix can be located immediately above or below the components being checked. This method must be completely automatized and can be used on the bench or in flight if the associated processing equipment can be kept light in weight. Fault isolation in this case depends on either locating hot spots or locating thermal pattern changes. The original matrix pattern can be stored in a computer and subsequent patterns compared with it by means of the computer. As pointed out previously, an optical arrangement is required if the detector is removed to a distance from the component or radiation source. Even with a matrix placed as close as 3 or 4 inches from the infrared source, it is doubtful that much more than hot spots could be sensed and located accurately without an optical arrangement. The application of this matrix technique will be determined by the amount of detectivity that is required, that is, the degree of infrared change that you are trying to detect and the extent to which you are willing to employ auxiliary equipment.

11-1732FU. AREA DETECTORS. Area

detectors are primarily pattern recognition devices. The components are scanned and an image is formed of the radiation intensity of the components. As with the matrix technique, a computer stores the original image, compares it with subsequent patterns, and prints out significant variations. With this method the components to be observed must be exposed; they cannot be hidden by other components, covers, etc. The scanner is likely to be in the form of a television camera, and its distance from the circuit would depend on the area to be scanned and the focal length of the lens.

11-1732FV. Based on the preceding discussion, the items that must be considered in selecting an infrared detector and the processing techniques are: the component operating temperatures and the degree of infrared changes to be monitored; the location of check-out (bench, inflight, etc.); the component accessibility; the required location of the detector with respect to the radiation sources; the surrounding temperatures; the ventilation of chassis under test; the detector parameters; and the associated equipment required.

11-1732FW. DETECTOR CHECKOUT PROCEDURE. IR detectors have evolved into highly sensitive and complex instruments; consequently, the equipment required to test, calibrate, and evaluate these detectors has become highly sophisticated. Early methods of testing IR detectors were simple to perform, but did not provide measurement data within specified limits. In one early method of testing, an IR detector was placed inside one end of a light-tight box, and an ohmmeter was attached to the detector to indicate the resistance of the IR detector. This measurement, which is called the dark resistance value, was noted by the technician. A hole was then cut into the other end of the box so that a flashlight beam could strike the detector. The resistance of the IR detector was noted again, this time being known as the light-resistance value. The two resistance values were compared and the ratio of dark-to-light resistance value was calculated. If this ratio fell within a specified range, the IR detector was considered to be acceptable.

11-1732FX. As the number of IR detector applications increased, additional types of IR detectors were developed, thus resulting in a requirement for more specific and more thorough test and evaluation techniques. The technician needed the means for making accurate signal, bias, frequency response, spectral response, and noise measurements. Consequently, test equipment was designed to present this information. Typical IR detectors that must be evaluated in connection with the use of detector measurement equipment include cooled detectors, multi-element arrays, detectors used with optical elements, and hardware-wound detectors. One of the later methods used for complete evaluation of IR detectors is illustrated by the block diagram in figure 11-259U. In this method, detectors to be tested were mounted on plug-in fixtures and inserted into a temperature-monitored compartment to permit control of the black-body temperature range during the test procedure.

11-1732FY. Detector bias is supplied from a dc source which includes meters for monitoring both voltage and current outputs. The value of these two quantities indicates the stability of the detector under test. The black-body temperature range can be varied from 50 to 1000 degrees C. A precision potentiometer and a calibrated thermocouple are built into the black-body (box interior) for monitoring the temperature. If the modulator housing and baffle are water-cooled, you can measure long wavelength response detectors, which are very sensitive to dc background radiation. The variable speed modulator operates over the

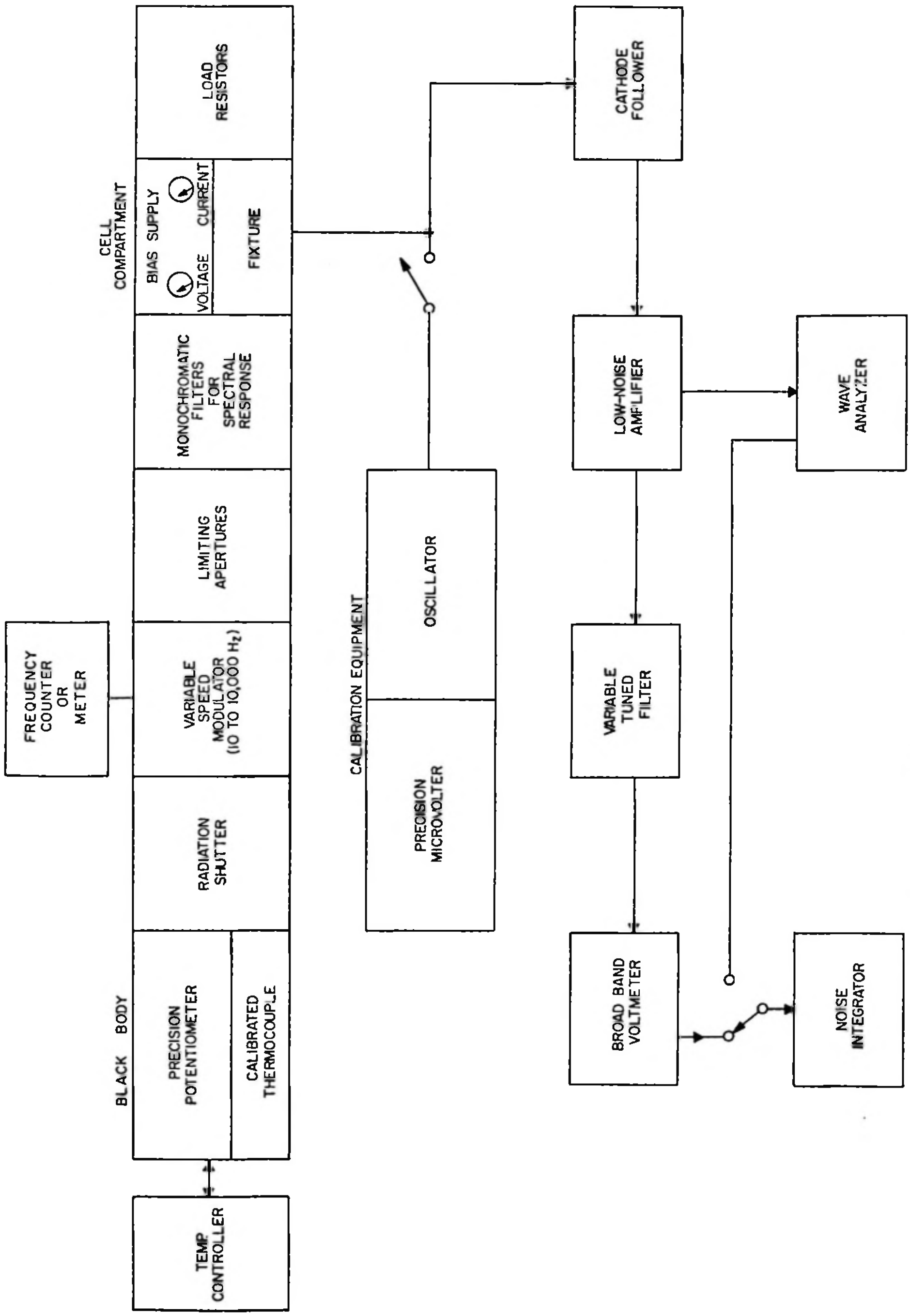


Figure 11-259U. IR Detector Evaluation Block Diagram range of 10 to 10,000 hertz.

11-1732FZ. The black-body energy emitted from the detector can be varied by mounting limiting apertures on a rotating wheel so that different size apertures are available. Spectral response measurements are facilitated by a set of monochromatic filters normally spaced at 0.5-micron intervals. A finer resolution of spectral response may be obtained by using a monochromator.

11-1732GA. The calibration equipment is composed of a precision micro-voltage divider and an oscillator. A special low-noise preamplifier amplifies the signal and noise of the detector and the output is split for two readouts. One output is applied to a wave analyzer for narrow-band signal and noise power spectra measurements. The other output is filtered through a variable-tuned filter and applied to a broadband voltmeter to simulate actual equipment bandwidths. Use of the noise integrator improves the efficiency of noise readings.

11-1732GB. Since the temperature of objects in space varies with altitude, internal power dissipation, and radiation and reflection qualities of the surface, it is essential for you to know the effects of temperature upon resistance, sensitivity, wavelength, frequency response, etc. The testing of an IR detector at low temperatures may be accomplished by mounting the detector in a vacuum dewar flask and pouring a liquid coolant into the flask. This coolant may be either liquid nitrogen or a combination of dry ice and acetone. These two coolants produce temperatures of -196 degrees C and -78 degrees C, respectively.

11-1732GC. In order to test an IR detector at temperatures varying from +20 to -100 degrees C, the equipment should contain a temperature chamber capable of holding the detector to a predetermined stable temperature and at the same time maintaining a moisture-free atmosphere around both the detector and the detector transmitting window to prevent frost from accumulating.

11-1732GD. A console has been built which can measure entire detector arrays for reconnaissance type satellites. This console includes automatic calibration and gain controls and can be pre-programmed for recalibration as often as needed. Signal, noise, bias voltage and current, and signal-to-noise ratio are all read out directly on a digital voltmeter or, at the same time, they can be put on tape, typed, or punched into cards. A simple computer program will provide you with all of the figures of merit. Optimum bias at each temperature is determined automatically by programming a selection of bias voltages into the programmer. Noise integration time is selected to coincide with the noise bandwidth being used. An optical focusing system enables the operator to automatically measure contour sensitivity of the detectors. Contour output is printed on a strip chart recorder.

SECTION XI

COMPUTER EQUIPMENT TESTING

11-1733. GENERAL.

11-1734. COMPUTER TYPES.

11-1735. GENERAL. Computers are used extensively by the Air Force for various applications, such as air defense data processing, supply data processing, weapons control, flight control, and telemetering. A computer may be a small unit consisting of a few circuits, or it may be a large complex equipment involving thousands of circuits. Two general types of computers are used; these are the digital and the analog computers. Information applied to a digital computer is separated into discrete units for processing, whereas the information applied to an analog computer is handled as continuously variable quantities. The digital computer performs mathematical operations by repeated addition, while several methods are used in analog computers. Large computers which require a high degree of accuracy are normally of the digital type. Although analog computers are not accurate enough for many applications, voltage or current representing specific conditions may be developed by analog circuits or devices and then converted to a digital form. After mathematical operations are completed, the resulting data may be converted to an analog voltage or current to operate certain controls or indicators.

11-1736. ANALOG COMPUTERS.

11-1737. Analog computers are used for such applications as determining the altitude of targets that are detected by radar equipment, and solving specific types of mathematical equations for design calculations. Such computers are usually composed of summing, multiplication, differentiating, and integrating circuits. Figure 11-260 is a block diagram of the type of computer that can be used to calculate the altitude of a radar target. In this example, an electromechanical computer receives values of slant range from the radar receiver and values of elevation angle from a synchro connected to the antenna, and converts this data into altitude information.

11-1738. DIGITAL COMPUTERS.

11-1739. GENERAL. The basic sections of a digital computer are shown in figure 11-261. These perform the input, output, memory, arithmetic, and control functions. All data enters the computer through the input section, where it may be converted to a digital form and placed in a buffer storage

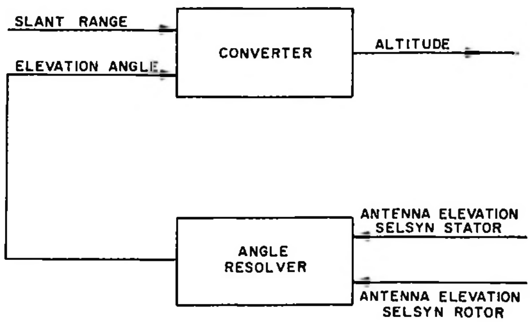


Figure 11-260. Analog Computer for Calculating Altitude of Radar Targets, Block Diagram

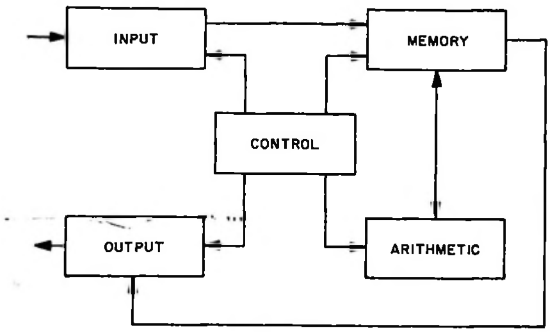


Figure 11-261. Basic Digital Computer Sections, Block Diagram

device. The memory section accepts this data at specific intervals and places it in the proper storage location. Two paths lead out of the memory element, one to the control section and one to the arithmetic section. Instructions are transferred from the memory to the control section, in which they are decoded and certain commands are set up for operation of the computer.

11-1740. INPUT. The input section accepts information in various forms and converts it to a form which can be used by other sections of the computer. Input information is applied to computers by many different methods, such as punched card readers, magnetic tapes, paper tapes, typewriters, and telephone lines that transmit data from distant sources.

11-1741. OUTPUT. The results of the computer operation are fed to the output for either local or distant applications. Some examples of output units are line printers, card punchers, magnetic tapes, counters, cathode-ray tubes, and telephone lines that are connected to distant equipment. The output section varies widely from one computer to another.

11-1742. MEMORY. The memory section stores information until it is needed by one of the other sections of the computer. Memory devices are assigned addresses, and the appropriate address is specified in the computer signal when information is needed or is sent to be stored. Several types of memory devices are used in computers; the most common are magnetic cores, ultrasonic delay lines, electrostatic tubes, multivibrator circuits, and magnetic drums. At present, magnetic cores are the most popular device, primarily because of their high speed, stability, and ability to retain information if there is a power failure.

11-1743. ARITHMETIC. Addition is the basic operation of the arithmetic section of a digital computer; other arithmetic functions are simply variations of the addition function. For example, multiplication in the arithmetic section is simply a repetitive addition. The basic circuit in the arithmetic section is usually the multivibrator, which is used to store or transfer the results of computation.

11-1744. CONTROL. The control section generates signals at the proper time to cause a desired action to take place within the computer. This is normally accomplished by decoder circuits, along with multivibrators and other switching devices. The control section keeps track of the instructions that are to be decoded and may perform part of the decoding. In addition, this section provides timing pulses that synchronize all sections of the computer.

11-1745. MAINTENANCE TECHNIQUES.

11-1746. To meet reliability requirements and provide efficient trouble-shooting procedures, special maintenance techniques have been developed for large computers. Maintenance program tests provide loop circulation of a simulated data pattern.

This type of testing will determine whether the test signal becomes distorted in the circulation process and whether the control circuits in the loop allow passage of the entire succession of signals. Marginal checking circuits are included in some computers to detect aging of the circuit parts before a failure occurs. In addition to these special techniques, normal signal-tracing methods are used to locate the specific circuits in which malfunctions exist. Basic measurements are used to detect faulty parts.

11-1747. MAINTENANCE PROGRAMS.

11-1748. GENERAL.

11-1749. Computers can be given an over-all check by means of maintenance programs. A maintenance program provides a thorough and rapid method for you to detect failure in a special portion of a computer. This type of over-all maintenance check is very flexible and efficient. These programs use the same type tape, memory, computing, and drum circuits as the operational programs. A program can be changed when the computer or auxiliary components are changed, and the program can be constantly improved. No extra test equipment is required since the computer circuits are used to perform the test. Testing by means of maintenance programs also results in the computer circuits being used in a more normal manner than during signal-tracing procedures. When a program has been checked and accepted as a good maintenance tool, it is not subject to deterioration. In contrast, test equipment may be checked and accepted only to become unreliable shortly after being placed in actual use.

11-1750. Maintenance programs are divided into three main classes: reliability, diagnostic, and utility programs. Maintenance programs that are used to detect the existance of errors are called reliability programs. Reliability programs should be arranged to check as many computer circuits as possible. Maintenance programs that are used to locate the circuits in which computer malfunctions originate are called diagnostic programs. An effective diagnostic program should locate the source of trouble as closely as possible. Actually, in many cases reliability programs have diagnostic features, and diagnostic programs have reliability features. For convenience, a program is called either a reliability or diagnostic program depending on its intended emphasis. In general, programs that check rather than diagnose are shorter and simpler.

11-1751. BASIC PROGRAMS.

11-1752. GENERAL. A program is a series of instructions which control the operations of a computer. Each instruction is used to cause some action which is a part of the over-all task that the computer must perform. Therefore, an instruction may be considered to be the basic building block of a computer program.

11-1753. An efficient program makes full use of the instructions which are available to accomplish the task in the shortest possible time and uses the least number of instructions. In most cases, one criterion, either time or the number of instructions, has to be chosen over the other, and the program is developed along this line. If time is important, you should try to write a maintenance program which uses instructions of short duration but may use quite a few memory locations for storage. On the other hand, if time is relatively unimportant, but only a few locations are available, you must choose instructions which do a number of things or cause the computer program to run through the same program more than once.

11-1754. To write a satisfactory maintenance program, it is necessary to have a thorough knowledge of the instructions that can be used. This includes execution time, the over-all purpose of the instruction, when the instruction may be used, and the state of the computer after the instruction has been carried out. In addition, you should know whether the instruction can be indexed and what internal conditions must be satisfied before it can be executed.

11-1755. HALT INSTRUCTION. The halt instruction causes the computer to stop executing instructions under program control. However, any operation which is in progress at the time the halt instruction is decoded will be completed first. For example, if information is being read into the memory from a deck of punched cards, all the cards will be read before the computer halts, even if the halt instruction was issued just after the reading operation began. The address portion of the halt instruction is not used; therefore, indexing is not possible. When the computer is halted by instruction, the program counter retains the address of the instruction immediately following, so that restarting the computer will cause this next instruction to be executed.

11-1756. CLEAR AND ADD INSTRUCTION. The clear and add instruction is used to enter a quantity into the accumulators from the memory section without changing the sign or magnitude of the words. This instruction is normally used when it is desired to begin a type of addition problem. The accumulators are first cleared, and then the information selected from the memory section by the address portion of the clear and add instruction is transferred to the computer registers. Addition of the data in the registers and the data in the accumulators is started; however, since the accumulators are cleared to +0, this addition has the over-all effect of transferring the word from the memory section into the accumulators unchanged. The memory area used is unchanged, and the registers are cleared to +0 after execution of the clear and add instruction.

11-1757. ADD INSTRUCTION. This instruction is similar to the clear and add instruction except that it does not provide for clearing the accumulators before the addition process begins. Thus, the add instruction will generate the sum of the data contained in the specified memory address and the data that is in the accumulators. This sum is placed in the accumulators, and the registers are cleared to +0. It should be noted that the add instruction can cause an overflow if the numbers added together are sufficiently large. If this happens, the result in the accumulator is meaningless.

11-1758. FULL STORE INSTRUCTION. The full store instruction is used to transfer words from the accumulators into the memory area specified by the address portion of the instruction. Thus the results of any operation performed by the arithmetic section are placed in the memory section for future use. The contents of the specified register are first cleared, and then the contents of the accumulators are transferred to the memory section buffers.

11-1759. RELIABILITY PROGRAMS.

11-1760. GENERAL. Reliability programs are used in both preventive and corrective maintenance tests to detect circuit failures rapidly and to discover failures that may occur only under particular operating conditions. Examples of troubles that are not evident at all times are failures that appear at specific repetition rates or for certain combinations of bits. In order to detect such failures, it is necessary to use reliability programs which check logical operation, paths of information flow, timing, ability of the computer to perform all functions, execution of instructions, etc.

11-1761. TYPES. Reliability programs check either the logical functioning of an entire computer section or the logical functioning of individual circuit groups in a section. Whichever method is used, it is assumed that associated circuits which are not directly checked by the program are in satisfactory operating condition. Thus these programs can be considered to fall into two categories, first order and second order. First-order reliability programs check the operation of an entire computer section, while second-order programs check the operation of assemblies or circuit groups, such as registers, counters, etc. In most cases, first-order programs are merely a combination of several second-order programs.

11-1762. INTERPRETATION. A reliability program provides a good-or-bad indication regarding the ability of the tested computer section or circuit to perform its operating functions. For example, consider a reliability program that checks the switching time of relays within a specific section of a computer. As long as the switching time of the relays is within normal limits, the reliability program will indicate satisfactory operation. When the switching time is excessive, however, there is an indication that maintenance is required. If the program runs successfully, there are no failures within the checked area. In the event of a failure indication, the failure may be in the area being checked or in another area that has been assumed free of trouble. Diagnostic maintenance programs should then be used to locate the source of trouble.

11-1763. DIAGNOSTIC PROGRAMS.

11-1764. GENERAL. To be efficient, maintenance programs for diagnostic applications must enable you to narrow the area of a failure down to the smallest possible number of circuits. This can be accomplished by employing increasing-area, decreasing-area, overlapping-area, and large-area checks. The most effective method will depend on the particular type of computer being tested.

11-1765. INCREASING-AREA CHECK. A maintenance program using the increasing-area check initially tests a small number of circuits. If a check indicates that all tested circuits are operating properly, successive checks are run in which progressively greater numbers of circuits are added. By this method, circuits which are found to be operating correctly are used to check other circuits. This process is continued until all circuits that can be checked by a maintenance program have been tested.

11-1766. DECREASING-AREA CHECK. When this method is used to find a trouble, a large number of circuits are initially checked by the maintenance program. If trouble is detected in a large area, additional checks are made of successively smaller portions of the equipment until the stages affected by the failure are not included in the test area. You should then be able to determine which stages are defective. If the check of a large area reveals no error, the remaining large areas of the equipment are checked until the trouble is detected. In many cases, trouble can be located more rapidly by this procedure than by the increasing-area method.

11-1767. OVERLAPPING-AREA CHECK. Another efficient method of locating trouble to within a small section of the equipment is the overlapping-area check. The routines of this type of maintenance program overlap each other. Thus, a failure is located at the overlapping portions of the routines which indicate the presence of trouble.

11-1768. LARGE-AREA CHECK. You may not be able to program an effective maintenance test for some small sections of a computer. A maintenance program can then be used only to detect the general area in which the malfunction occurs. When the general area is located, conventional trouble shooting will be necessary to find the circuit in which a failure has occurred.

11-1769. UTILITY PROGRAMS.

11-1770. Utility programs are used as aids for both operation and maintenance programming procedures. This type of program is used to print out information from magnetic cores, magnetic drums, or other storage devices within the computer memory section. It is also used to transfer maintenance programs from punched cards or magnetic tape into the computer memory section. Utility tracing programs provide a printed record of the contents of various computer registers to enable you to follow maintenance program operations.

11-1771. MARGINAL CHECKING.

11-1772. GENERAL.

11-1773. Marginal checking is a preventive-maintenance technique that is used for some Air Force and commercial computer equipment to detect the decrease in reliability of circuit parts due to aging. Aging circuit parts almost invariably change in value, current-handling capabilities, or voltage limitations. Generally, the changes brought about by aging are gradual, and you will not notice any variation in the normal operation of the equipment. For maximum equipment reliability, parts that are beginning to deteriorate must be detected and replaced before a failure occurs.

11-1774. Marginal checking is usually controlled by a maintenance program. The program directs the computer to perform the normal computer operations of addition, subtraction, etc, while the program varies certain circuit parameters about their normal values. In this way, the computer is made to perform normal functions under adverse operating conditions.

11-1775. To accomplish marginal checking, certain operating conditions are changed from their normal values. Since circuit part values normally change with age, the variations that can be introduced before a failure occurs become less as the parts age. The amount of variation, from the normal value, that can be introduced before equipment failure occurs is called the margin of reliability of the circuit or group of circuits being tested. If the margin is regularly checked and its gradual decrease noted, as shown in figure 11-262, the time of circuit failure can be anticipated. Three methods of changing operating conditions for marginal checking are as follows:

a. Variation of dc supply voltages.

b. Variation of circuit values.

c. Variation of electron-tube filament voltages.

11-1776. DC SUPPLY VOLTAGE VARIATION.

11-1777. The most versatile method of marginal checking is by variation (excursion) of the dc supply voltage for one or more circuits. Causing an excursion of a circuit's dc supply voltage will simulate the changes that normally result from the aging of circuit parts. Gradually increasing the excursion of the supply voltage to a circuit will eventually result in a circuit failure regardless of the circuit's age. Figure 11-263 shows the relationship between circuit reliability and the excursion

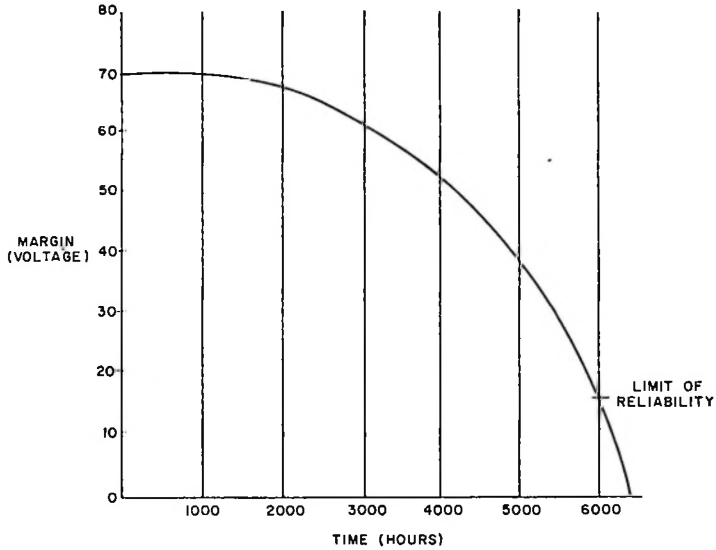


Figure 11-262. Typical Circuit Part, Life Curve

voltage required to cause a circuit failure. The magnitude of the voltage excursion necessary to cause a failure is called the margin of the voltage on the circuit. This margin becomes smaller as the circuit ages. When the circuit fails at the normal operating voltage, the margin is zero.

11-1778. As long as the possibility of circuit failure is low the circuit is considered satisfactory. For the example shown in figure 11-263 the circuit reliability of 80 percent is acceptable. When the voltage excursion necessary to cause a circuit failure decreases so that the circuit reliability is below the 80-percent value, maintenance must be performed to replace parts or an entire plug-in assembly. The level at which the circuit reliability is acceptable must be determined for each circuit or circuit group that is tested by marginal-checking methods.

11-1779. A life curve for a circuit can be drawn, as shown in figure 11-262; you can then estimate the time for a circuit to age to the point where failure is expected. Maintenance will be performed on the equipment when the marginal check indicates that a failure will probably occur before the next marginal check is due. Such a life curve is useful for maintenance applications when the margin limit is approached gradually during each successive testing period. Certain circuits will not have useful life curves because they age gradually but fail suddenly. These circuits may operate for long periods with practically no change in margin, and then the margin drops sharply in a short period of time. Since no appreciable decrease in margin takes place before the circuit fails, it is difficult to determine when a faulty part should be replaced and marginal checks are not effective. The circuits for which useful life curves cannot be established, however, represent only a small percentage of general computer circuits. Most circuits have life curves that enable you to determine the time at which a circuit part must be replac-

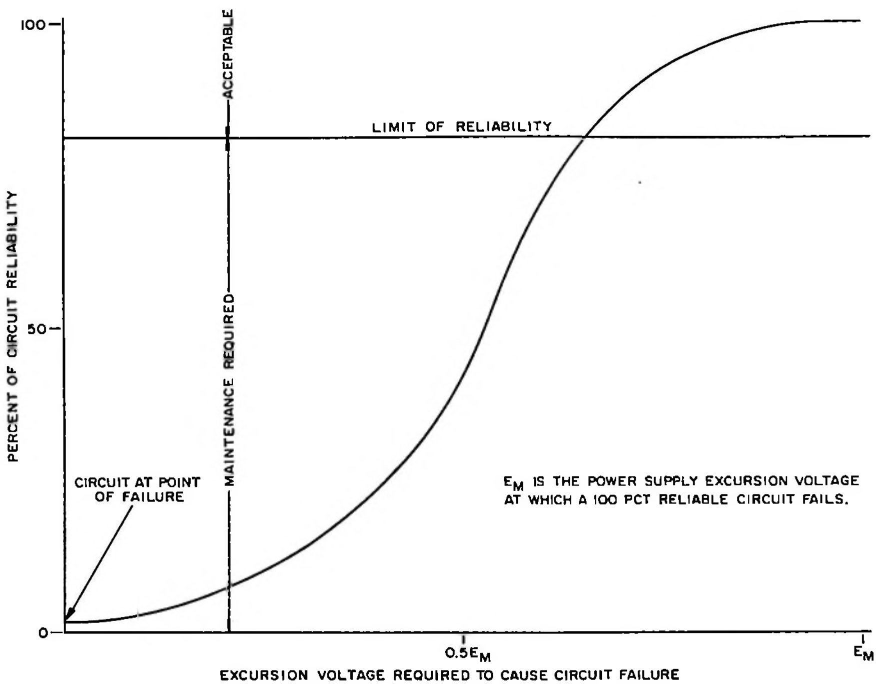


Figure 11-263. Circuit Reliability Versus Excursion Voltage Required To Cause Circuit Failure

ed to prevent failure during computer operating time.

11-1780. CIRCUIT PART VALUE VARIATION.

11-1781. Failure of circuit parts other than electron tubes can be anticipated by periodically simulating the aging of the parts. Aging can be simulated by changing resistor, capacitor, or inductor values of a circuit and noting the effects of such changes. This can be done when adjustable electrical parts are included in the circuits to be tested. By periodically measuring the amount of change necessary to cause circuit failure, and by determining how much normal component aging will produce such a change, it is possible to establish the time at which a part must be replaced to prevent circuit failure before the next scheduled test.

11-1782. FILAMENT VOLTAGE VARIATION.

11-1783. Reducing the filament voltage of electron tubes is a method used to simulate a condition of low cathode emission which causes circuit failure. Anticipation of this condition enables you to replace a deteriorating tube before failure occurs.

11-1784. For marginal checking applications, low cathode emission can be simulated by periodically decreasing the filament voltage from its normal operating value to a value at which useful emission ceases. Each time a marginal check is made, the difference between the normal filament voltage and the filament voltage at which the equipment fails is noted. Over a period of time, this difference, or margin, will become smaller. Eventually, a point is reached where the margin is so small that you can expect emission failure before the next periodic check. The weak tube should then be replaced to prevent its failure during normal operating time. This method of marginal checking is seldom used because it is limited to the prevention of computer failures caused by tube deterioration.

11-1785. PROCEDURE.

11-1786. Figure 11-264 represents a circuit that can be selected by a maintenance program for marginal checking. The operation of this circuit is such that successive pulses place a "1" in FF1, transfer it to FF2, and then clear both flip-flop stages.

11-1787. During the marginal checking operation, an excursion is applied to the supply voltage line of FF1. Assume that, at some voltage value, the computer senses that a "1" was not transferred to FF2. The excursion is stopped and the margin noted. To determine which stage has failed, you might first make voltage and resistance checks of the circuit parts of FF1. If this check indicates that FF1 is functioning correctly, it is possible that the output pulse from gate AG1 is so small that a slight change in the supply voltage applied to FF1 causes the circuit to fail. It is also possible that gate AG2 has aged to the point where any decrease in the signal output level of FF1 will result in failure of the pulse to be transferred to FF2.

11-1788. MARGINAL CHECKING UNITS.

11-1789. Marginal checking units for variable supply voltage applications include a voltage-regulator unit, relays to connect the test voltage to circuits, and control circuits necessary to initiate marginal checking. When high currents must be furnished by the regulator, an amplidyne may be used to provide a variable dc supply voltage. This device provides a well-regulated output voltage that can be accurately controlled over a wide positive-to-negative range.

11-1790. A simplified block diagram of marginal checking units is shown in figure 11-265. The relays, which are controlled by maintenance programs, can connect any one of three circuit groups to the marginal-checking voltage regulator. Circuit group

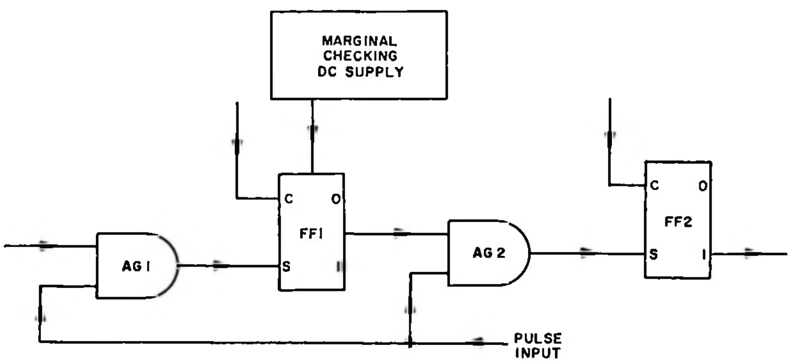


Figure 11-264. Typical Circuit Selected for Marginal Checking, Logic Diagram

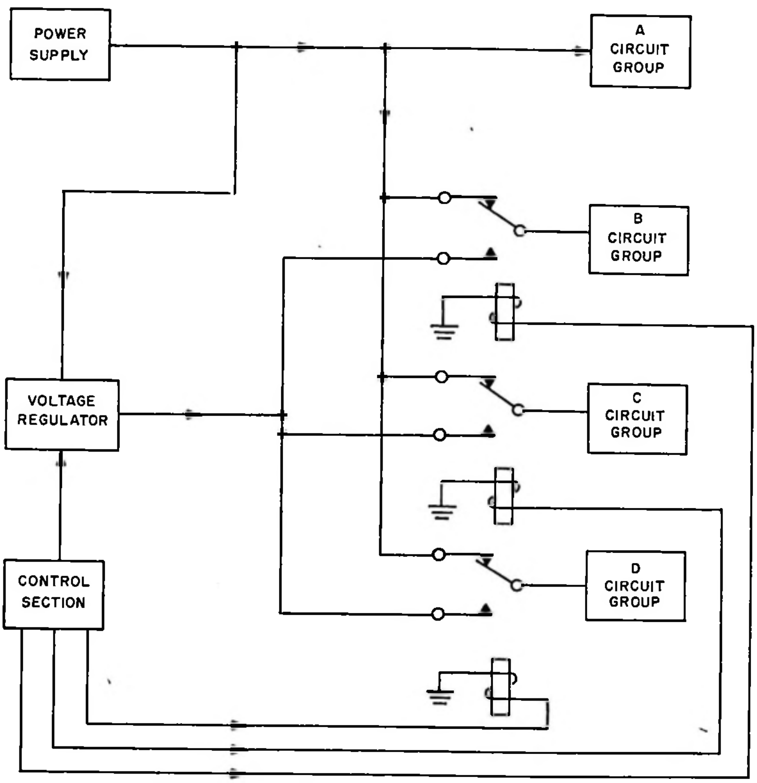


Figure 11-265. Marginal Checking Units, Simplified Block Diagram

A represents circuits that cannot be effectively tested by this method; therefore, it is not connected to a relay in the marginal checking circuit.

11-1791. COMPUTER DIAGRAMS.

11-1792. GENERAL.

11-1793. Logic diagrams are usually included with computer maintenance instructions to show the paths of data flow through the computer circuits. The use of logic diagrams will often simplify over-all circuit tracing and thus help you to locate trouble in equipment that has many circuits. When you have determined that a trouble is in a specific circuit or group of circuits, conventional schematic diagrams should be used as an aid to locate faulty parts.

11-1794. LOGICAL SYMBOLS.

11-1795. GENERAL. Symbols used in logic diagrams are geometric figures that represent the various types of computer circuits. Common electrical symbols are used to represent circuits and devices such as amplifiers, rotating machinery, and test points. Symbols peculiar to computer applications are used for circuits and devices such as magnetic drums, flip-flop stages, and gating circuits. Basic symbols that are used for Air Force technical manuals are shown in figure 11-266, and common reference designations for these symbols are listed in table 11-14. Variations of these symbols are often used in commercial and older Air Force technical manuals, and abbreviations are usually added to the symbols. Therefore, many computer technical manuals that include logic diagrams with nonstandard symbols contain tables of symbols and abbreviations.

11-1796. Each symbol represents a function or, in special cases, a combination of functions. The symbols are interconnected by lines that indicate information or control signal paths, but not actual wire connections. As a rule, there is no correlation between the logical functions and the individual modules of a computer. One module may be used to perform several logical functions, or a logical function may require several modules.

Table 11-14. Logic Diagram Reference Designations

DESIGNATION	NAME
AG	And Gate
CT	Counter
FF	Flip-Flop
OS	One Shot
OR	Or Gate
RG	Register
ST	Schmitt Trigger
SR	Shift Register

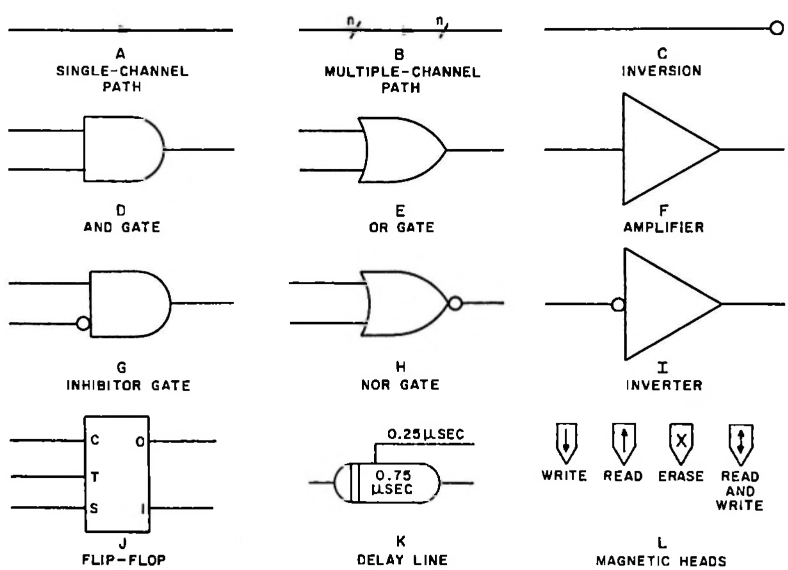


Figure 11-266. Logical Symbols for Computer Diagrams

11-1797. Summarization logic diagrams are sometimes included in a technical manual to further simplify explanations of equipment operation. Such a diagram may show groups of circuits that perform a related function, such as an accumulator, storage register, or full adder, as one block. All signal inputs and outputs of the block are identified, and the function may be marked on the block.

11-1798. SIGNAL PATHS. Single- and multiple-channel paths that connect logical symbols are shown in parts A and B of figure 11-266. A multiple-channel path is distinguished by cross-ties, and the number of channels is indicated by a circled number. The direction of signal flow is indicated by arrowheads superimposed on the channel-path line. You should keep in mind that these lines indicate data or control-signal paths, not wires; and the arrowheads indicate the direction of data flow, not voltage or current.

11-1799. BASIC CIRCUITS. AND-gate, OR-gate, and amplifier symbols are shown in parts D, E, and F of figure 11-266. These symbols can be modified by the addition of a small circle, as shown in part C of the figure, to indicate signal inversion. The inversion symbol can be placed adjacent to the inputs or outputs of symbols to represent circuits such as inhibitor gates, nor gates, and inverters, as shown in parts G, H, and I of the figure.

11-1800. The symbol for a flip-flop multivibrator is shown in part J of figure 11-266. This circuit is a bistable multivibrator that has three possible inputs: set S, clear (reset) C, and trigger T; and two possible outputs: "1" and "0". The flip-flop circuit assumes the "1" state when an effective signal appears at the S input; it assumes the "0" state when an effective signal appears at the C input. When an effective signal appears at the T input, the flip-flop output state is reversed. If signals appear simultaneously at more than one input, you can not be certain which will trigger the circuit and the output has no meaning.

11-1801. COMMON DEVICES. The symbol in part K of figure 11-266 represents a passive delay device such as an electromagnetic or ultrasonic line. The total duration of the delay is written on the symbol. If the delay device is tapped, as indicated in the figure, the duration of the delay from the input to the tap is written within parentheses adjacent to the tap output. These delay lines are used for both timing and memory applications. Twin vertical lines indicate the input side.

11-1802. Part L of figure 11-266 shows the symbols for magnetic heads. The functions of the heads are indicated by the symbols; the first symbol is for reading only, the second is for writing only, the third is for erasing only, and the fourth is for reading and writing. Magnetic drum assemblies can be drawn with the angular relationship of the heads marked in degrees.

11-1803. SUMMARIZATION SYMBOLS. Figure 11-267 illustrates summarization symbols that are used for binary registers. Part A of the figure shows a storage register consisting of a group of four flip-flop circuits connected in parallel. The symbol on the left side shows the individual input and output paths. In the symbol on the right side, the four S inputs, the four "1" outputs, and the four "0" outputs are represented by multiple-channel lines.

11-1804. Part B of figure 11-267 shows a four-stage binary register; the contents of this register can be shifted, one stage at a time, to the right or left when triggered by a shift input signal. The symbol on the left side shows four individual parallel input and output paths. The symbol on the right

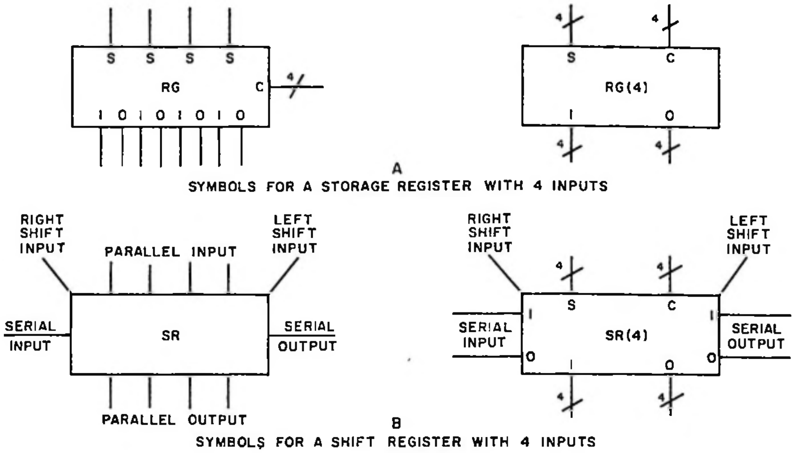


Figure 11-267. Summarization-Type Logical Symbols

side combines the parallel S inputs, C inputs, "1" outputs, and "0" outputs into multiple-channel lines. Figure 11-268 is a more detailed logic diagram of a four-stage flip-flop shift register. This diagram also represents different functions by a single symbol since gate circuits may be included as part of a flip-flop symbol.

11-1805. STYLIZED PULSE WAVEFORMS.

11-1806. Pulse waveforms are used on logic diagrams to indicate the nature and timing of the signals. Waveform symbols are used to represent the signal level, time of occurrence, beginning and ending times, and rise and fall times, as shown in table 11-15. Examples of positive- and negative-going pulses are shown, and representative voltage and time values are used. A pulse may be represented as a single or double line, and, in the case of information-bearing pulse trains, not all pulses are necessarily present.

11-1807. Figure 11-269 shows a representative pulse waveform with level and time designations that are used in computer manuals. V_d represents the maximum voltage amplitude of the pulse. The rise time, t_r, is the time required for the beginning of the pulse to increase from $0.1 \times V_d$ to

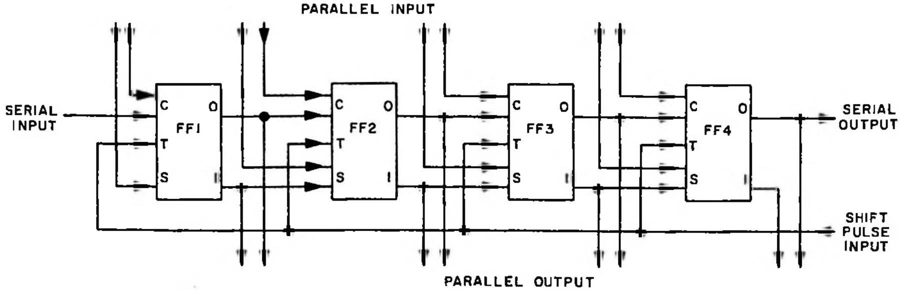
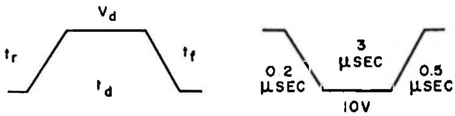


Figure 11-268. Flip-Flop Shift Register, Detailed Logic Diagram

Table 11-15. Logic Diagram Pulse Waveform Notations

WAVEFORM NOTATIONS	DESCRIPTION
	LEVEL STARTING AT PULSE TIME A AND ENDING AT PULSE TIME B
	PULSE OCCURRING AT PULSE TIME C
	PULSE TRAIN WITH FIRST PULSE OCCURRING AT PULSE TIME D AND LAST PULSE AT PULSE TIME E
	PULSE WITH VOLTAGE LEVEL V_d, DURATION t_d, RISE TIME t_r, AND FALL TIME t_f
$\frac{V_d}{t_d}$ $\frac{10}{3}$	PULSE NUMERICAL RELATIONSHIP USED WITHOUT A WAVEFORM SYMBOL

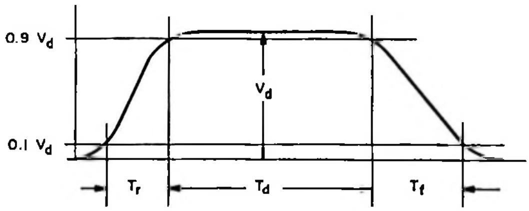


Figure 11-269. Representative Pulse Waveform

$0.9 \times V_d$. The pulse duration, t_d, is the time that the pulse is equal to or greater than $0.9 \times V_d$. The fall time, t_f, is the time for the end of the pulse to decrease from $0.9 \times V_d$ to $0.1 \times V_d$.

11-1808. Figure 11-270 shows the waveforms and notations that may be used with an AND-gate circuit. Two pulse signals with amplitudes of 20 volts and durations of 1 and 10 microseconds are applied to the gate. The resulting output is a pulse signal having an amplitude of 20 volts and a duration of 1 microsecond. The output pulse rise time

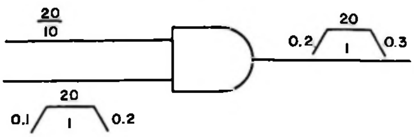


Figure 11-270. Waveform Designations for Gating Circuit

is 0.2 microsecond, and the fall time is 0.3 microsecond.

11-1809. LOGICAL EQUATIONS.

11-1810. Equations listed on logic diagrams provide additional information regarding the signal and circuit. Such logical equations may indicate the type of gate circuits used and the conditions that must be present to produce an output signal from a particular gate. Figure 11-271 shows a schematic diagram, the related logic diagram, and the logical equation of a diode network with AND-gate and OR-gate circuits. The equation is read, E is equal to a binary 1 if A and B are 1 or C and D are 1. The equation consists of two and functions that form an overall or function. Only one of the and functions must equal a binary 1 for E to equal 1.

11-1811. Other relationships that can be expressed in equation form include signal designations with switch numbers and positions marked above them as follows:

$$\begin{array}{ccccccc} & & S_1 - N & & S_1 - 1 & & S_1 - 0 \\ a_1 & = & A & + & GND & + & (-13\ v) \end{array}$$

This equation means: signal a_1 is the same as A (-13 volts or 0 volts) if switch S_1 is in the N position, or signal a_1 is at ground potential (0 volts) if S_1 is in the 1 position, or signal a_1 is at -13 volts if S_1 is in the 0 position.

11-1812. Flip-flop inputs can also be expressed as a logical equation. A flip-flop stage with and-gate circuits in the input might have the following equation:

$$FF_1{:}1 = f_1f_2f_3$$

$$0 = f_4f_1f_5$$

That is, the 1-side inputs to flip-flop stage FF_1 are f_1 and f_2 and f_3, and the 0-side inputs are f_4 and f_1 and f_5.

11-1813. ALIGNMENT.

11-1814. GENERAL.

11-1815. Circuit adjustments may be necessary when input or output units are either connected or disconnected from a computer. These adjustments may vary the gain of amplifier stages, the frequency of a timing circuit, or the voltage furnished by a power supply. The need for other adjustments will be indicated by preventive-maintenance checks. Power-supply outputs must be adjusted to furnish the correct current to magnetic cores and the correct voltage to amplifier and oscillator stages. The speed of magnetic drum and tape devices must be adjusted to provide computer signals having the correct pulse width and timing.

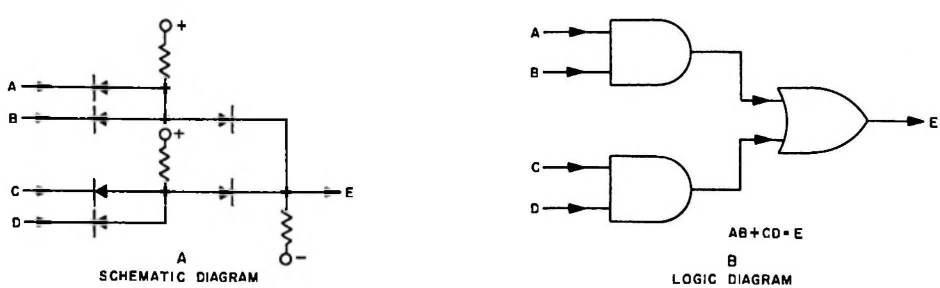


Figure 11-271. Diode Gating Network Diagrams

11-1816. CORE MEMORIES.

11-1817. Core memory circuits may require realignment because of the following reasons:

a. Changes in operating currents due to the aging of parts.

b. Replacement of plug-in units, such as digit plane drivers, memory gate generators, or timing stages.

c. Accidental changing of the original potentiometer settings.

The procedures for adjusting a core memory require checks of the memory timing cycle, the digit-plane-driver current, the read current, and the write current.

11-1818. The memory timing cycle can be measured with an oscilloscope calibrated to measure time. The time between the clear-memory, start-memory, set-read, clear-read, set-inhibit, set-write, clear-write, and clear-inhibit pulses must be checked. If a discrepancy exists in the timing of any of these pulses, an adjustment must be made. In some memory units this is done by changing the taps on the memory clock delay line.

11-1819. An oscilloscope calibrated to indicate voltage can be used to determine the digit-plane-driver, read, and write currents. Maintenance programs can be used to inject the proper computer signals into the memory-coil arrays. You can then calculate the current through each drive line after measuring the voltage drop across the terminating resistor for each set of windings. Potentiometers located in the memory units are used to set the current to the correct amplitude. Recheck and, if necessary, readjust the read and write currents after the first adjustments are completed. This readjustment may be necessary because of interaction between the two circuits.

11-1820. After the read and write currents have been adjusted, you can check the balance by means of an oscilloscope and, if necessary, make a fine adjustment. Connect the oscilloscope probe to a read-write drive line. Run a maintenance program to apply synchronized bursts of read and write pulses to the drive line. If the read and write currents are equal, the oscilloscope display should be similar to that shown in part A of figure 5-216. If any unbalance is present, the oscilloscope display will resemble the display shown in part B or C of the figure. Either the read or write current potentiometer should be adjusted slightly to provide a balance.

11-1821. MAGNETIC DRUMS.

11-1822. GENERAL. Magnetic-drum read and write head adjustments are always required in the initial stages of installation and after replacement. When the read and write functions are accomplished by two drum heads, the heads must be adjusted with respect to both signal amplitude and timing. Since these two adjustments are interacting, they must be made concurrently. The interaction of these adjustments results from the construction and location of the heads, head bars, and rotor drum. When writing and reading are accomplished by a single drum head, the timing will always be correct and only the amplitude adjustment must be made.

11-1823. The use of two oscilloscopes is recommended for drum head adjustments. One oscilloscope is connected to the input of the drum read amplifier to monitor the head amplitude, and the other is connected to the output of the drum read amplifier to provide a check of timing. Typical wave-

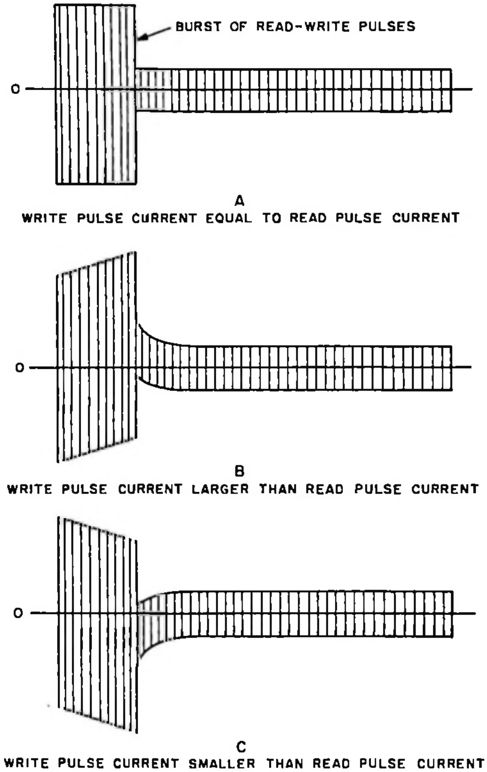


Figure 11-272. Magnetic Core Read and Write Current Balance Check Waveforms

forms that may be obtained are shown in figure 11-273.

11-1824. TIMING ADJUSTMENT. For a coarse adjustment, loosen the head-retaining screws and move the head the desired amount. Then tighten the screws until they are just snug enough to retain this position. Observe the timing pattern on the oscilloscope, and use the retaining screws to rock the head for a fine adjustment. One screw should be tightened and the other loosened to tilt the head to the position which provides accurate timing. Movement of the screws

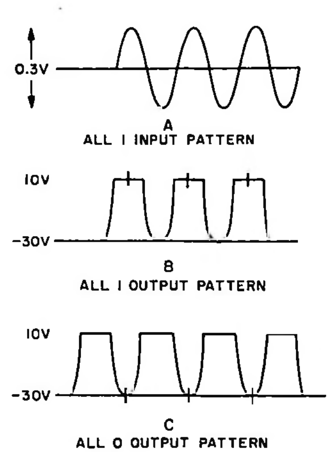


Figure 11-273. Typical Drum Read Amplifier Test Waveforms

in either direction should be slight. Excessive tightening of a screw may cause the head to be cocked an abnormal amount and result in a distorted input signal to the read amplifier.

11-1825. SIGNAL AMPLITUDE ADJUSTMENT. The output signal amplitude of a read head is adjusted by varying the air gap between the drum surface and the read head core. On some magnetic drums, the read heads are adjusted to provide a signal output which is 75 percent of the amplitude that is measured when the head is in contact with the drum surface. To make this adjustment, proceed as follows:

a. Connect an oscilloscope to the output of the read head.

b. Loosen the lock nut on the amplitude screw just enough to permit a snug fit.

c. Each time an adjustment is made, tap the amplitude adjustment screw lightly to overcome static friction in the associated mechanism.

d. Turn the amplitude adjustment screw until the head core just makes contact with the drum surface. The head core should remain in this position only long enough to measure the head peak-to-peak voltage output with the oscilloscope; otherwise the head may be damaged.

e. Turn the amplitude adjustment screw until the amplitude of the head output waveform is reduced to 75 percent of the value measured in step d. For example, assuming that the contact amplitude is 300 millivolts, you should adjust the head to provide an output equal to 75 percent of 300 millivolts, which is 225 millivolts.

f. The acceptable limits for the output of a drum head is usually listed in the equipment technical manual. For the example above, the final amplitude setting must fall within the limits of 125 to 300 millivolts; that is, if the 75-percent value is higher than 300 millivolts, it must be reduced to this maximum value. If the resultant is lower than 125 millivolts, the head is probably defective and may have to be replaced.

g. If noise spikes appear at the output of the drum head at the 75 percent setting of the amplitude adjustment, they should be minimized by lowering the amplitude below this value. However, the amplitude must not be decreased below the lower limit described in step f.

h. When the desired amplitude is obtained, tighten the lock nut, taking care not to disturb the adjustment.

11-1826. TIMING AND AMPLITUDE RELATIONSHIP. A definite timing and amplitude relationship exists between corresponding read and write heads, and this relationship must be taken into account when adjustments are deemed necessary. If a timing error is detected in a read head and you cannot move the head far enough to obtain correct timing, the following procedure is recommended:

a. Move the corresponding write head in a direction to correct the timing error. Write a new test signal with the head in the new position.

b. Adjust the write-head amplitude. This is accomplished by removing the write-head plug from the write-head-amplifier cable connector and applying the output of the write head to the vertical input of an oscilloscope. Then using the write head as a receiver, adjust the gap between the write-head core and the drum surface to provide an amplitude equal to 75 percent of the contact amplitude. This must be within the limits specified in the equipment technical manual. The limit for the write head may be narrower than for the receive head; for this example assume the limits to be 175 and 200 millivolts.

c. Reconnect the write head to the write-head amplifier, and write a new test pattern. Readjust the read head for correct timing and amplitude. If the minimum limit remains unattainable, replace the head.

11-1826A. TIME-CODED GENERATORS.

11-1826B. GENERAL. A time-coded generator is a precision timer or generator, used for performing precise timing with recirculation registers.

11-1826C. RECIRCULATION REGISTER. A recirculation register is a dynamic-serial storage register as shown in part A of figure 11-273A. The recirculation register can also be a magnetic drum track with two heads spaced a given distance apart as shown in part B of the figure, or it can be used as a delay line as shown in part C of the figure.

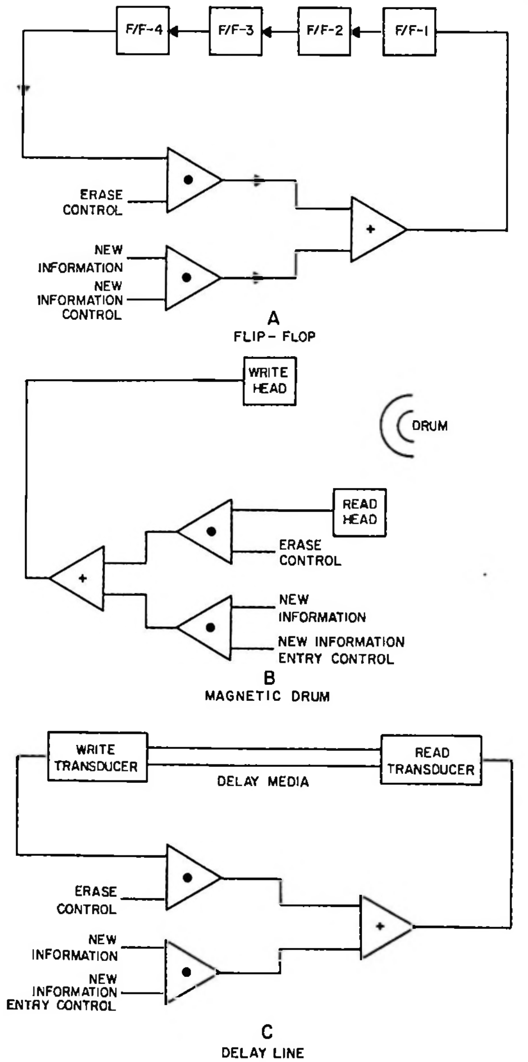


Figure 11-273A. Recirculation Register

A representative portion of the output of all recirculation registers is coupled back to the input. Therefore, when you read information at the output of the register, the information is also replaced in the register.

11-1826D. New data is written into the computer by inhibiting the recirculation path of the stored data and substituting the new data. The new data (digital words) being stored is normally made the same length as the length of the recirculating register work, or as multiples of that length. Bit one (1) of the stored information will always appear at the bit one (1) time of the new word being stored. Therefore, if the register is made shorter than the new word, the information in the register will rotate forward.

11-1826E. PRECESSION. Precession of recirculation registers is a situation that occurs when stored information appears at the output in advance time. In situations where the register is foreshortened by one bit, you can assume that the word length is 10 bits long and that the register is 9 bits long. The largest usable word that can be stored is 9 bits long. Therefore, the largest usable word which can be stored is 9 bits long. During word time one, as illustrated in part A of figure 11-273B, the data enters the register and bit one of the word is available at the output of the register at bit 9 time. At bit one time, bit number 2 of the stored word will be at the output. After 2 word times have passed, bit number 3 will appear at bit one time. In other words, the stored information is going forward with respect to the register clock, and this phenomenon is called forward precession. If the register is made longer than the standard word, as illustrated in part B of figure 11-273B, the information will go backwards with respect to the register clock, and this phenomenon is called backward precession.

11-1826F. As an example of typical register operation, assume a situation where the digital word is 000 000 0001, you will find that from the time of entry into the register it will take 10 word times for the "1" of the word to appear in bit one time. If the bit one time of each word is examined, and a pulse is put out every time that the "1" of the digit word is sensed, then the output

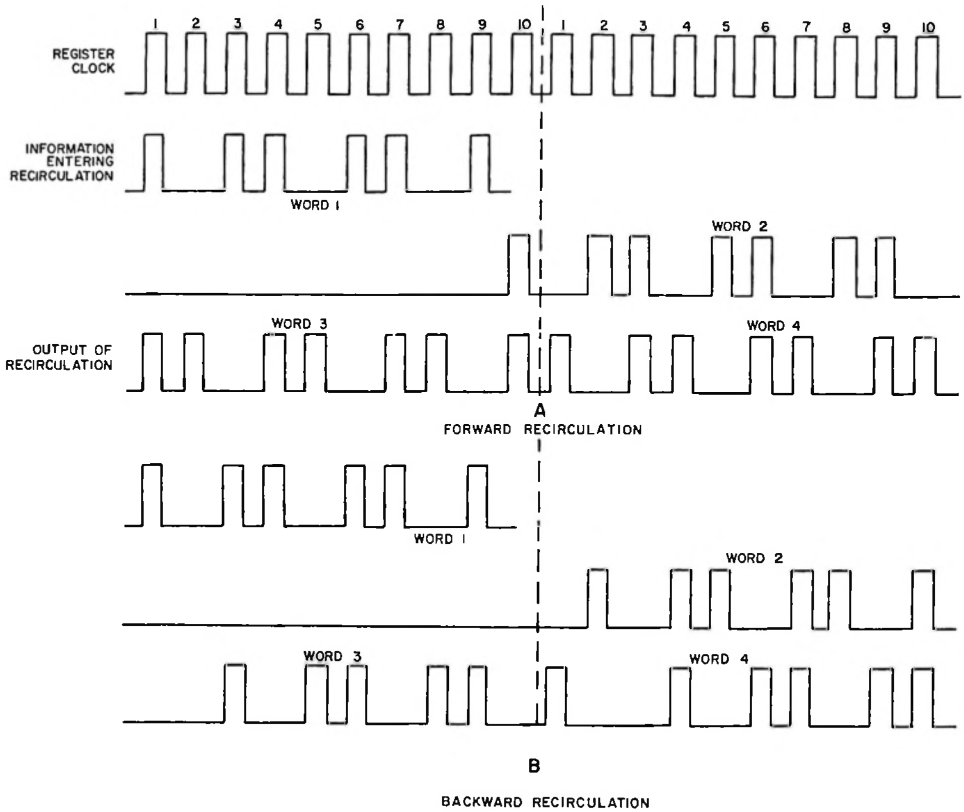


Figure 11-273B. Precession of Recirculation Register

will have one pulse every 10 word times.

11-1826G. To increase the periods between pulses, you may employ one of several methods. The first method that you may use is to increase the length of the word. Since the elapsed time between pulses is a multiple of the word length, the larger the number of bits in the word, the longer the time. However, this is not always practical. There are many cases where the word length is already fixed.

11-1826H. Another method that you may use is to increase the length of the register. If the register is made twice as long as the word length, it will take the single "1" 2 J word times to appear at the output in bit one time; here the letter J expresses the number of bits in the word. This relationship is linear, and therefore you can determine the elapsed time between pulses from the formula

$$T = NJP$$

where N represents the number of word lengths of the register, where J expresses the number of bits per word, and where P is the time period of the register clock.

11-1826I. If the time required should be less than NJP, there are two methods which can be used to obtain this time. You can change the precession rate by foreshortening the register by 2 or 3 bits instead of by one bit. However, foreshortening the register may not be desirable if the delay is to be programmable.

11-1826J. Another method that you may use to obtain an elapsed time less than NJP is to reduce the number of revolutions needed to reach completion by changing the initial digital word. Basically, you are circulating the word 000 000 0001 and precessing one bit per word time. If the initial word had been 000 0000 100, there would be 7 word times instead of 10. Thus, by controlling the starting point of the bit, you can control the measured time. Such a method requires the use of precision techniques for developing time markers.

11-1826K. Another method of developing time markers is the use of a recirculation register strictly as a storage register. A recirculation register is a series device; therefore, a full word time is consumed in examining the contents of the register shown in figure 11-273C. This illustration provides a diagram of an adder which is only capable of adding one to the number stored in the recirculation register. The addition of one is accomplished by presenting the carry flop to a one at the end of each word time. The equation for the operation is: T = NJP. The operation is as follows: The 2's complement of N is entered into the register. As the number is circulated in the register, one is added to the number each word time. A detector senses that the word has reached zero indicating that the required time has been reached. The major advantage of this technique is that the maximum time which can be measured is quite large even with a small register. The maximum time is defined as $2^J J_p$, where J is

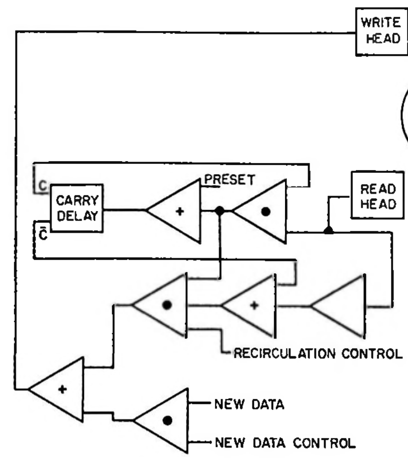


Figure 11-273C. One Bit Adder

the number of bits per word and p is the clock period.

11-1826L. Unless the time required is exactly a multiple of the word time, the accuracy of the time marker will be $\pm 1/2$ word. For most applications, this is sufficient, but for many it is not. To remedy this condition the recirculation register can be used to generate a marker to the nearest multiple of a word time which is shorter than the time required. The output would then be delayed the required additional time using a standard delay line.

11-1826M. A time-coded generator is a device that lends itself to the functions of elapsed-time methods, and in addition, stores the constants and eliminates the need of an additional delay line to obtain the resolution of one bit.

11-1826N. OPERATION. The basic concept of a conventional torsional-mode delay line, as illustrated in figure 11-273D, is composed of a pulsing coil surrounding a magnetostrictive material. When a signal

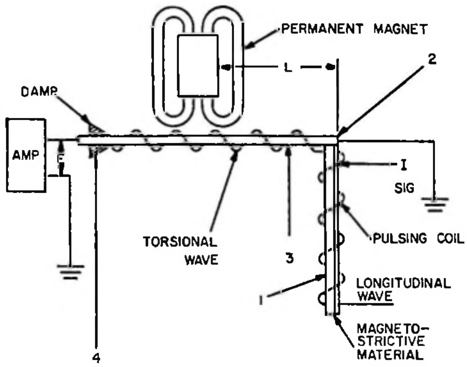


Figure 11-273D. Conventional Torsional-Mode Delay Line

is generated and applied to the coil, the mechanical stress on the wire directly beneath the coil changes, thereby inducing a longitudinal shock wave that generates downward at point 1 in the figure. However, at point 2, the longitudinal shock wave is transformed into a torsional shock wave. This torsional shock wave now generates towards point 3, and is finally absorbed by the damping process at point 4. When the permanent magnet is placed in the circuit, the torsional wave intersects the magnetic field, and a voltage E is produced. The occurrence time of the voltage waveform is approximately equal to the delay time/unit length of the shock wave at point 3, also a delay which is kept deliberately small occurs at point 1. If a second magnet is placed along the line, another voltage pulse will be generated by the same acoustic wave, and the 2 pulses will appear at the output of the amplifier in serial form. However, if you place a single magnet across the amplifier instead of across the line, the only voltage at the output will be the voltage generated by the single magnet, no matter how many other magnets are present on the line.

11-1826O. The advantages of a conventional torsional-mode delay line are: accurate variable frequency generation, pattern generation, adjustable length dynamic storage, constant storage, target simulation, frequency multiplication, frequency division, and accurate delay time. All applications can be efficiently mechanized because of the ease of adjustment, the relatively long delay per unit length of line, and the extreme constancy of delay with temperature changes.

11-1826P. A self-contained, periodic time-coded generator can be designed by dividing a section of the line into several zones, each representing one bit of the register clock and by placing an amplifier across the entire section. A serial word can now be generated if you place a magnet in the zones where a "1" appears in the word. You should now place an extra magnet on the line outside the section where the "1" appears, and also place an amplifier across the section where the "1" appears. The conventional-mode delay line coupled with the additional magnet then becomes a self-contained periodic time coder.

11-1826Q. To obtain the resolution to one bit time, place a double-width magnet on the line at the required point. The double-width magnet has the following property: the pulse generated is twice the width of the standard, therefore, the pulse can be separated from the others. After reaching the required multiple of words, the line is pulsed and the output is taken from the double-width magnet. A number of constants could be stored if the length of the line is longer than the basic word time. These constants could then be operated either simultaneously or sequentially, providing a number of precise time-coded generators from a given reference.

11-1826R. A circuit employing four magnets with spacing corresponding to a pulse period of T_1 seconds, is illustrated in figure 11-273E. The arrangement of the circuit is such that it serves as the equivalent of a tuned circuit, giving maximum output only when an impulse rate corresponding to T_1 is applied to the input terminals. A pulse rate equal to $1/T_1$ results in an output that is 4 times the normal signal amplitude. Any deviation causes a rapid drop-off in output amplitude approaching a minimum of unit amplitude. Therefore, various harmonics and subharmonics of T_1 cause an output signal amplitude between one and four times nominal. The effective Q is increased by decreasing the effective output pulse deviation, or by using a greater number of magnets with a spacing that corresponds to the period of the selected pulse rate, resulting in an exceedingly high Q and consequently resulting in a narrow bandpass configuration.

11-1826S. A conventional torsional-mode delay line is capable of readout without the use of an external permanent magnetic field. It is possible to magnetize a spot on the line surface, and to effect readout via the intersection of the torsional wave and the field. This arrangement results in a somewhat smaller output signal (approximately 50 per cent of normal output), which is, nevertheless, of usable quality. The lines, as used in this mode of operation, require special treatment to maximize their magnetic retentivity. The great advantage of this concept is the simplicity of line construction which is permitted where fixed constants are involved. The basic torsional-mode delay line then takes on the geometry of a time-coded generator.

11-1826T. Combination lines are possible when the same line stores constants which can be read out nondestructively at the standard torsional-mode delay line output, and when the same line functions as a time-coded generator with its isolated fixed output. A line of this type serves in a dual capacity, both as a serial-constant store and as an interim-serial data store, as illustrated in figure 18-273F.

11-1826U. Ability to program time-coded generators is provided by the use of electromagnetic control tops. The fixed permanent magnets are replaced with small electromagnets which generate an impinging magnetic field when electrically excited. The specific number of control tops can be altered without physical removal or repositioning the magnetic top assembly shown in figure 18-273G. This illustration is a schematic of a simple ten-bit pattern time-coded generator.

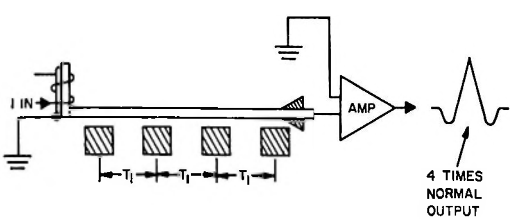


Figure 18-273E. Tuned Circuit Equivalent

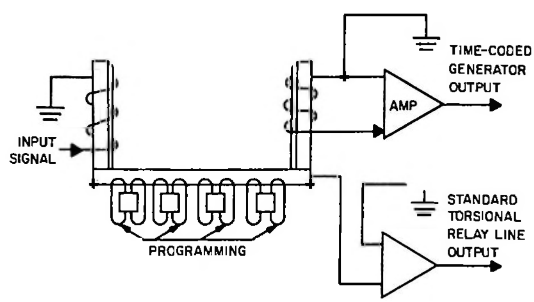


Figure 18-273F. Dual Capacity Timer

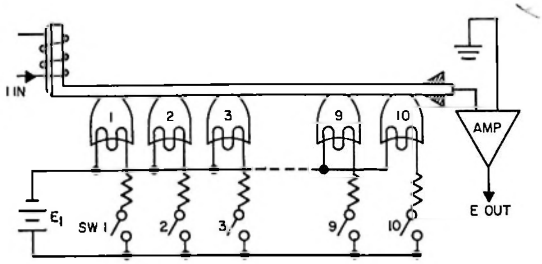


Figure 18-273G. Ten-Bit Time-Coded Generator

11-1826V. MAGNETOSTRICTIVE DELAY LINES. Magnetostrictive delay lines produce very long delays. These types of delay lines are constructed of a transmitting coil and a receiving coil, placed on opposite ends of a magnetostrictive rod such as nickel. When the transmitting coil is pulsed, the nickel contracts and a rarefaction is projected down the rod to the receiving coil. The delay time is based upon the pulse speed of the material times the length of the material. Ultrasonic delay lines are designed in the same manner, except quartz is used as the magnetostrictive rod as opposed to nickel.

11-1826W. Magnetostrictive delay lines have been designed to operate at a bit rate of 300 kHz. This type of magnetostrictive delay line consists of 7 delay lines, each having a delay of approximately 212 microseconds. The drivers and amplifiers are transistorized, and there is a 3.3 microsecond recirculating delay line that serves the purpose of a clock.

11-1826X. Ultrasonic delay lines using magnetostrictive principles for operation have been designed to supply lines with a delay range of 5 to 3000 microseconds. Delay ranges from 3,000 to 12,000 microseconds are also available, but must be designed to order. Adjustments on these types of delay lines are available to 8 microseconds.

11-1826Y. DISTRIBUTED CONSTANT DELAY LINES. Distributed constant delay lines are constructed of a coil wound glass or ceramic rod that has been coated with silver. Inductance is obtained from the coil, and capacitance is obtained from the dielectric area between the wire and the silver coating which is grounded.

11-1826Z. Subminiature lumped constant delay lines are designed for use in missile, airborne, and commercial computers. These types of delay lines have a delay time of 2 microseconds, a rise time of 0.22 microseconds, an impedance of 1,000 ohms, and an insertion loss of 0.001 db maximum.

11-1826AA. GLASS ULTRASONIC DELAY LINES. Glass ultrasonic delay lines have been designed that exhibit an information rate of 3 to 30 megabits per second. Delay times as high as 150 microseconds are possible. Glass ultrasonic delay lines offer the advantages of speed, stability, and simplicity.

11-1827. PREVENTIVE MAINTENANCE.

11-1828. GENERAL.

11-1829. Preventive maintenance for computer equipment is basically the same as for other electrical equipment; it consists of inspection, cleaning, preservation, lubrication, and performance checks. However, a unique feature of computer preventive maintenance is that you can use reliability programs to quickly determine whether a computer is operating properly. In some computers marginal checks can be controlled by the reliability program, or performed manually to detect deterioration of computer parts before a failure occurs.

11-1830. WIRING.

11-1831. Because of the extensive amount of wiring in large computers, you should carefully inspect wiring and cabling. Check the wires for loose or broken lacing, and frayed insulation. Check the cables and connectors for improper placement that might subject them to strains or kinks. Inspect the connectors and wires for burned or charred parts, dirt, cracks, and breaks.

11-1832. RELAYS AND SWITCHES.

11-1833. INSPECTION. Inspect relays to see that they are securely mounted, that the contacts are not pitted, that the springs have sufficient tension, that the armature does not stick, and that there are no signs of overheating and corrosion. You can check the action of the armature by operating the relay manually. Relay contacts can be examined with the aid of a flashlight and a mirror. Avoid bending the springs, and do not open sealed relays.

11-1834. Inspect switches for loose mountings and connections. Examine the contacts for dirt, pitting, and corrosion. Test the action of the switches and see that they operate without binding. In gang and wafer switches, see that the movable blade makes good contact with the stationary member. Make sure that the stationary contact leaves spread as the movable blade slides into them. Some switches have contacts that are impossible to reach without damaging the switch assembly. Check these switches for defective mechanical action and looseness of mountings and connections.

11-1835. CLEANING. Clean the exteriors of relays and switches carefully by blowing away the dust with approximately 5 psi of air pressure. If the connections are dirty, clean them with trichlorethylene. If it is necessary to remove covers from relays or switches, make sure that no dirt, lint, or other undesirable material is present that might get into the contacts. You can remove dust and lint from a switch or relay with a soft-bristled brush.

11-1836. There are various types of contacts in the relays used in a computer. For cleaning purposes they are divided into hard-surfaced contacts, made of palladium or platinum and soft-surfaced contacts, which are either silver or silver-plated.

11-1837. Clean hard-surfaced relay contacts, when necessary, by using a flat blade in a burnishing tool. Clean the blade by wiping it with a lint-free cloth moistened with trichlorethylene, and then move the blade two or three times between the relay contacts to brighten them. Open contacts can be pressed gently together with the fingers or an orange stick to apply pressure against the blade of the burnisher. Closed contacts will usually apply enough pressure against the blade of the burnisher. Therefore, you can open the relay armature manually, insert the blade of the burnishing tool between the contacts to be cleaned, release the relay armature, and burnish.

CAUTION

Avoid excessive burnishing. When too much of the contact metal is removed, the contact movement is altered and readjustment is necessary. Separate contacts carefully, when necessary, and use care never to bend the springs.

11-1838. If burnishing does not correct the contact trouble, deposit a few drops of trichlorethylene, with a toothpick, on the contacts. Before the solvent has had a chance to dry, add a few more drops to flush away dirt loosened by the first application. Allow the solvent to dry on the contacts, and then burnish the contacts as described above to remove any remaining residue. Always burnish the contacts after cleaning with trichlorethylene.

11-1839. Clean solid-silver contacts by inserting a strip of hard-surfaced bond paper between the contacts and then withdrawing the paper while pressing the contacts lightly together. Repeat with fresh strips of paper until the contact surfaces are clean. If this does not adequately clean the contacts, apply trichlorethylene as described above, and again polish them with a paper strip.

11-1840. Silver-plated relay contacts are ordinarily not cleaned. However, if such maintenance is necessary, they are cleaned in the same manner as for solid-silver contacts. Extreme care must be taken not to wear away the thin silver plate.

CAUTION

Never use newspaper or any soft paper, emery cloth, or highly abrasive material such as coarse sandpaper to clean relay or switch contacts.

11-1841. MAGNETIC DRUM UNITS.

11-1842. To eliminate noise which may be generated by the rotating drum, you must periodically inspect the static-grounding brush that is mounted at the end of the drum assembly. If the brush wears to less than half its original length, you should replace it. After a brush has been replaced, check to make sure that there is low resistance between the end plate of the rotor and the drum unit. This check will assure you that the brush is seated properly in the holder and is making good electrical contact.

11-1843. The drum rotor, drive motor, bolt, and pulleys are all subject to mechanical wear during normal operation; therefore, they must be regularly inspected and maintained. Check the drum drive motor pulley for tightness. An indication of a loose pulley is an accumulation of belt rubber on the pulley and pulley guard. If the pulley is allowed to remain loose, the belt will be chafed and its life shortened.

11-1844. TAPE DRIVE UNITS.

11-1845. Many digital computers include tape handling equipment. This equipment must be cleaned, lubricated, and periodically checked to ensure proper operation. Daily maintenance includes cleaning the head assembly at the beginning of each day's operation and running a short reliability maintenance program, if the computer has this feature.

11-1846. Weekly maintenance may include the cleaning of items such as the friction drive clutch and the air filters. You should check the machine's forward and reverse transfer time, the high-speed rewind, creeping of the tape reel, and the tape break circuitry. You may also be able to check the ability of the equipment to reproduce test pulses, to measure the tape speed, and to measure the moving coil and erase coil currents. A more comprehensive reliability maintenance program than is used for the daily check can then be run to determine whether error-free operation can be achieved for a period of 15 minutes.

11-1847. MAGNETIC TAPE.

11-1848. Dust and dirt can reduce the intensity of the reading and recording pulses by increasing the gap between the tape and the head. Therefore, you should take care of tape in the following manner:

a. Keep tape in a dust-proof container whenever it is not in use on a tape unit.

b. While the tape is on the machine, keep the container closed and store it where it is not exposed to dust or dirt.

c. Store tapes in some type of cabinet that is elevated from the floor and away from sources of paper or card dust. This should minimize the transfer of dust from the outside of the containers to the tape reel during loading and unloading operations.

d. Do not use the top of tape units as a working area. Placing materials on top of the units may expose the tape to heat and dust from the blowers in the unit.

e. When identifying tape reels, use a material that can be removed without leaving a residue. Adhesive stickers, that can be easily applied and removed, are satisfactory. Never alter a label identification by means of an eraser, since particles from the eraser may come in contact with the tape.

f. Inspect tape containers periodically and remove any dust by washing with a regular household detergent.

g. When necessary to clean tape, gently wipe it with a clean, lint-free cloth moistened with an approved cleaning fluid, such as trichlorethylene.

11-1849. Recorded information normally comes very close to the edge of a tape. Therefore, for proper operation, the edges of the tape must be free from nicks and kinks. To accomplish this, observe the following precautions:

a. When you remove a tape reel from a recorder, handle the reel near the hub whenever possible. If there is resistance in removing the reel, press it from the rear with your hands as near to the hub as possible. Under no circumstances should the reel be rocked by grasping the outer edge.

b. Avoid throwing or dropping reels, and do not make contact with the exposed edge of the tape. Dropping a reel can easily damage both the reel and the tape. The use of a reel and tape after the reel has been dropped is usually unsatisfactory. Therefore, never throw or mishandle reels, even while they are in their containers.

c. When mounting reels on the recorder, push them firmly against the stop on the mounting hub to ensure good alignment.

d. When placing the tape on the takeup reel, carefully align the tape to prevent damaging the edge on the first few turns.

e. If a tape break occurs, wind the resulting pieces onto two smaller reels. Splicing is not recommended unless it is necessary to make a temporary splice to recover information.

11-1850. Magnetic tape is sensitive to changes in humidity and temperature. Recommendations for tape storage are as follows:

a. If possible, store the tape in the computer room where it is to be used. Location of tape storage near the tape drives reduces both handling and variations in atmospheric conditions.

b. The storage area should be kept at a temperature of $65^{\circ}F$ to $80^{\circ}F$ and at a relative humidity of 40 to 60 percent.

c. If the tape must be removed from the computer room, you can hermetically seal it in a plastic bag to reduce the effect of temperature and humidity changes on the tape's physical dimensions. If the tape is not hermetically sealed, before it is reused it should be allowed to remain in the computer room for a length of time equal to the time it was out of the computer room. If the tape is out of the computer room for a period longer than 24 hours, it should be conditioned to the computer room for 24 hours before being reused.

d. For long-term storage, enclose the reel and container in a hermetically sealed plastic bag. Store in an area of constant temperature. Either cold or hot temperature can harm tape. A temperature between $40^{\circ}F$ and $120^{\circ}F$ is satisfactory.

11-1851. TROUBLE SHOOTING.

11-1852. GENERAL.

11-1853. Computer circuits such as multivibrators, clock oscillators, and gates can be tested in the same manner as similar circuits in radar, radio, and multiplex carrier equipment. Test equipment such as oscilloscopes, voltmeters, ohmmeters, frequency meters, tube testers, and transistor testers can be used to determine which parts are actually defective and, if possible, what has caused the failure. However, large computers have such enormous numbers of circuits that efficient trouble shooting to locate defective circuits can be accomplished only with the aid of maintenance programs, built-in test equipment, and special test sets designed to be used with specific computers.

11-1854. In computers which feature maintenance programs, reliability checks may be used to determine the area of a computer in which a failure has occurred. You can then run diagnostic programs to localize the trouble to a group of circuits or plug-in units. For example, if trouble appears to be in a plug-in matrix assembly, the suspected matrix assembly can be replaced by a unit known to be operating correctly. If this results in proper computer operation, you may then be able to test the defective matrix assembly with the aid of a matrix test set to locate the defective parts. If a special matrix test set is not furnished for the computer, the assembly must be tested with general-purpose test equipment to find the defective parts.

11-1855. Test equipment such as test-pattern generators, trouble indicators, and memory units are furnished with some computers as built-in equipment. Test-pattern generators can be used for signal-tracing applications or to provide a simulated signal for maintenance purposes. Visual and audible alarms are provided to indicate errors in calculations or the existence of conditions that might cause damage to the computer. Neon indicators are often connected to flip-flop stages to indicate the status of registers. By observing the indicators and alarms, you can often decide what action should be taken to correct a malfunction.

11-1856. MEMORY UNITS.

11-1857. Trouble within a memory unit can usually be located by means of diagnostic maintenance programs. An example of a trouble-shooting procedure is as follows:

a. Apply power to the computer and prepare the computer for test mode operation.

b. Check the supply voltages to the memory units; if they are not set properly, adjust them to their correct values.

c. Determine which memory unit is faulty by analyzing the computer error indication, program error, or error printout.

d. Run a diagnostic maintenance program. When an error printout occurs, check the computer condition indicators to determine in which routine the failure occurred.

e. Set the maintenance memory unit to repeat the routine which produced the first error. Run this routine to make sure that the error will occur when the routine is run separately. If the test routine does not produce the error when run separately, set the maintenance program to the previous routing so that program operation will switch to the first error printout.

f. If the trouble is intermittent, you can use the marginal-check version of the program to apply prescribed margins to the memory unit. If the original memory failure is detected under marginal-check conditions, verify the failure margin by applying a manually controlled voltage excursion to the tested marginal unit while running the routine.

g. Check the memory unit for the remaining errors that were detected by the maintenance program. This is accomplished in the same manner as in steps e and f.

h. Analyze the error printouts to determine whether the error is a bit failure or a selection failure. If the error is a bit failure, determine which bit is causing the failure, and check the associated sense amplifier or digit plane driver. If the error is a selection failure, further analyze the printout to determine which selection line or lines are causing the failure, and check the associated driver stage.

i. Refer to the equipment technical manual trouble-shooting guides to determine which circuits should be tested using an oscilloscope. Calibrate an oscilloscope to make the necessary voltage and timing measurements. Then rerun the maintenance routine and observe the action of the suspected memory circuits.

11-1858. MAGNETIC DRUM UNITS.

11-1859. GENERAL. The general procedure for trouble-shooting the magnetic drum units of a computer is to verify the existence of a failure and determine its source and cause. Maintenance programs are the chief means of determining the particular areas in which malfunctions have occurred. However, in equipment where maintenance programs are not effective or not available, manual testing must be performed at the computer maintenance console or directly at the magnetic drum units. Marginal-checking procedures are applicable for trouble-shooting intermittent malfunctions.

11-1860. MAINTENANCE PROGRAMS. If the computer has this feature, you can run a diagnostic program whenever a drum unit is not functioning correctly. A drum unit maintenance program can be used to check the fields, status controls, disconnect counter, field switch, and information circuitry. To make a test, you place the drum unit in the computer-test mode and initiate the maintenance program which is stored on cards or magnetic tape. The program first clears the drum fields, then writes a test pattern. The computer reads and checks information transfer. Errors are indicated if the program halts before the program is completed or if incorrect information is printed out. Maintenance tables are normally furnished with the equipment technical order as aids to determine the source of trouble from the results of a maintenance check.

11-1861. WRITE-READ CHECK. The bits of information that are recorded on a magnetic drum by a test pattern may be displayed by lamps on a check register in the drum unit. These indicators are used in conjunction with tables in the equipment technical manual to help you analyze the cause and location of a malfunction. Such tables are arranged according to the types of bit errors which may be encountered; ie, missing bits common to two or more fields on one drum, a missing bit in one field, or two adjacent bits in error in one field. When one of these conditions is present, the corresponding table should be used to locate the trouble within a defective circuit or device. An arrangement of computer circuits for testing a magnetic drum is shown in figure 11-274.

11-1862. Several test patterns should be used individually or in combination to eliminate the possibility that an error may go undetected or be misinterpreted. Examples of such test patterns are as follows:

a. Straight "0" bit pattern.

b. Straight "1" bit pattern.

c. "0" and "1" bits alternating in the first register and complementing in the following register; ie, first register, 010101, second register, 101010, etc.

d. All "1" bits in the first register, all "0" bits in the second register, etc.

11-1863. A test pattern applies a "1" or "0" bit to each write head for the purpose of receiving identical information from the corresponding read head. The use of one test pattern may not prove sufficient, in all cases, to thoroughly identify a trouble. You may then require one pattern or a combination of two or more patterns to locate the trouble which causes bit errors. For example, a straight "0" bit pattern, which is applied to the drum write heads, may appear as all "0" bits on the check register, and yet one of the read heads may be defective and produce only "0" bits. The use of a straight "1" bit pattern, in this case, will reveal the particular read head which is at fault. You must then make further checks to determine whether the failure is caused by a defective head, a misadjusted head, or the read amplifier stages that are connected to the head. If the read head and corresponding amplifier stages are operating correctly, you should test the corresponding write head and its associated circuitry.

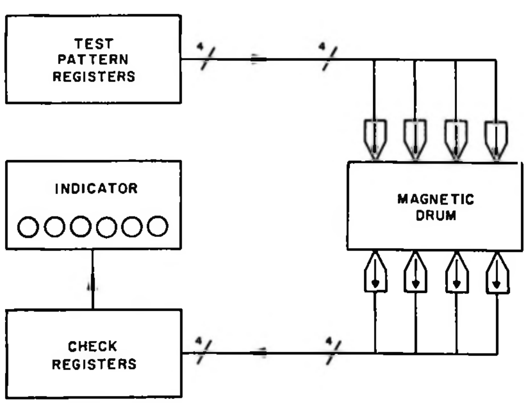


Figure 11-274. Magnetic Drum Write-Read Test, Block Diagram

11-1864. RUNOUT TEST. Excessive bearing wear, unbalance, and ovalness of a drum rotor can be checked by means of a runout test. To perform this test, connect an oscilloscope to a read head as shown in figure 11-275. Operate the drum motor at its normal speed, and apply an all "1" bit test pattern to the write head. The read head waveform should resemble the waveform shown in figure 11-276. For the drum to be considered satisfactory, any change in the amplitude of the test pattern bits, V_{MAX} minus V_{MIN}, must be less than 20 percent of the maximum bit amplitude. This measurement should be made from read heads located at each end of the drum, and also from a read head at the approximate center of the drum.

11-1865. NOISE TEST. You can determine whether noise is being generated by a magnetic drum by using an oscilloscope in the same manner as in the runout test. For this test, disconnect the test-pattern gen-

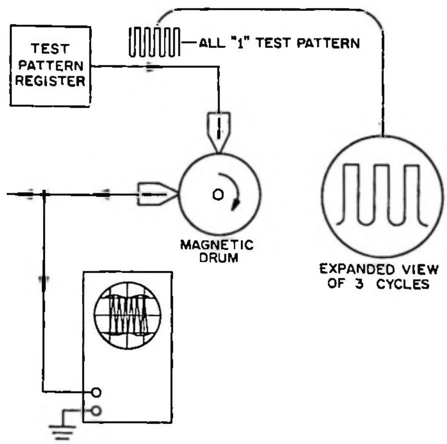


Figure 11-275. Magnetic Drum Runout Test, Block Diagram

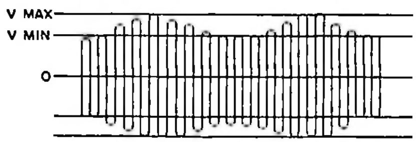


Figure 11-276. Runout Test Waveform

erator from the write head, and erase all signals from the drum. Then start the drum rotor and observe the display on the oscilloscope. The oscilloscope display may resemble the waveform shown in figure 11-277. Any voltage that is present is a result of residual electrical noise or spurious magnetization of the drum rotor. This test must

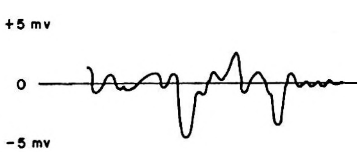


Figure 11-277. Noise Test Waveform

be performed for each recording track on the drum rotor. You will have to refer to the equipment technical manual to determine the noise spike amplitude that can be tolerated and the number of tracks that can be noisy before the drum rotor must be replaced. For the example shown in figure 11-277, 5 millivolts is considered to be the maximum noise pulse amplitude that can be permitted.

11-1866. CROSSTALK TEST. A crosstalk test determines whether each channel processes data free of interference from other channels. To perform this test, erase all information from the track to be checked, and connect the read-head output to the oscilloscope. Then operate the drum and apply test patterns to all channels except the channel being tested. Use a calibrated oscilloscope to measure the output noise level due to crosstalk. This test must also be performed for each track on the drum. The noise level and number of noisy channels that can be tolerated will be listed in the equipment technical manual.

11-1867. MAGNETIC TAPE UNITS.

11-1868. The trouble-shooting of computer magnetic tape units can also be accomplished by means of diagnostic maintenance programs. Such a program may use the control units, memory units, line printer, and marginal check equipment to test the tape units. Such a program may be composed of three test routines, as follows:

a. Routine 1 uses four different test patterns to check the timing of the write, read, backspace, and rewind instructions, as well as their operation. This routine also checks the write and read end-of-file circuits, the read zero word instruction, the disconnect circuits for the conditions either read more or read less than the number of words contained in a record, and all information transfer paths involved.

b. Routine 2 uses a random number pattern to check the timing of the read, write, and rewind instructions and their operation. This test is accomplished by checking the rewind operation, the stepping of the tape clock, the tape speed, and the effects of a number of short random-number records.

c. Routine 3 is used to check the file-protect-on circuitry.

11-1869. The routines to be run will depend on the error indication. These three maintenance program routines can be run in five different modes, as follows:

a. Mode 1 uses routine 1 only.

b. Mode 2 uses routine 2 only.

c. Mode 3 automatically applies excursions for prescribed margins or margins to failure, using either routine 1 or routine 2 depending on which marginal-check word is being used.

d. Mode 4 selects the tape drives available and uses routines 1 and 2 on each tape drive automatically. This mode is normally used for preventive maintenance and not for trouble shooting.

e. Mode 5 uses routine 3 only.

11-1870. When these routines are run, all errors will be indicated by a printout, which will include the type of error, the trouble location, and possible plug-in units that caused the failure. In modes 1, 2, and 5, errors will also be indicated by a programmed halt, with the computer accumulator designating when the error occurred. In mode 3, the marginal check words will also be printed, designating the selected line and excursion that caused the failure. In mode 4, the program will print out errors, but will halt only after all tape drives have been tested.

11-1871. REPAIR PROCEDURES.

11-1872. GENERAL.

11-1873. When the cause of a computer failure has been found by trouble-shooting procedures, you will normally perform corrective action, such as adjustment, alignment, tuning, cleaning, and replacement of parts. The replacement of computer parts involves the same basic procedures as are used for radio, radar, and carrier equipment. When replacing parts with many wire connections, such as switches, relays, connectors, and transformers, each wire should be tagged and identified. You can then remove the wires from the defective part and make the replacement without making an error in the wiring. Many computer circuits are contained in modules, which may be either plug-in units or soldered to the equipment. The extent of repair for module units will depend on what is considered to be economically practicable. The units may be repaired in field maintenance shops or sent to depots. Appropriate soldering techniques must be used when repairing computers that use semiconductors, miniature circuit parts, and printed circuits to avoid overheating the parts. The voltages used in ohmmeter measurements of semiconductor circuits must be limited to safe values.

11-1874. MEMORY UNIT WIRE REPLACEMENT.

11-1875. On large magnetic core memory units, where many planes are used, replacing a broken wire or an entire plane must be done with extreme care to avoid creating new troubles. If one of the X or Y drive lines within a memory plane breaks, the plane will normally have to be removed from the memory unit to be repaired.

11-1876. If an X or Y drive line breaks close to the terminal on a memory plane, you may be able to replace the wire without having to remove the plane. A recommended procedure is as follows:

a. Disconnect the power to the memory unit, and remove necessary panels from the cabinet or rack.

b. Double-check the suspected wire by making a continuity check with an ohmmeter.

c. Cut a new wire of the correct diameter and length, and tin the ends.

d. Remove the jumper wires, on both sides of the plane, that are connected to the same terminals as the broken drive line. Then remove the portion of the broken end of the drive wire that is wound around the terminal.

e. Have another person hold the broken end of the drive line so that it cannot be pulled into the plane. Carefully unsolder and unwind the wire on the terminal at the unbroken end of the drive line.

f. As well as possible, straighten out the portion of the wire that has been unwound from the terminal at the unbroken end of the drive line. Place the wire end-to-end with the new wire. Overlap the ends approximately 1/16 inch, and tape them to a support such as a computer card, as shown in figure 11-278. Solder the wires together to make a temporary mechanical bond, and remove the tape.

g. Smooth the solder joint with a sharpening stone until there are no ends or sharp bulges that will catch as the wire is pulled through the cores.

h. Gently pull the broken end of the wire out of the plane. If the solder joint binds within the plane, try gently jockeying the wire back and forth. Excessive pull on the

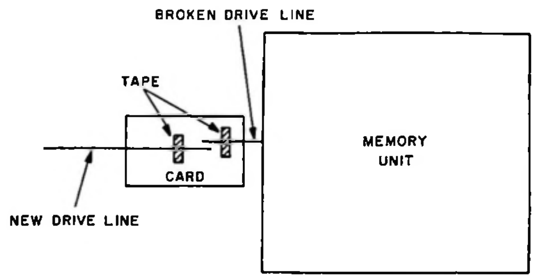


Figure 11-278. Method of Connecting Replacement Wire to Broken Drive Line

wire can cause it to break in the plane, damage a core, or damage another winding; in such cases, the entire plane must be removed. If jockeying does not help, pull the wire back and smooth the solder joint for another try.

i. After a new wire is pulled into place, clip off the old wire at the solder joint. Solder the new wire and the jumper wires, that were previously removed, to the proper terminals.

11-1877. MEMORY PLANE REPLACEMENT.

11-1878. If a memory plane is found to have a defective core or a wire having a break that cannot be reached from the outside, it must be removed and either repaired or replaced. Since this is a time-consuming job which, if not done carefully, can create other troubles, you should make every possible check to prove that the suspected plane is really defective before removing it. Make continuity checks of suspected wires, and make certain that the trouble is not caused by a poor solder connection at one of the terminals. When you are certain that a plane must be removed, proceed as follows:

a. Disconnect the power to the memory unit and remove the necessary panels from the cabinet or rack.

b. Shield the assemblies below the planes to be worked on by attaching a cheesecloth with masking tape.

c. Use a grease pencil to mark the defective plane on all four sides so that you will not make a mistake and remove any wrong wires.

d. Cut all jumper wires to adjacent planes by cutting close to the defective plane terminals.

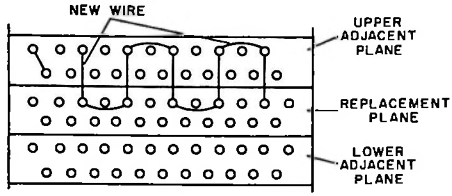
e. Remove the cut jumper wires from adjacent planes by grasping the wire with needle-nose pliers and unsoldering at the terminal. Wipe the terminals clean of surplus solder.

f. Remove any retaining screws and air-seal material from the defective plane.

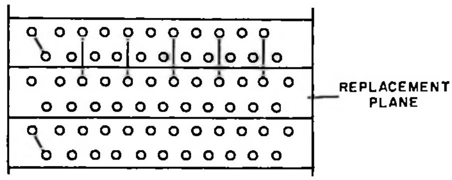
g. Make a note of the proper position for the defective plane so that you will be able to insert a replacement plane in the same position. Then slide the defective plane out of the memory unit.

h. Insert a replacement plane, replace the retaining screws, and replace the air-seal material that was removed during step f.

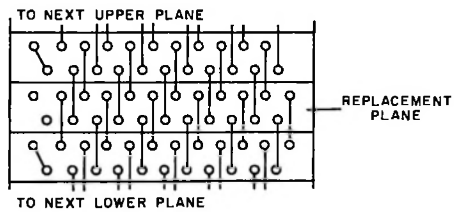
i. Replace all jumper wires that were removed in steps d and e. Figure 11-279 shows an efficient method of doing a good wiring job. Choose two horizontal rows of terminals that must be connected together with jumper wires. Connect these terminals with a single wire, as shown in part A of the figure. Keep the wire taut, and wrap it around each terminal. Then solder the wire to the terminals to be used, and cut out the horizontal portion of the wire. The plane wiring will then be similar to that shown in part B of the figure. This procedure should then be repeated for the other sets of horizontal terminals that must be connected by jumper wires. Part C of the figure shows the jumper wires reconnected to one side of the replacement plane.



A
CONTINUOUS RUN OF WIRE ALONG ENTIRE SIDE OF PLANE



B
SOLDER TERMINALS AND CLIP OFF HORIZONTAL PORTION OF WIRE



C
SIDE OF PLANE COMPLETELY WIRED

Figure 11-279. Replacement of Plane Jumper Wires

j. When all four sides of the plane are completely rewired, remove the cloth shield and check for wire scraps and solder splashes that may have fallen past the cloth. Then replace the cabinet panels and check the operation of the memory unit.

11-1879. COMPUTER APPLICATIONS.

11-1880. GENERAL.

11-1881. Isolation of computer malfunctions involves the use of a variety of techniques. These techniques include diagnostic and marginal checking routines, employment of conventional test equipment such as oscilloscopes, voltmeters, ohmmeters, frequency meters, and tube and transistor testers. In addition to the items mentioned above, special test equipment designed for particular computers is frequently used. Such special test equipment provides complete test programs for subassemblies or, as in the case of most digital computers, for the removable printed-circuit plug-in cards (or boards) making up the circuitry. Other items of special test equipment may be used to test other computer components, such as the magnetic drum and its associated circuits.

11-1882. Special test equipment may be fully automatic, and involving the use of elaborate equipment which requires a minimum of manual operation. It may also be semiautomatic, involving the use of certain manual control and response interpretation operations. In its simplest form, special test equipment may be a test jig or fixture constructed to perform an individual testing operation; such equipment is understandably of limited flexibility in its use. Test programs recorded on magnetic tape, or punched card, are supplied to the test set. In the simpler test sets, you may set up the program through manual operation of switches, in accordance with the instructions provided.

11-1883. The following paragraphs discuss the general characteristics of typical special test equipment. Most items of test equipment of this type operate in essentially the same manner, varying only in their application to specific computers.

11-1884. PRINTED-CIRCUIT-CARD TEST SET.

11-1885. GENERAL. A printed-circuit-card test set provides the testing capability for the removable digital, analog, and servo printed-circuit, plug-in cards (or boards) which compose the circuitry of most digital computers. The test set simulates the normal operational parameters of the card under test. The testing of each card circuit is performed on an individual basis. The card test set provides you with a programmed setup of test conditions, including operating voltages, input signals, loads, and test equipment connections which permit you to view normal circuit operation. Tests are provided for a variety of circuits, including flip-flops, gates, drivers, single-shots, matrices, multivibrators, amplifiers, oscillators, inverters, summation circuits, adders, converters, controls, emitter-followers, delays, and other circuits necessary for the operation of a computer.

11-1886. FUNCTIONAL DESCRIPTION. The test set employs semiautomatic operation, and is programmed through the use of 80-column IBM-type punched cards. These cards are read by a card reader device which is part of the card test set. The card reader, in turn, supplies signals to the card test set, enabling you to set up the test program for application to the printed-circuit card under test. You should progress through the prescribed test program step-by-step, observing the required indication for each step on an oscilloscope, digital voltmeter, or vacuum-tube voltmeter. Each step of the program is initiated by depressing and releasing a control switch; however, all setup operations are automatic.

11-1887. The test set consists of ten sections, as shown in figure 11-280. These sections are identified as the punched card reader, punched card reader control,

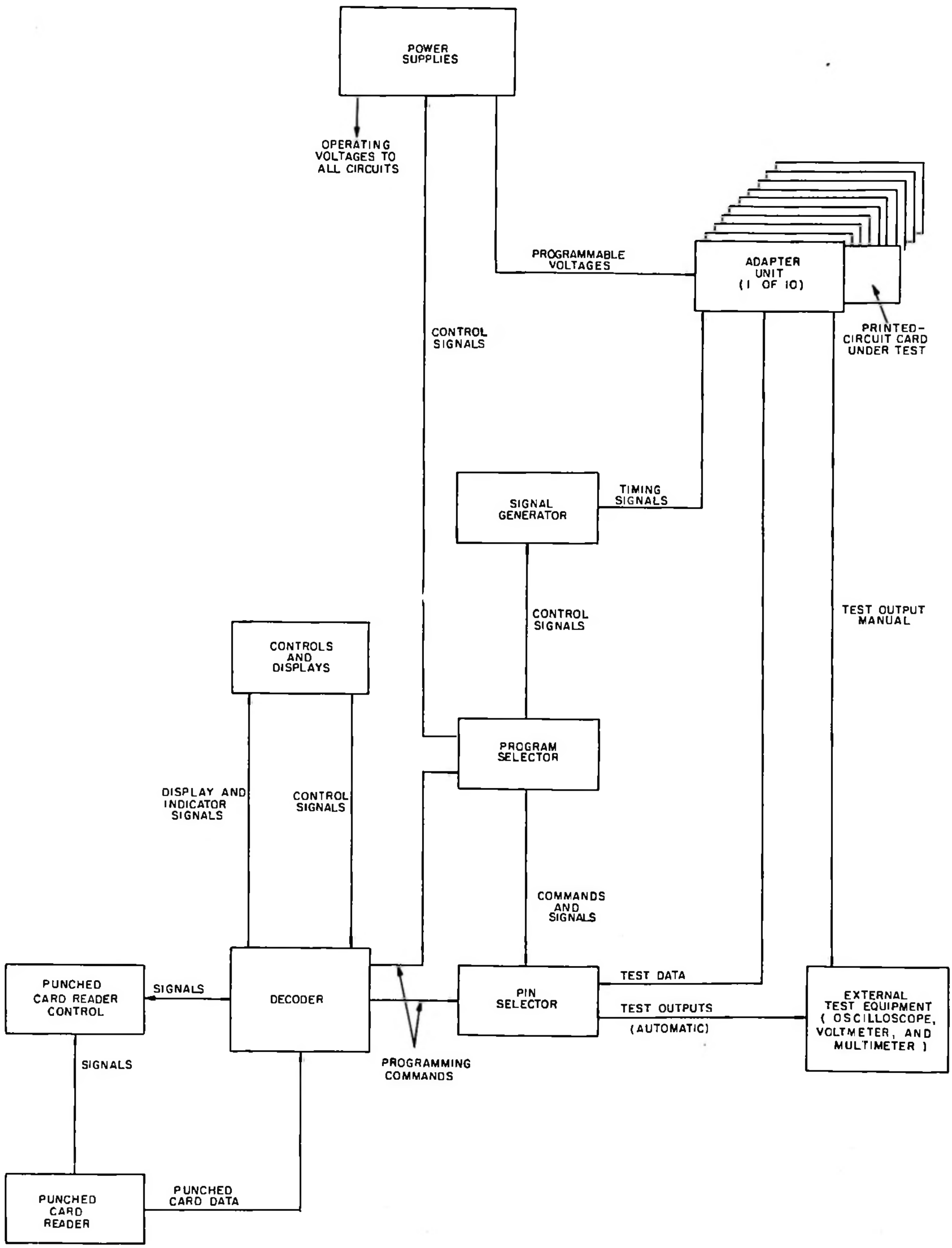


Figure 11-280. Block Diagram of Printed-Circuit Card Test Set

decoder, program selector, pin selector, signal generator, power supplies, adapter units, controls and displays, and external test equipment.

11-1888. The punched card reader provides you with the programming for the test set. Every printed-circuit card tested by the test set has one or more associated punched cards. The information contained on each punched card consists of binary coded data which is sensed by the card reader. The punched card reader control and the decoder receive signals and coded data, respectively, from the punched card reader. The punched card reader maintains control over the operation of the punched card reader and the decoder. Within the decoder, the data is decoded from binary form to a command in order to control test set operation, or to a signal used to energize a relay in one of the relay matrices of the program selector or pin selector sections. The program selector routes the signals and voltages internally within the test set, while the pin selector routes the test signals and voltages to the adapter unit in use. The signal generator produces pulses of various frequencies and pulse widths used in the test program. The power supply section provides operating voltages, along with certain programmable voltages required for testing.

11-1889. The adapter units of the test set provide interface, test circuits, and active and passive loads for an assigned group of printed-circuit cards to be tested. Only one of the available 10 adapter units is in the operating position at any one time. Each adapter unit performs the following functions:

a. Connects the output of the signal generator to the parameters (impedance and signal level) required by each printed-circuit card under test.

b. Programs the programmable power supplies for the specific voltages required by each test.

c. Provides active and passive load terminations to each printed-circuit card under test.

d. Provides convenient front panel test points which you can use with appropriate test equipment for observing the operation of the printed-circuit card under test.

11-1890. The controls and displays section controls the application and removal of power to the test set, and the sequence of the punched card progression through the punched card reader, Alphanumeric readout displays mounted on the front panel of the test set permit you to view the status of each card command; this feature also assists you in locating malfunctions within the test set.

11-1891. USE OF PRINTED-CIRCUIT-CARD TEST SET. Although the test set itself is a complicated piece of equipment, you will find that actually using it in testing printed-circuit cards is very easy. Energize the test set, but place it in a standby or non-operational status. Then select the printed-circuit card to be tested ana the associated adapter unit installed in the test set. Insert the binary punched cards containing the test program into the punched card reader. You now have the test set fully energized, and you can begin testing. Next, depress a pushbutton switch to advance the program from one test to another. You can then compare the test output indication (oscilloscope, digital voltmeter, or vacuum-tube voltmeter). When your actual results compare with those specified, the program is advanced to the next test. If the results do not favorably compare, apply trouble-shooting techniques to the defective circuit. Since all tests dynamically check the printed-circuit cards, the problem of isolating faults is greatly simplified. Accompanying test data in printed form describes the type and content of each test of the program; this information is also used in analyzing the malfunction. After you replace the defective component, test the printed-circuit card again to make sure that the trouble is cleared. Through the use of this test set, most printed-circuit cards can be dynamically checked. Those printed-circuit cards which the test set cannot check can be tested statically (non-operationally) by a conventional multimeter on a "point-to-point" basis.

11-1892. MAGNETIC DRUM TEST SET.

11-1893. GENERAL. A magnetic drum test set provides calibrated signals for aligning the magnetic drum used in computer equipment employing this type of device. The test set generates calibrated pulses which are written on the drum. After a certain time interval, the pulses are read off the drum and returned to the computer. The read and write pulses are applied to an oscilloscope used in monitoring the operation of the magnetic drum.

11-1894. FUNCTIONAL DESCRIPTION. The test set consists of four sections, as shown in figure 11-281. These sections are as follows: power control, signal generating, switching, and output.

11-1895. The power control section controls the application and removal of ac and dc voltages supplied to the test set. In addition, the power control section has protective circuits which prevent damage to the test set in the event of current overloads.

11-1896. The signal generating section generates, amplifies, shapes, and times the signals required for testing and aligning the magnetic drum. The output signals of the signal generating section for this typical test set are a single pulse, 10-kc pulses, and a 1-megacycle pulses. The desired output is selected and applied to the appropriate connectors of the output section by means of the switching section. In addition, read pulses from the magnetic drum are amplified and shaped by the signal generating section and then applied to the switching section.

11-1897. The switching section is used in selecting the available output signals from the signal generating action. In addition, an externally supplied computer trigger pulse can be applied to the output section, where it will be available to trigger the sweep circuit of an external oscilloscope.

11-1898. The output signals from the signal generating section and the externally supplied computer trigger pulse are applied through the switching section to the appropriate output connectors, where the signals are available for external use.

11-1899. USE OF MAGNETIC DRUM TEST SET. The test set is used for performing the following magnetic drum adjustments: long-channel parallel, servo loop delay, head-to-drum gap, and short-channel storage time. Since the test set is permanently connected to the parent computer equipment, the use of the test set involves no change in installation prior to use. The various controls of the test set enable the test set program to be applied to the magnetic drum via the parent computer equipment.

11-1900. SPECIAL CONSIDERATIONS.

11-1901. GENERAL.

11-1902. Analog and digital computers require special methods of trouble-shooting, which differ from the testing methods used with other types of electronic equipment.

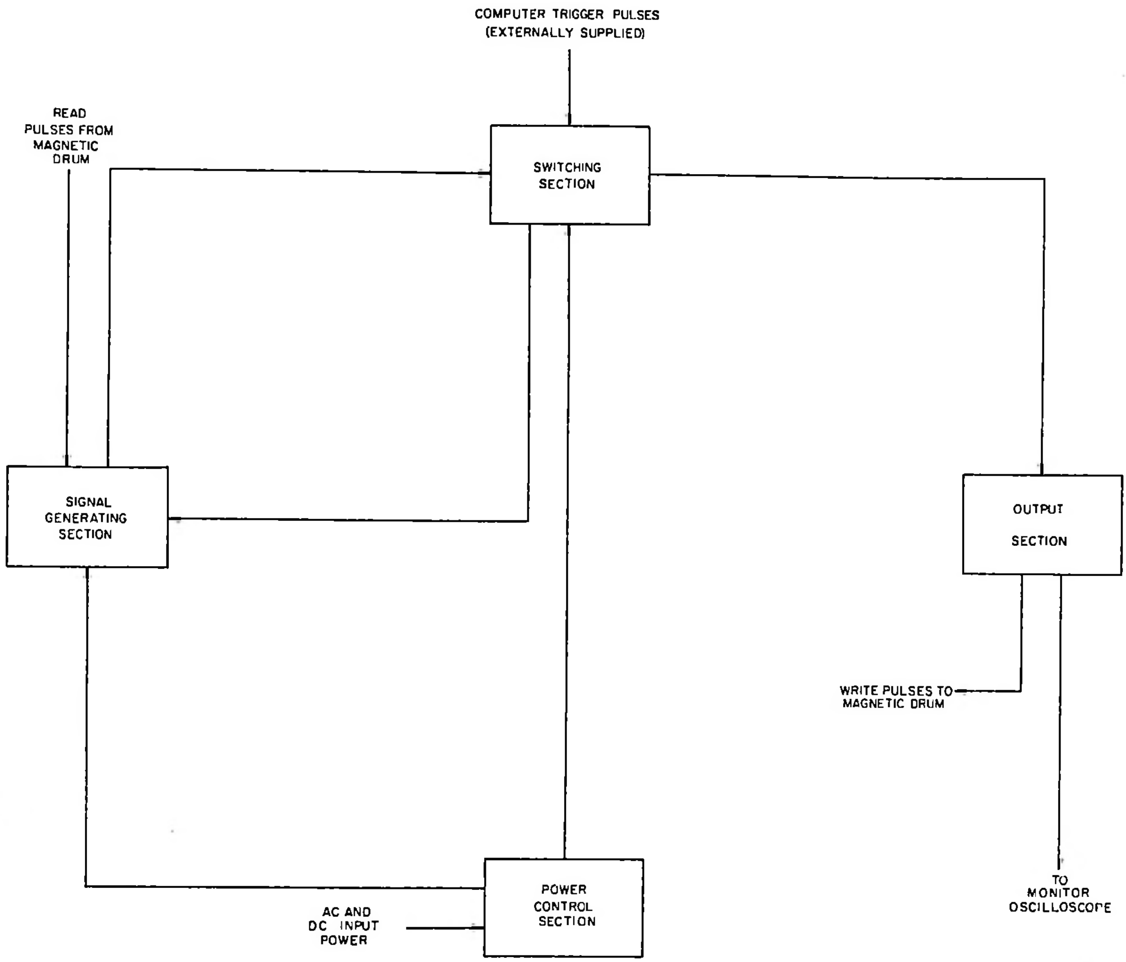


Figure 11-281. Block Diagram of a Magnetic Drum Test Set

Such special methods are contained in maintenance technical orders provided for the computer equipment. The general objectives of the methods are: (1) to isolate failures to a replaceable part (normally a component card), (2) to discuss the removal and replacement of the faulty part, and (3) to aid in the verification that the fault has been corrected.

11-1903. Analog and digital computers, for the purposes of maintenance, can be divided into the following two special categories: (1) computer equipment designed to perform arithmetic operations programmed by the operator (an equipment of this type is termed an arithmetic computer), and (2) computer equipment designed to perform a fixed program, usually as part of a facility (an equipment of this type is termed a data processor).

11-1904. Diagnostic routines direct equipments, which normally perform computer functions, to act in a maintenance capacity. The normal computing actions of some equipment may be interrupted and the equipment then programmed to test or exercise itself or associated components. The desired end result of the diagnostic routines is to detect malfunctions and isolate these malfunctions to specific areas of the equipment. Once a particular area is identified, further isolation and subsequent correction of the faulty circuit are achieved by substitution of the printed-circuit card (in equipments using cards) or replacement of the faulty component(s). Diagnostic routines apply equally to arithmetic computers and data processors. The results of the diagnostic routines are usually recorded on the computer "print-out". Subsequent analysis and interpretation are required for localizing the equipment malfunction. Once the suspected fault is localized, you can employ test methods which pinpoint the faulty component.

11-1905. Four major trouble-analysis methods (other than diagnostic routines) are available to aid you in isolating equipment malfunctions. These methods are: (1) observing the trouble or error indicators provided with the equipment, (2) use of special purpose test equipment, (3) following standard "signal tracing" procedures, employing common external test equipment such as oscilloscopes, meters, etc, and (4) substitution by the process of selective elimination of printed-circuit cards containing the suspected faulty circuit.

11-1906. USE OF FLOW CHARTS.

11-1907. GENERAL. Flow charts are usually provided to direct you in following a certain defined set of test procedures, whether it be a diagnostic routine or any other type of test program. Computer and data processing problems are characterized by extremely complex combinations of elementary steps. The operator, therefore, runs the risk of becoming bogged down in an incomprehensible mass of details. The flow chart helps to visualize the complex series of steps for this program.

11-1908. In general, there are two types of flow charts, as follows: (1) "functional", or "top-level", which defines the over-all aspects of a problem and enables you to grasp the principal technical idea of the procedure you are ab out to follow, and (2) "detailed", which breaks down the functional flow chart into individual steps, thus providing detailed instructions for performing the test program.

11-1909. TYPICAL FLOW CHART. An example of a typical functional flow chart is illustrated in figure 11-282. This chart is used in a diagnostic routine for a computer memory section. The illustration shows the main functions of the component programs and their interrelationship. The numbering of the blocks is arbitrary, and does not necessarily indicate the sequence given in a particular technical order.

11-1910. SERVICING TECHNIQUES.

11-1911. GENERAL.

11-1912. Generally speaking, servicing techniques for the hardware of computers follow those methods used for other items of electronic equipment. However, there are some special techniques which apply specifically to computers, as explained in the following paragraphs.

11-1913. REPAIR OF PRINTED-CIRCUIT CARDS.

11-1914. Current practice in computer equipment construction is to make wide use

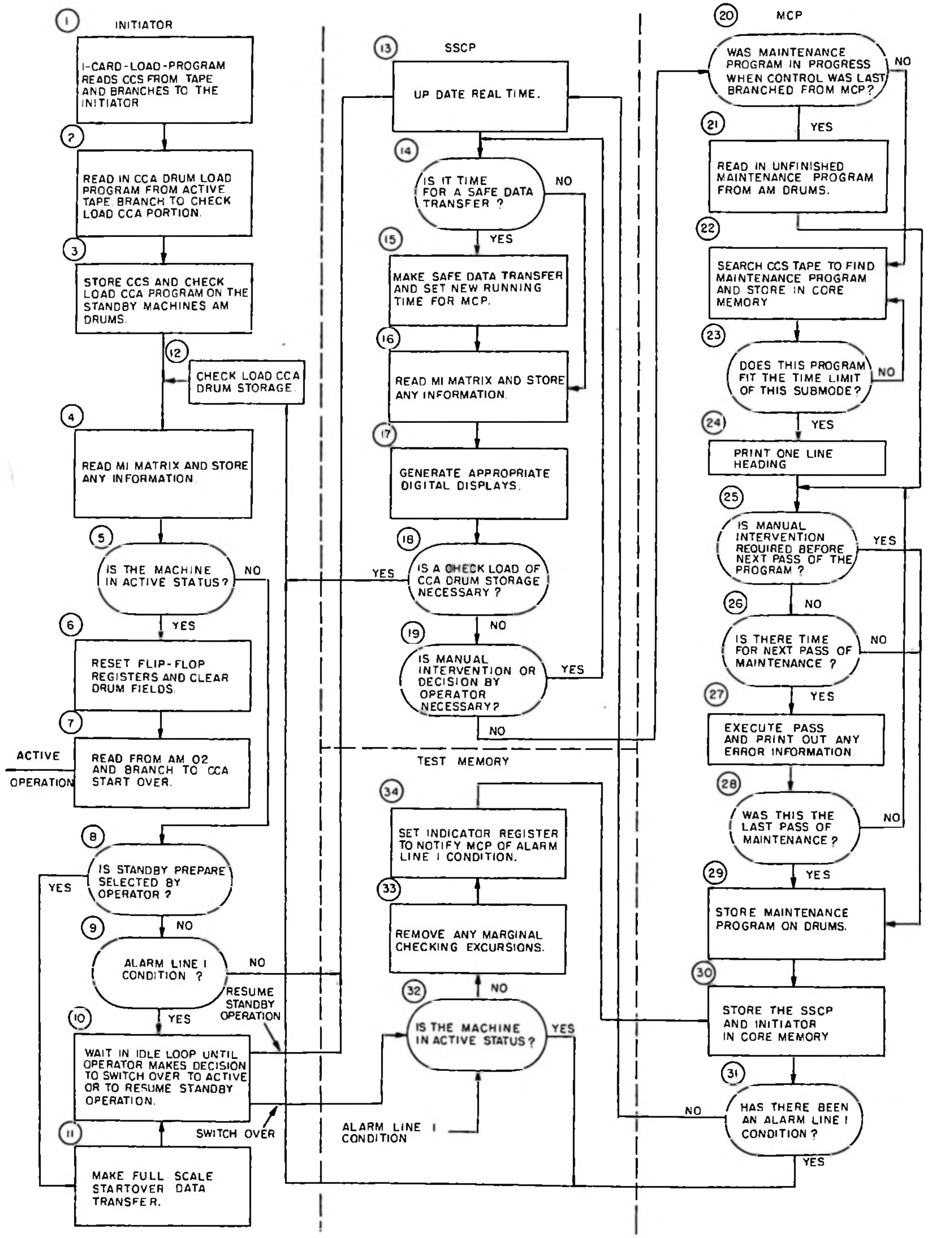


Figure 11-282. Typical Functional Flow Chart

of removable printed-circuit cards which contain most of the logic or other operational circuits. This technique keeps downtime to a minimum when a failure occurs, because the cards can be easily replaced. The circuits themselves may be composed of individual components, such as transistors, resistors, diodes, capacitors, inductors, etc; or they may be elaborate circuits which comprise many components, all of which are encapsulated into a small form. These latter circuits often appear in microminiature form, depending on the physical size of the computer equipment. The following paragraphs contain detailed instructions for repairing printed-circuit cards, installing new parts, and testing after repair. These instructions include the use of tools and materials needed for printed-circuit-card maintenance.

11-1915. PRELIMINARY INFORMATION. Printed-circuit-card repair is difficult, and requires more skill than conventional-equipment repair. Since the cost of a printed-circuit card is high, especially as compared with the cost of an individual component part, never try to save a component part at the expense of a card. Most component parts can be clipped from the card, thereby protecting the card's printed-circuit conductor (i.e., the copper foil beneath the visible solder coating) and preventing any undue component part damage. Exercise care when using a soldering iron to remove the leads of a chipped-off component part, to connect a new component part, or to service the card itself. Since printed-circuit cards are easily damaged by heat, prolonged application of heat will destroy the adhesive quality of the bonding agent that holds the printed-circuit conductor to the card. The following discussion describes acceptable procedures for replacing component parts and servicing the printed-circuit card, and identifies the tools and materials required for performing these procedures. Read these instructions carefully before attempting any printed-circuit repair. For satisfactory repairs on printed-circuit cards, the use of the following tools and materials is recommended:

a. Pencil-type soldering iron, 25 watts maximum

b. Twist drills, numbers 30 through 60

c. Small metal pick, knife, or equivalent

d. Wire clipper

e. Pliers

f. One-half-inch brush

g. Rosin-alcohol solder flux

h. Alcohol

11-1916. REPLACEMENT OF DIODES AND TRANSISTORS. Transistors are extremely sensitive to heat, and may be destroyed if subjected to exessive temperatures for even short periods of time. For this reason, an understanding of the soldering techniques used in transistor circuits is very important.

a. Whenever possible, use a low-wattage soldering iron, preferably 25 to 50 watts; in addition, provide a suitable heat sink between the point of contact of the soldering iron and the transistor. This is best done by grasping the transistor lead being soldered with long-nosed pliers, just above the point of soldering iron contact. The pliers will dissipate any excess heat before it is conducted to the transistor. A heat sink is essential and should be employed whenever a soldering iron contacts a transistor lead, no matter how short the period of contact time.

b. Apply the soldering iron to the transistor lead only long enough to melt the solder into a workable state. Never bring the soldering iron into contact with the body of the transistor.

c. If only a high-wattage soldering iron is available, wrap a piece of heavy-gauge wire around the tip of the iron. This extension of wire allows the body of the soldering iron to be held away from the transistor and also aids in dissipating excess heat.

d. The transistor connection points in a given circuit may not be keyed. It is possible, therefore, when replacing the transistor, to insert the substitute backwards, thereby reversing the pin connection to the emitter and collector leads. For this reason, before the transistor is unsoldered from the circuit, identify the emitter and collector terminals in the circuit. This can be done by marking the emitter terminal connection point in the circuit with a pencil, a piece of chalk, or a crayon before removing the transistor.

e. Some soldering irons, when plugged into an ac line, have a voltage existing between the metal of the soldering iron and earth ground. This voltage causes a leakage current that can seriously damage a transistor when the soldering iron is brought into contact with the transistor lead. This effect can be avoided by connecting a jumper lead from the body of the soldering iron to the ground point of the circuit being repaired.

11-1917. REMOVING DEFECTIVE PARTS. To remove a defective component part, proceed as follows:

a. Using a wire clipper, cut (close to lead hole but allowing some of lead to extend through hole) both leads of the component part, and remove the part. Carefully straighten the lead end that extends through each hole so that the lead may be easily withdrawn, as described in step b.

b. Exerting a slight pressure, apply the tip of a hot pencil-type soldering iron to the tip of the lead end (keep soldering iron away from printed-circuit foil). As the lead end absorbs heat, the solder will melt and the lead will break away from its junction with the printed-circuit foil. Remove the soldering iron immediately, and, using pliers, quickly pull the lead free; brush away any excess solder. Do not force or twist the lead to remove it from the printed-circuit card.

c. As a lead end is removed, solder may flow into the open hole in the printed-circuit foil; to remove this solder, tap the card gently while the solder is soft. If this should fail to clean the hole, carefully drill out the solder (using a twist drill of appropriate size). If possible, apply the drill to the printed-circuit side of the card. Drilling from the opposite side may ruin the card by loosening the foil as the drill passes through.

11-1918. REMOVING FLUSH-MOUNTED COMPONENTS. An axial-lead part that has been bonded to a printed-circuit card (with an epoxy resin or similar compound) can be removed by breaking the defective part or by applying heat to the bonding compound. The method to be used depends on the type of part and its location. If a defective axial-lead part cannot be removed by heat, cut or break the part away from the bonding compound. In some cases, the part to be replaced is so closely positioned between other parts that one lead must be cut close to the body of the defective part to permit the application of a prying tool. Wherever possible, cut the defective part with end-cutting pliers or diagonals.

11-1919. Regardless of which tool is employed (round-pointed or spade type), great care must be used in its application to prevent the printed-circuit card or other parts from being damaged or broken. Apply the point of the tool against the bonding compound, between the part and the printed-circuit card. Use the tool in such a manner that it works away the bonding compound from the part to be broken away until enough has been removed for the tool to exert pressure against the part. Keep the leverage surface area of the tool flat against the surface of the printed-circuit card, to prevent the tool from gouging or breaking the board.

CAUTION

Never apply excessive pressure against a printed-circuit card.

11-1920. After the defective part has been removed from the bonding compound, remove the leads or tabs from their terminals on the printed-circuit card. Clean the area thoroughly before installing the new part. Do not remove the compound left on the card (under the removed part) unless required. The mold left in the compound should fit the new part; thus, inserting the new part in this mold helps to prevent vibration of the part.

11-1921. INSTALLING NEW PARTS. To install a new component part in a printed circuit, proceed as follows:

a. Using a knife or a suitable scraper, scrape the leads of the part to be installed.

b. Bend the leads so that they fit snugly into the holes where the component part is to be installed. Mount the component part on the card, gently pushing the leads through the holes. Bend each end of the leads close to the foil.

c. Apply flux to the joint. Touch the lead with the tip of a hot soldering iron, and apply a small amount of fluxless 60/40 solder to the junction.

d. Remove the soldering iron as soon as solder flows into the joint. Hold the part firmly until the solder sets.

e. Using a small amount of alcohol, remove the excess flux.

11-1922. REPAIRING PRINTED-CIRCUIT CONDUCTOR. The printed-circuit conductor is the thin copper foil which is bonded to the card proper and usually covered by a solder coating. Although the printed-circuit conductor can withstand normal handling and will operate with no trouble under rated service conditions, it is likely to damage if handled carelessly. Should any part of the conductor be damaged (i.e., split or raised from the card proper), perform the required repairs as follows:

a. When part of the conductor has been raised from the card, remove it by clipping both conductor ends close to the card; a split in the conductor requires no such preliminary preparation.

b. Bend a piece of a 20-gauge copper wire into the shape of a staple. The staple should be long enough to span the defective portion of the conductor and to allow clinching approximately 1/4 inch from each end, once placed in position on the printed-circuit card.

c. From the printed-circuit side of the card, drill two holes (of appropriate size) to receive the ends of the wire staple. Unless the printed-circuit conductor is approximately 1/4 inch wide, do not drill the holes directly into the foil, but drill them near enough that the staple is parallel to the foil. If the foil is 1/4 inch wide, drill the two holes into the printed-circuit conductor. If the repair is not in a congested area of printed circuitry, proceed to step d. Otherwise, insert the staple from the component side of the card and, on the opposite side, clinch each end of the staple diagonally across the printed-circuit conductor.

d. If space permits, drill two additional holes in the printed-circuit card, each close to one of the previously drilled holes, but on the opposite side from the printed-circuit conductor. From the printed-circuit side of the card, insert the ends of a wire staple into the two holes on one side of the conductor; then, holding the staple flush against the card, bend each end of the staple back through the adjacent hole closest to it. Pull the wire taut and clinch each end across the printed-circuit conductor; then proceed to step e.

e. Solder the two joints by the method previously described.

11-1923. REPAIRING CRACKS IN PRINTED-CIRCUIT CARDS. A broken printed-circuit card may have to be repaired in an emergency (temporary status) where no replacement is available. Before repairing a broken printed-circuit card, assess the damage for the extent of the break and the amount of damage to the parts involved. If the card is not too complicated or the damage too extensive, the card can probably be repaired.

11-1924. If a small portion or nonsupporting corner of the board is broken off, it may be rebonded to the larger section with a nonconductive cement. If cementing is not feasible or does not hold satisfactorily, the pieces can be fastened together with wire staples, cut from solid conducting wire of the diameter and length required, depending on the width of the conducting strip to be repaired.

11-1925. To insert the staples, drill holes about 1/4 inch from each side of the break. The holes should be just large enough to accommodate the wire used for stapling. (The size of the wire required may vary, depending on the width of the conductive strip to be repaired.) Drill the holes through the conducting strips so that the staples will provide a good electrical contact across the break; this method will permit the use of enough staples to hold the pieces together, without danger of shorts between conductors. If the break is very large, position additional staples at all points possible to give the board more support.

11-1926. Where the adhesive and stapling method described above does not provide sufficient structural strength or rigidity, consideration should be given to the use of splints or a doubler. Strips of thin card material are glued across the fracture by means of nonconductive adhesive material. Where additional strength is required, a small section of the card material may be glued to the splints by means of the nonconductive adhesive.

11-1927. Rebond any loose conducting strips with a nonconductive bonding cement; then apply nonconductive cement to both sides of the break, and join the sections together. Insert half of the measured and pre-cut wire staples from top to bottom, and the other half from bottom to top, bending the ends flush against the card. Solder these staples to the conducting strip.

11-1928. If the card is not completely broken, but is only cracked, drill a hole at the end of each crack to prevent further lengthening of the break. Then repair the crack in the same manner as for the complete break discussed above.

11-1929. After the repairs are completed, clean both sides of the repaired area with a stiff brush and solvent such as methyl chloroform (GM 6810-664-0387). Allow the board to dry thoroughly, and then coat the repaired area with an epoxy resin or similar compound. This coating will not only protect the repaired area, but will also help to strengthen it.

NOTE

When a board is broken, it is much better to replace the entire board. The repair techniques described here are for emergency repair only.

11-1930. TESTING AFTER REPAIR. As far as possible, performance tests of the plug-in card subassemblies are performed by using the operating equipment as the test unit. If operating spare components are available, repaired plug-in cards can be brought closer to operating standards by this method. The plug-in cards can also be tested by using computer test equipment which simulates the operating conditions of the card.

SECTION XII

TELEPHONE AND TELEGRAPH EQUIPMENT TESTING

11-1931. GENERAL.

11-1932. A telephone or telegraph arrangement must be maintained with a minimum of service interruptions. Periodic testing will provide information which will permit you to determine the condition of the telephone or telegraph equipment and perform maintenance to keep interruptions to a minimum.

11-1933. Some knowledge of the theory of the test methods used for telephone and telegraph equipment is essential for correct application of each testing method. This section discusses the various methods used in telephone and telegraph equipment testing.

11-1934. TELEPHONE TESTING.

11-1935. HANDSET TESTING.

11-1936. If the transmitter of the handset is suspected of being faulty, the following test procedure should be used: Unscrew the transmitter mouthpiece and remove the transmitter element. Connect an ohmmeter across the contact electrodes of the transmitter element, and speak into the transmitter element. If the transmitter is working properly, the ohmmeter will indicate fluctuating resistance. If the contacts are dirty, or the transmitter is open-circuited, the ohmmeter will indicate a very high or infinite resistance, respectively.

11-1937. If the receiver of the handset is suspected of being faulty, the following procedures should be used: Disconnect the output of the receiver transformer, and connect an ohmmeter between the disconnected lead and the output of the receiver circuit. If the circuit to the handset is complete and the receiver is not open, the ohmmeter will read a certain value of resistance, depending on the type of receiver you are testing. If this reading is not obtained, remove the receiving element from the handset receiver and connect an ohmmeter across the contact electrodes. If the ohmmeter indicates a greater resistance, check the contacts for dirt and for proper tension. If a smaller resistance is indicated, the contacts may be partially shorted through a dust/grease path. In any case, clean the contacts. A typical handset is shown in figure 11-283.

11-1938. NOISE MEASURING.

11-1939. A noise-measuring set, as shown in figure 11-284, is a device which provides rapid and accurate determination of noise values associated with telephone equipment. The noise-measuring set indicates the noise magnitude by the combination of a dial setting and the deflection of a meter needle.

11-1940. The noise-measuring set may be used to measure metallic-circuit noise, noise in a telephone receiver, noise to ground, noise and cross-talk volume on a bridging basis, and volume on message or program telephone circuits.

11-1941. Different kinds of simple frequency noise voltages have different inter-

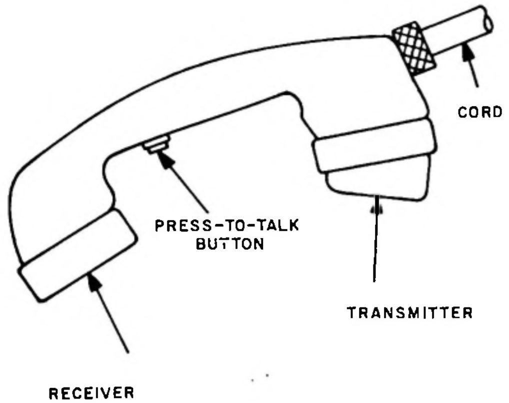


Figure 11-283. Typical Handset

fering effects for various applications. To obtain the relative interfering effect of values for different applications, equalizer networks are included in the noise-measuring test set; they are used in combination with other characteristics of the set to provide frequency weighting characteristics so that, for a given application, two noises that are equal with a given frequency weighting are approximately equal in interfering effect.

11-1942. LINE TESTING.

11-1943. The procedures used for locating the cable to be tested, determining talking and calling circuits, and performing the tests for various situations are discussed herein. This information applies to telegraph circuits as well as to telephone circuits.

11-1944. BATTERY-RECEIVER METHOD. The battery-receiver method is used to detect and locate shorts, crosses, opens, and grounds. In each case, the method of detection and location depends on your interpretation of a click heard in the receiver.

11-1945. DETECTION. The battery-receiver method may be used to detect individual faulty conductors of a cable. To test for a short or cross, disconnect the circuit

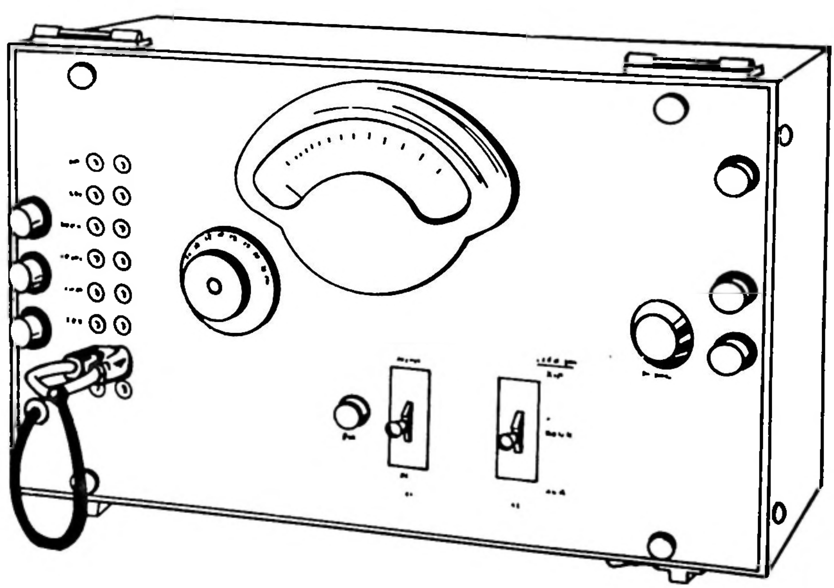


Figure 11-284. Typical Noise-Measuring Set.

at the termination point as shown in parts A and B of figure 11-285. Connect the battery and receiver to the ends of the conductors being tested. A click which is proportional to the resistance of the fault will be heard in the receiver.

11-1946. To test for an open, first strap and ground the conductors which are to be tested at the termination point as shown in part C of figure 11-285. Connect the receiver and battery in series to ground. Connect the receiver test lead at point B. A capacitance click will be heard in the receiver if the conductor is open. Disconnect the receiver test lead from point B, and reconnect it to point A. In this case, a fault click only will be heard.

11-1947. To test for a ground fault, connect the receiver and battery in series to ground as shown in part D of the figure, and connect the receiver test lead to point A. A capacitance click should be heard. When the receiver test lead is connected to point B, a fault click will be heard on both make and break circuits, indicating a faulty conductor.

11-1948. LOCATION. When the faulty conductor or conductors have been detected, the fault should be located systematically with the battery and receiver. This can be accomplished by testing between two points. One of these points must be the central office. The second point must be taken successively at various distances from the central office. If the fault is not between the initial second point and the central office, the tester picks another second point at a greater distance from the central office. The procedure is repeated until the general area of the fault has been determined. Then, the exact location of the fault can be determined by one of the location methods discussed in this section.

11-1949. RESISTANCE-MEASUREMENT METHODS. The Wheatstone bridge and its

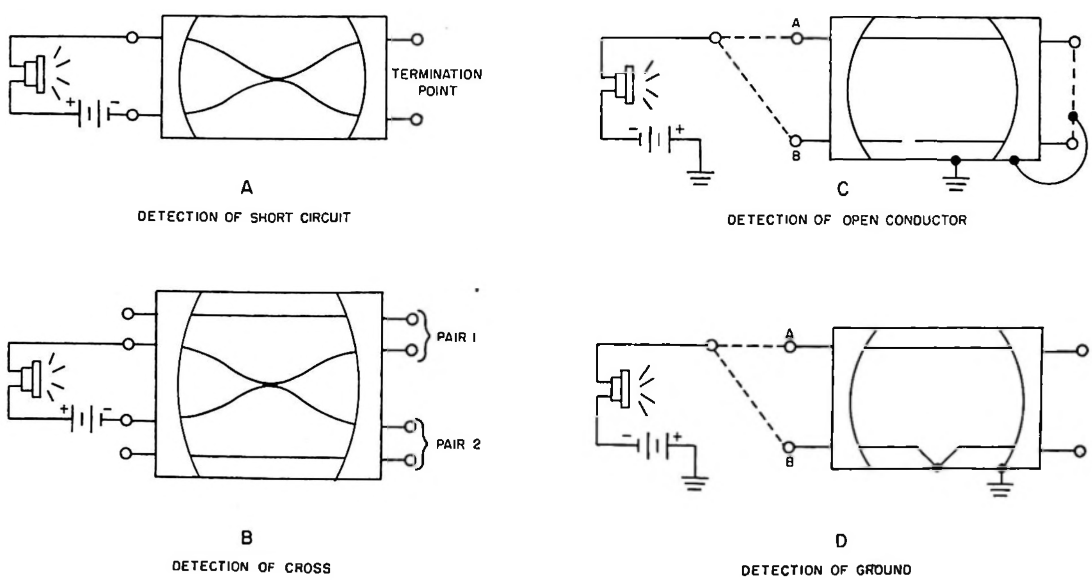


Figure 11-285. Battery-Receiver Methods of Detecting Faulty Conductors

adaptations may be used to locate faults. The particular test method to be used depends upon such factors as the length and composition of the cable to be tested and the nature of the fault to be located. The adaptations of the Wheatstone bridge method are the simple loop, the Varley loop, and the Murray loop.

11-1950. The simple loop test is used to determine the total loop resistance of a circuit. In figure 11-286 the ends of the conductors are strapped. Line 1 and line 2 will therefore be connected in series. Since this is a series circuit, line 1 and line 2 are equivalent to a single resistance, R_X, as shown in figure 11-287. The total loop resistance may then be calculated by applying the equation:

$$R_X = \frac{AR}{B}$$

where A, R, and B identify the resistance arm values in figure 11-287. This equation is also the basic equation used for both the Varley loop and Murray loop test methods. It is used throughout the procedures discussed in the following paragraphs.

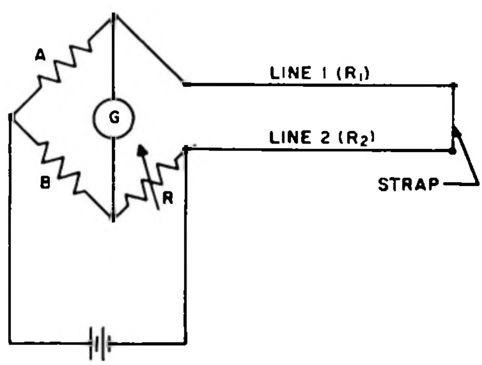


Figure 11-286. Simple Loop Test for Determining the Total Loop Resistance of a Circuit

11-1951. The Varley loop test is accomplished by connecting the distant end of the faulty conductor to a good conductor. Figure 11-288 illustrates the circuit connec-

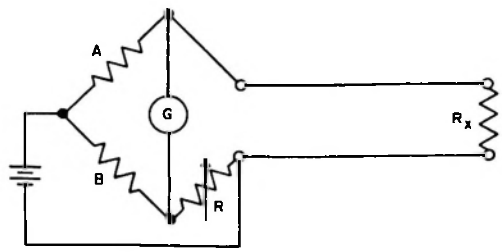


Figure 11-287. Equivalent Total Loop Resistance

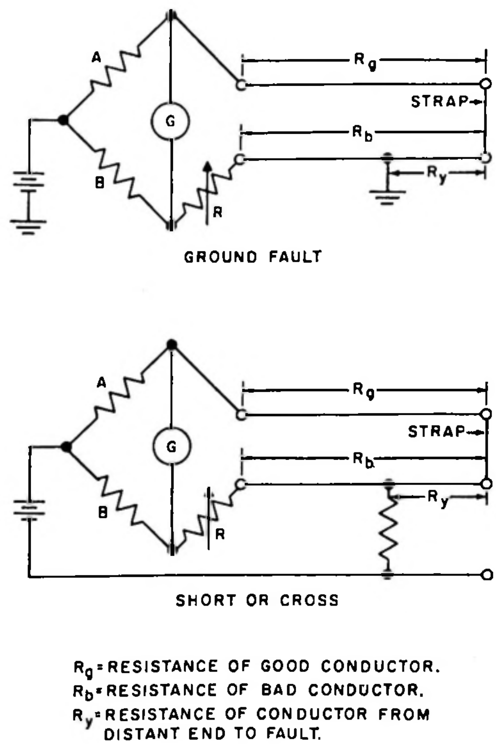
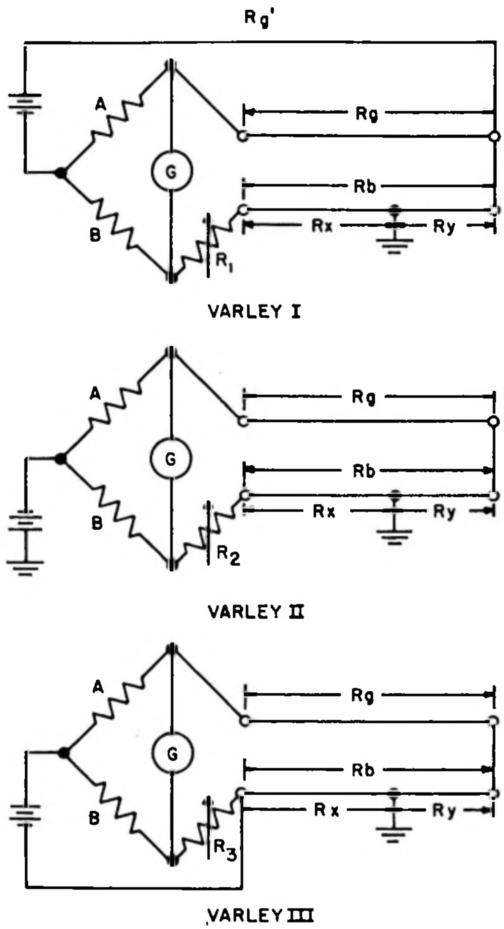


Figure 11-288. Regular Varley Loop



Rg' = Rg = RESISTANCE OF GOOD CONDUCTOR.
Rb = RESISTANCE OF BAD CONDUCTOR.
Ry = RESISTANCE OF CONDUCTOR FROM DISTANT END TO FAULT.
Rx = RESISTANCE OF CONDUCTOR FROM TESTING END TO FAULT.

Figure 11-290. Three-Varley Loop Method for Determining the Distance to a Ground Fault

$$\frac{A}{B} = \frac{R_g + R_y}{R_2 + R_b - R_y} \qquad \text{(Varley II)} \qquad (2)$$

$$\frac{A}{B} = \frac{R_g + R_b}{R_3} \qquad \text{(Varley III)} \qquad (3)$$

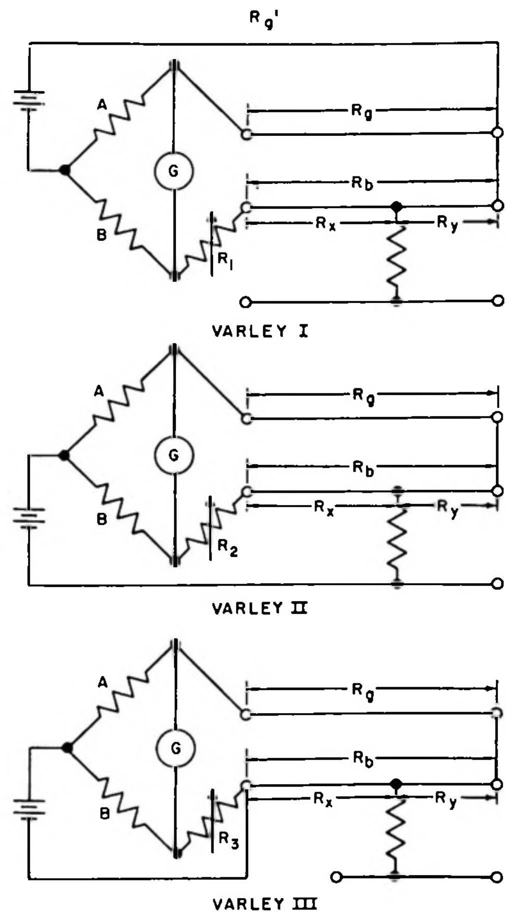


Figure 11-291. Three-Varley Loop Method for Determining the Distance to a Short or Cross Fault

When the bridge is balanced, reading R_2 will differ from reading R_1 by the amount of resistance of the bad conductor from the strap to the fault, or R_y. Therefore, determine R_y as follows:

Find the common denominator of Varley I (1) and Varley II (2):

$$AR_1 + AR_b = BR_g \qquad (4)$$

tions for a ground, short, or cross fault. The strap, which connects the faulty conductor to the good conductor, has a negligible resistance. The circuit connections of the figure form a looped circuit which the fault divides into two parts. For this test, the resistances of the good and bad conductors must be known, either from a previous test or from tests made before fault location is attempted. The simple loop method is one method of finding the resistances of the conductors. When the respective resistances of the good and bad conductors of a similar pair of conductors are equal and resistances A and B are equal, a simple expression for R_y may be obtained:

$$R_y = \frac{R}{2}$$

where R is the set value of the variable resistance shown in the figure. The value of the R arm equals the loop resistance from the fault to the distant end of the conductors.

11-1952. The check test shown in figure 11-289 may be made by reversing the good and bad conductor connections. The figure illustrates the check test for a ground fault. However, a check test for a short or cross may also be performed in a similar manner.

11-1953. The three-Varley loop method is more advantageous than the regular Varley loop and the Murray loop method, because in the three-Varley loop method the resistance of the good and bad conductors need not be known. In addition, the three-Varley loop method eliminates the effects of lead resistance, which makes the method especially useful for tests from a central office desk. This method uses two good conductors of any resistance and a ground connection. The three-Varley loop test method is also used for locating faults in a coaxial cable. When using the Wheatstone bridge for any type of Varley loop, you must read the applicable operation handbook of the Wheatstone bridge test set and make the proper connections for the tests.

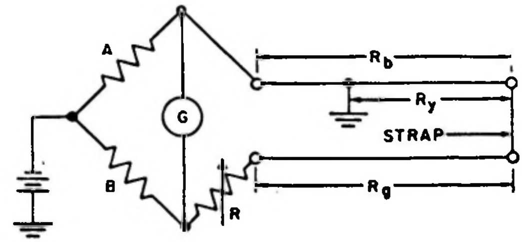


Figure 11-289. Regular Varley Loop Check Test Circuit

11-1954. Figures 11-290 and 11-291 illustrate the circuit connections for determining the distance to a fault by the three-Varley loop method. Figure 11-290 illustrates the circuit connections for determining a ground fault, and figure 11-291 illustrates the circuit connections for determining a short or cross fault. The conductors in figures 11-290 and 11-291 are strapped at their distant end. This strap should be tight and offer a negligible resistance. The three measurements which are taken for this method are called Varley I, Varley II, and Varley III.

11-1955. After connecting the circuits as illustrated in figures 11-290 and 11-291, the three-Varley measurements are taken. These Varley measurements determine the values of R_1, R_2, and R_3. The following equations will provide you with the value of A/B for each of the three-Varley measurements. The ratio of A/B must be the same for all three measurements.

$$\frac{A}{B} = \frac{R_g}{R_1 + R_b} \qquad \text{(Varley I)} \qquad (1)$$

$$AR_2 + AR_b - AR_y = BR_g + BR_y \qquad (5)$$

Subtracting (4) from (5):

$$\begin{array}{l} AR_2 + AR_b - AR_y = BR_g + BR_y \\ -AR_1 - AR_b \qquad\quad = -BR_g \\ \hline -AR_1 + AR_2 - AR_y = BR_y \qquad (6) \end{array}$$

Add $+AR_y$ to both sides of the equation:

$$+AR_2 - AR_1 = AR_y + BR_y \qquad (7)$$

Factor:

$$A(R_2 - R_1) = R_y(A + B) \qquad (8)$$

Therefore:

$$R_y = \frac{A}{A + B}(R_2 - R_1) \qquad (9)$$

Similarly, the following equations may be derived:

$$R_x = \frac{A}{A + B}(R_3 - R_2)$$

$$R_b = R_y + R_x = \frac{A}{A + B}(R_3 - R_1)$$

11-1956. The most convenient arm ratios (A/B) are 1/9 and 1/4. When these arm ratios are used, the equations for R_y, R_x, and R_b become:

Ratio of 1/9:

$$R_y = \frac{(R_2 - R_1)}{10}$$

$$R_x = \frac{(R_3 - R_2)}{10}$$

$$R_b = \frac{(R_3 - R_1)}{10}$$

or:

Ratio of 1/4:

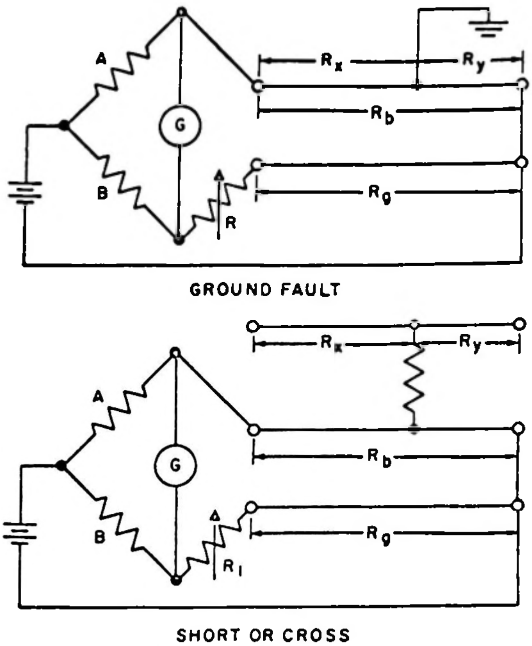
$$R_y = \frac{(R_2 - R_1)}{5}$$

$$R_x = \frac{(R_3 - R_2)}{5}$$

$$R_b = \frac{(R_3 - R_1)}{5}$$

11-1957. When the total resistance of the loop is more than 1100 ohms, it may not be possible to balance the bridge with arm ratios of 1/9 and 1/4. If this is the case, it may be necessary to use an arm ratio of 1. However, if the good conductor has a lower resistance than the bad conductor, and the arm ratio is 1, the bridge will not balance for the Varley I measurement. To make the bridge balance for the Varley I measurement, the good and bad conductor leads must be interchanged at the bridge terminals, as shown in figure 11-292.

11-1958. The modified Varley loop is used when the bridge will not balance because an arm ratio of 1 is being used and the faulty conductor has a larger resistance than the good conductor. To avoid interchanging the good and bad conductor leads for the Varley I measurement, the modified Varley loop can be used, as shown in figures 11-293 and 11-294. This circuit differs from that of the regular Varley loop in that a small resistance is connected between the test set and the good conductor. This resistance is used during all three Varley measurements. The value of the resistance must be large enough to make the resistance of the good conductor higher than that of the bad conductor. The value of the resistance need not be known, since it is not included in the derivation of the equations. The equations which apply to the modified Varley loop test when the arm ratio is 1 are as follows:



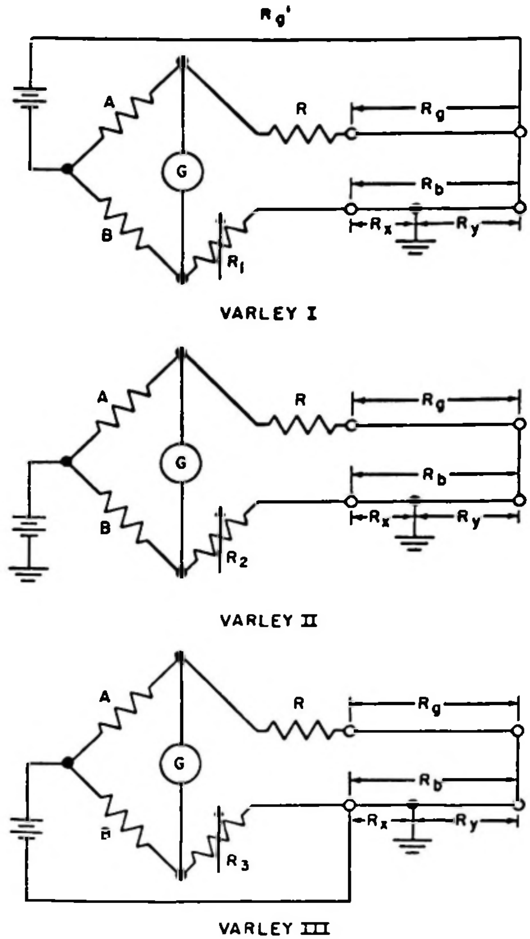
R_g = RESISTANCE OF GOOD CONDUCTOR.
R_b = RESISTANCE OF BAD CONDUCTOR.
R_y = RESISTANCE OF CONDUCTOR FROM DISTANT END TO FAULT.
R_x = RESISTANCE OF CONDUCTOR FROM TESTING END TO FAULT.

Figure 11-292. Three-Varley-Loop Method, Reversal of Conductor Leads

$$R_y = \frac{R_2 + R_1}{2}$$

$$R_x = \frac{R_3 + R_2}{2}$$

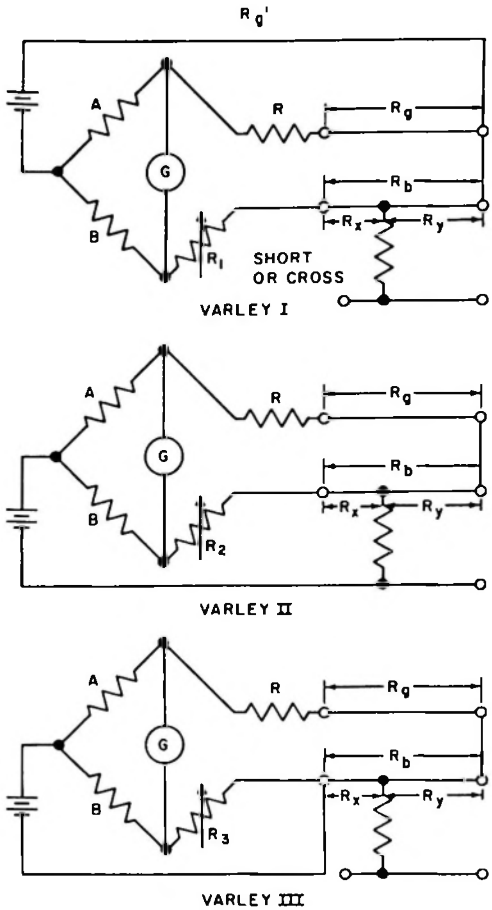
$$R_b = \frac{R_3 + R_1}{2}$$



$R_g' = R_g$ = RESISTANCE OF GOOD CONDUCTOR.
R_b = RESISTANCE OF BAD CONDUCTOR.
R_y = RESISTANCE OF CONDUCTOR FROM DISTANT END TO FAULT.
R_x = RESISTANCE OF CONDUCTOR FROM TESTING END TO FAULT.

Figure 11-293. Modified Varley Loop for Locating a Ground Fault

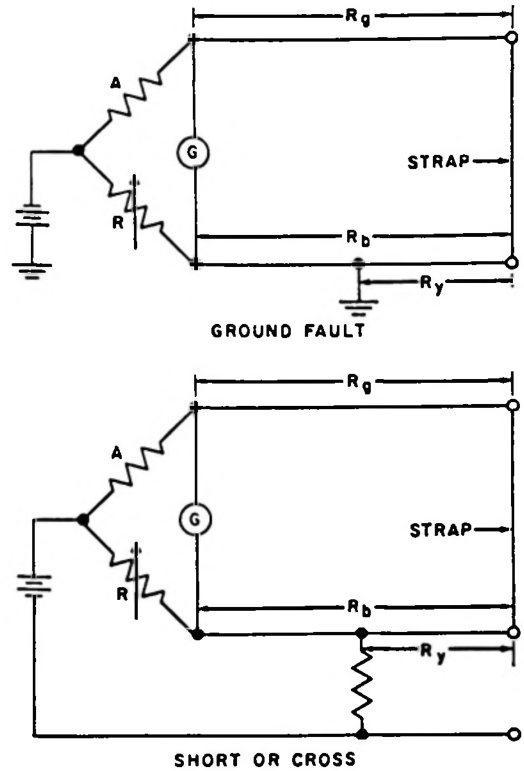
11-1959. The Murray loop test method is used for locating faults in short cable lengths involving relatively low resistances. When using this method, the loop resistances of the two conductors (faulty and good) must be known. The theory of the Murray loop is similar to that of the Varley loop. However, instead of setting arms A and B of the bridge to equal values and using the variable resistor R to compensate for the conductor resistance between the good and faulty conductor, arm B is completely eliminated and



$R_g' = R_g$ = RESISTANCE OF GOOD CONDUCTOR.
R_b = RESISTANCE OF BAD CONDUCTOR.
R_y = RESISTANCE OF CONDUCTOR FROM DISTANT END TO FAULT.
R_x = RESISTANCE OF CONDUCTOR FROM TESTING END TO FAULT.

Figure 11-294. Modified Varley Loop for Locating a Short or Cross Fault

variable resistor arm R is connected as illustrated in figure 11-295. This figure illustrates the circuit connections for locating a ground fault and the circuit connections for locating a short or cross fault. The conductors are strapped at the distant ends.



R_g = RESISTANCE OF GOOD CONDUCTOR.
R_b = RESISTANCE OF BAD CONDUCTOR.
R_y = RESISTANCE OF CONDUCTOR FROM DISTANT END TO FAULT.

Figure 11-295. Murray Loop Circuit for Locating Faulty Conductors

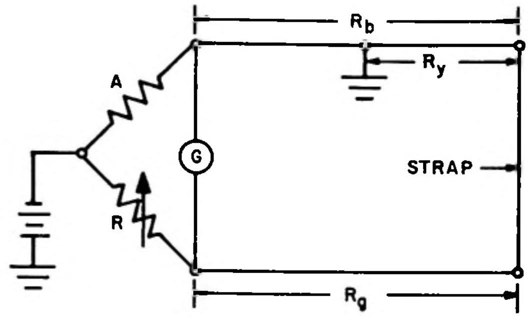
This strap must be tight and have negligible resistance. In this arrangement, the ratio of the reading R to the setting of arm A is equal to the ratio of the resistance of the defective conductor from the testing point to the fault, divided by the resistance of this same conductor from the fault to the distant end plus the resistance of the good conductor. This equation may be written as follows:

$$\frac{R}{A} = \frac{R_g - R_y}{R_g + R_y}$$

In this equation, it is assumed that the faulty conductor and good conductor have the same series resistance per 1000 feet, and are of the same length. This is, in all probability, the situation. However, if this is not true, the numerator and denominator of the right side of the equation will have to be multiplied by their respective resistances (taken from table 11-16). Therefore, the equation becomes:

$$\frac{R}{A} = \frac{(R_g - R_y)\, r_1}{(R_g + R_y)\, r_2}$$

where r_1 and r_2 equal the resistances per 100 feet multiplied by the applicable length of the faulty and good conductors, respectively.



R_g=RESISTANCE OF GOOD CONDUCTOR.
R_b=RESISTANCE OF BAD CONDUCTOR.
R_y=RESISTANCE OF CONDUCTOR FROM DISTANT END TO FAULT.

Figure 11-296. Murray Loop Check Test for a Ground Fault

11-1960. A check test may be made by reversing the good and faulty conductor connections. Figure 11-296 illustrates the check test for a ground fault. However, a check test for a short or cross fault may be performed in a similar manner.

11-1961. VOLTMETER METHOD. The voltmeter method is used to detect shorts, crosses, grounds, and opens. To test for a short, first disconnect the conductors of the cable at a cable terminal or at some convenient point in the circuit. At the test point, connect the voltmeter as illustrated in figure 11-297. The source of voltage for the voltmeter may be applied by an external battery at the test point or by the central office. If a reading is observed on the voltmeter, the conductors are shorted. This same test may be used for the detection of a cross.

11-1962. To test for a ground, first strap the conductors at a cable terminal as shown

Table 11-16. Resistance of Conductors

GAUGE (AWG)	DIAMETER (IN.)	RESISTANCE	
		Ohms/1000 feet at 20°C (68°F)	Feet/ohm at 20°C (68°F)
13	0.072	2.026	493
16	0.0508	3.977	251
19	0.0359	8.049	124
22	0.0253	16.190	62
24	0.0201	25.94	39
26	0.0159	41.67	24

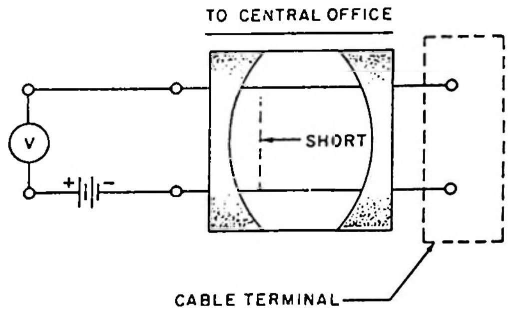


Figure 11-297. Voltmeter Method of Detecting a Short or Cross Fault

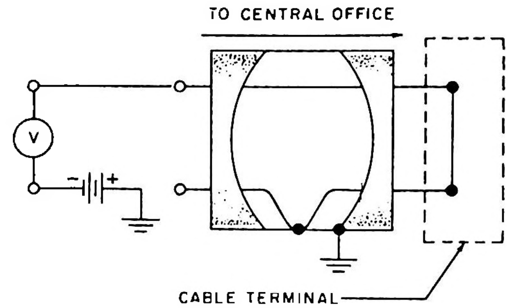


Figure 11-298. Voltmeter Method of Detecting a Ground Fault

in figure 11-298 or at any other convenient point in the circuit. Then place the voltmeter in the circuit by connecting one voltmeter lead to a conductor and the other lead (battery lead) to ground. If a deflection is observed on the meter, the conductor is grounded.

11-1963. To test for an open, first strap the conductors at a cable terminal as shown in figure 11-299 or at any other convenient point in the circuit. Then place the voltmeter in series with the circuit. If no reading is observed on the meter, there is an open in the circuit.

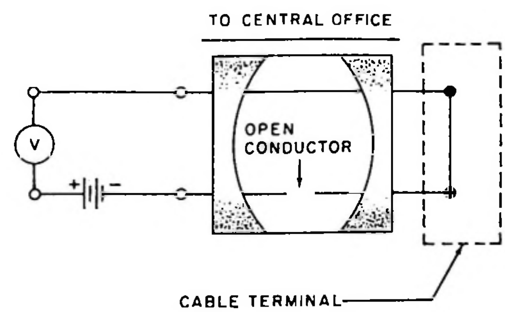


Figure 11-299. Voltmeter Method of Detecting an Open Circuit

11-1964. PRACTICAL LOCATION APPLICATIONS. The basic theory underlying the determination of an unknown distance by line resistance measurements can be applied in a practical manner to a cable circuit. The resistance of a cable circuit, as illustrated in part A of figure 11-300, is the resistance of line 1 plus that of line 2 between points BC and EF, respectively. The resistances between points AB and DE may be either cable or open-wire conductors. The line points B, E, C, and F may be binding posts of a cable terminal or any selected set of test points, such as T_1 and T_2. In this example, T_1 will be selected as the operating test point; it may be located any distance from the central office equipment.

11-1965. Select the proper voltmeter scale for the test. Connect the shunt lead of the voltmeter across points B' and E, and then depress the shunt switch, S. The reading ($V_{B'E}$) observed on the voltmeter is equal to $V_O - V_{ABED}$ (where V_O is the original voltage). The voltage V_{BE} may be considered the source voltage, which is equivalent to the total circuit voltage, V_B. If the shunt potential $V_{B'E}$ is replaced by an

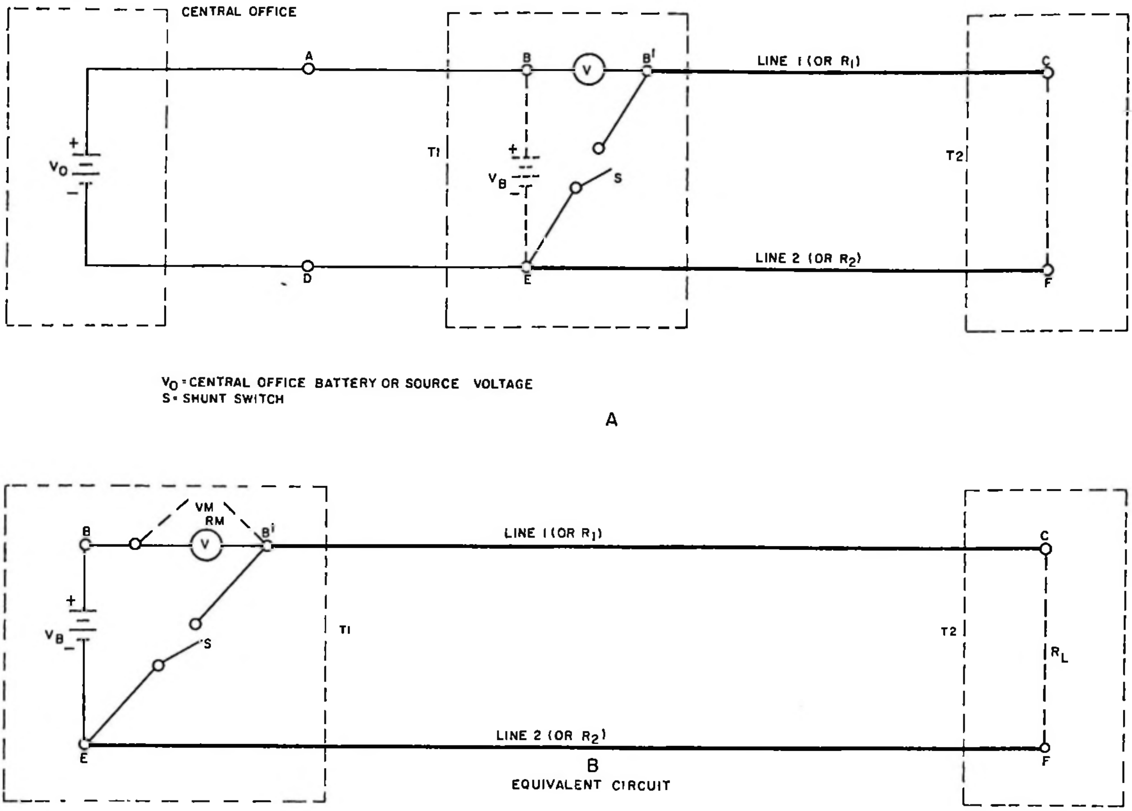


Figure 11-300. Determination of Distance by Line-Resistance Measurements

equivalent battery, as shown in part B of the figure, that battery voltage may be considered as V_B.

11-1966. Strap the circuit at points C and F of test point T_2. Since this connects lines 1 and 2 in series, lines 1 and 2 may be considered as a single resistance equal to R_L, as shown in part B of the figure. Observe the reading on the meter with the shunt switch, S, in the open position. This reading is the voltage drop across the meter, which is equivalent to voltage V_M. At this point the unknown resistance, R_L, may be calculated from the following equation:

$$R_L = \frac{R_M (V_B - V_M)}{V_M}$$

where R_M is the voltmeter resistance and V_M is the voltage across the meter. The only factor on the right side of the equation which is not known is the resistance of the voltmeter. This may be calculated by multiplying the sensitivity of the meter by the meter scale you use for this test.

11-1967. After the unknown resistance, R_L, has been calculated, the distance from point B to point C may be calculated by use of table 11-16. The tabulated data includes the diameter of each conductor in the cable, the resistance in ohms/1000 feet, and the reciprocal in feet/ohm. Therefore, select the proper resistance for the cable type of lines 1 and 2. Multiply this resistance by R_L to obtain the distance BCFE. To obtain distance BC, divide the total distance BCFE by 2.

11-1968. In many cases, the gauge of a length of conductor will be nonuniform. That is, part of the conductor may be 22 gauge, part 19 gauge, and part 24 gauge. To locate the cable fault accurately, it is necessary to convert a nonuniform cable to an equivalent length of single-gauge cable. For instance, 100 feet of 22 gauge is equal in resistance to 200 feet of 19 gauge or 62 feet of 24 gauge. Table 11-17 lists the conversion factors that are necessary to convert one gauge-size length of cable to its equivalent length in another gauge size.

11-1969. DETERMINATION OF DISTANCE TO A FAULT. When a fault such as a short or cross occurs in the circuit, the procedure previously discussed must be modified to determine the exact location of the fault. The procedure must be modified because otherwise the fault resistance would be included in the calculation of R_L, and, therefore, would result in an erroneous calculation of distance. The modification is twofold; a third conductor of the cable must be used in order to make two loop resistance measurements to determine the resistance of the fault, and a third loop resistance measurement must be made (as shown in figure 11-301) to accurately determine the distance to the fault.

11-1970. Strap the circuit at points C and H as shown in part A of figure 11-301. Connect voltmeter leads B and E to points B' and G, and determine loop resistance BCHE. Connect voltmeter lead B to point E' as shown in part B of the figure. Now determine the resistance of loop BE'KJCHE. The fault resistance (R_F) may be calculated by subtracting the loop resistance BCHE from loop resistance BE'KJCHE.

11-1971. Connect voltmeter leads B and E to points B' and E', respectively, as shown in part C of the figure, and determine the loop resistance BJKE. To determine the distance to the fault, proceed as follows: Subtract the fault resistance, R_F, from loop BJKE, and divide the difference by 2. The result of this subtraction and division is the resistance of line E'K. To determine the distance, d, which is the distance to the fault, convert the line resistance to the equivalent distance, using table 11-16.

Table 11-17. Gauge Conversion Data Table

ACTUAL GAUGE	FACTOR TO DETERMINE EQUIVALENT LENGTH FOR:					
	13 ga	16 ga	19 ga	22 ga	24 ga	26 ga
13	1.0	0.510	0.252	0.125	0.078	0.049
16	1.96	1.0	0.496	0.246	0.153	0.095
19	3.97	2.02	1.0	0.496	0.31	0.19
22	7.99	4.07	2.01	1.0	0.624	0.389
24	12.8	6.52	3.22	1.6	1.0	0.623
26	20.6	10.5	5.18	2.57	1.61	1.0

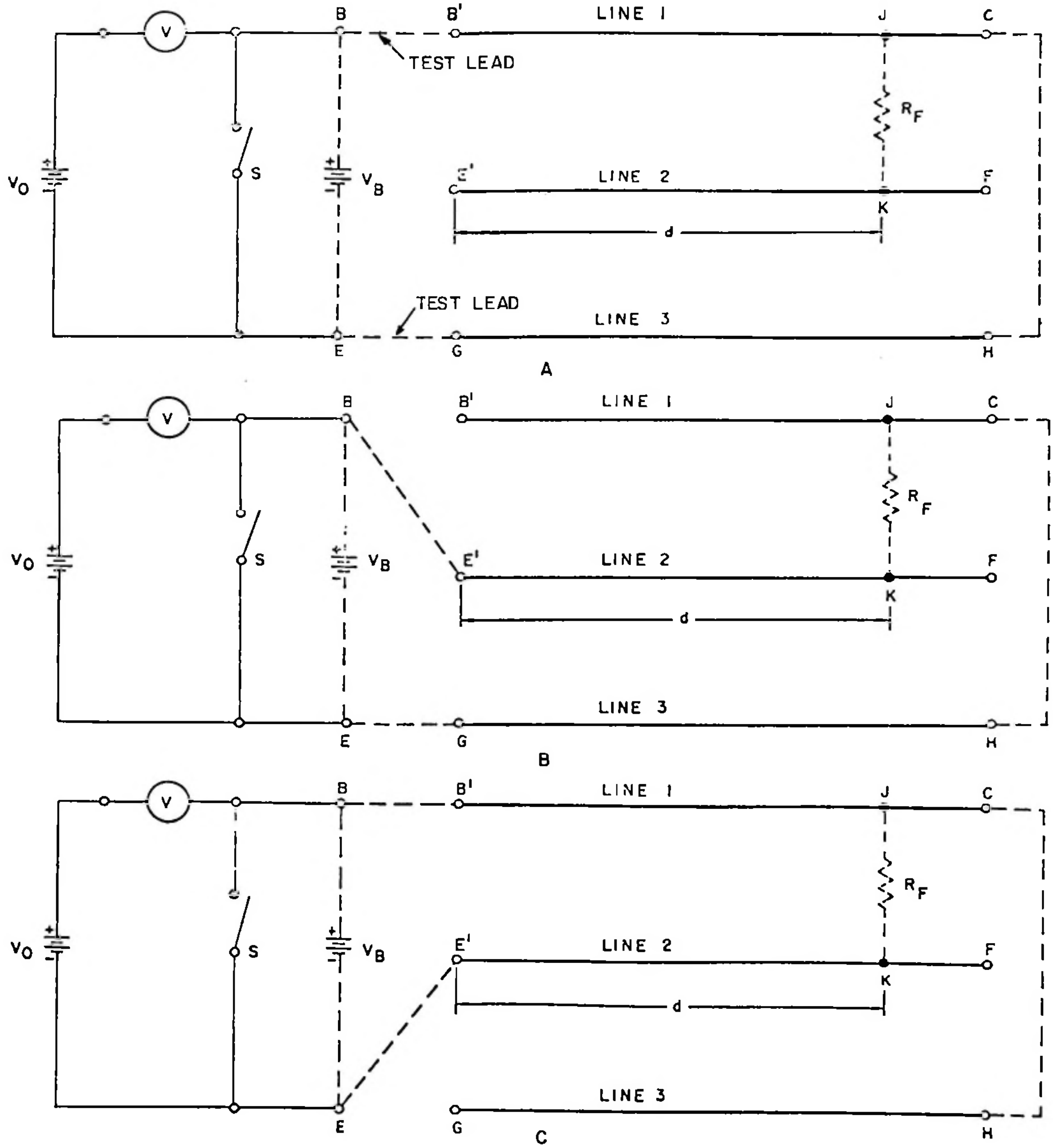


Figure 11-301. Determination of Distance to a Short or Cross Fault

11-1972. When a ground fault occurs in a circuit, the resistance of the fault need not be calculated. The resistance of a fault is written in the equation as R_{JG}, and will be cancelled when the loop resistances are calculated. To calculate the distance to the fault, proceed as follows (refer to figure 11-302 for the mathematical analysis behind each step):

a. Determine the source voltage, V_B. Then strap the circuit at points C and H.

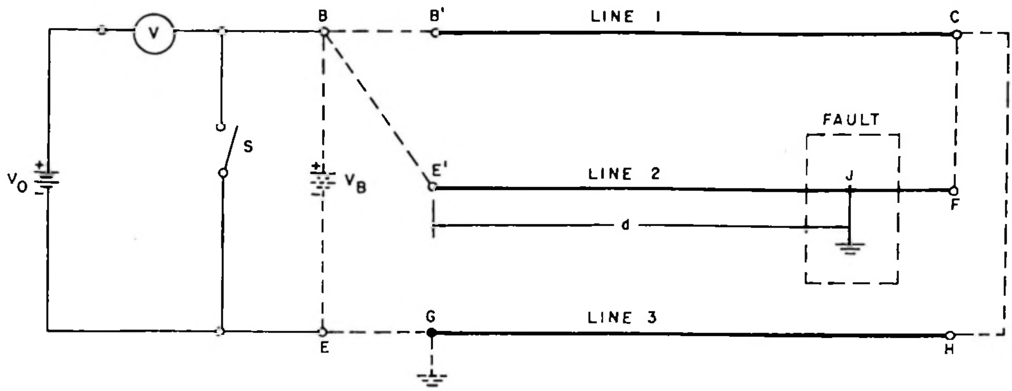


Figure 11-302. Determination of Distance to a Ground Fault

Connect voltmeter leads B and E to points B' and G, respectively. Determine the loop resistance B'CHG. Now remove the short circuit at points C and H, and strap points C and F. Connect point G to ground. Connect voltmeter leads B and E to points B' and G, respectively. Determine loop resistance B'CFJG.

b. Subtract B'CFJG from B'CHG, and identify the result as L3. After this is done, connect voltmeter leads B and E to points E' and G, respectively. Be sure that point G is still grounded. Now determine loop resistance E'JG, and identify this value as L_4.

c. Add L_4 to L_3 and divide by 2. The result will be the resistance of the distance to the fault.

11-1973. OHMMETER METHOD. The ohmmeter method for detecting and locating faults in a cable circuit is very similar to the voltmeter method. The two methods differ only in the procedures for determining the loop resistances. In the voltmeter method the loop resistances were calculated, while in the ohmmeter method the loop resistances are read directly on the meter scale. The detection of a fault for both methods, however, is indicated by a deflection on the meter scale. When locating faults, similar loop resistances must be measured.

11-1974. TONE COMPARISON OR BALANCE TEST METHOD. The tone comparison or balance test method is used for detecting shorts, crosses, grounds, and opens. This test is performed with a test set in conjunction with a standard receiver and resistances, or a test set in conjunction with a split-head receiver. In addition, a tone set must be supplied with each set of equipment. However, for all tests, the method of testing is identical. All conductors to be tested must be of the same gauge and length. If this is not the case, a balance cannot be obtained. This is true, even if all the conductors in the group are good. All faults detected by the tone comparison test method must be checked by the battery-receiver test method. A slight tone will be heard in the receiver for all conductors, good or bad, if the fingers of a tester touch either the test lead or the exploring lead.

11-1975. The detection of faults by the tone comparison test method, as shown in figure 11-303, is performed by first selecting a standard pair. For the example illustrated in figure 11-303, pair 1 is the standard pair. The standard pair must be of the same length and gauge as the conductor pairs to be tested. This standard pair must be tested and found free of shorts, crosses, grounds, or opens, and, in addition, must produce no tone or only a very slight tone in the receiver when the tip side is balanced against the ring side of a circuit. If the conductors under test are in a different cable group, and therefore of different length, this fact will be indicated by some

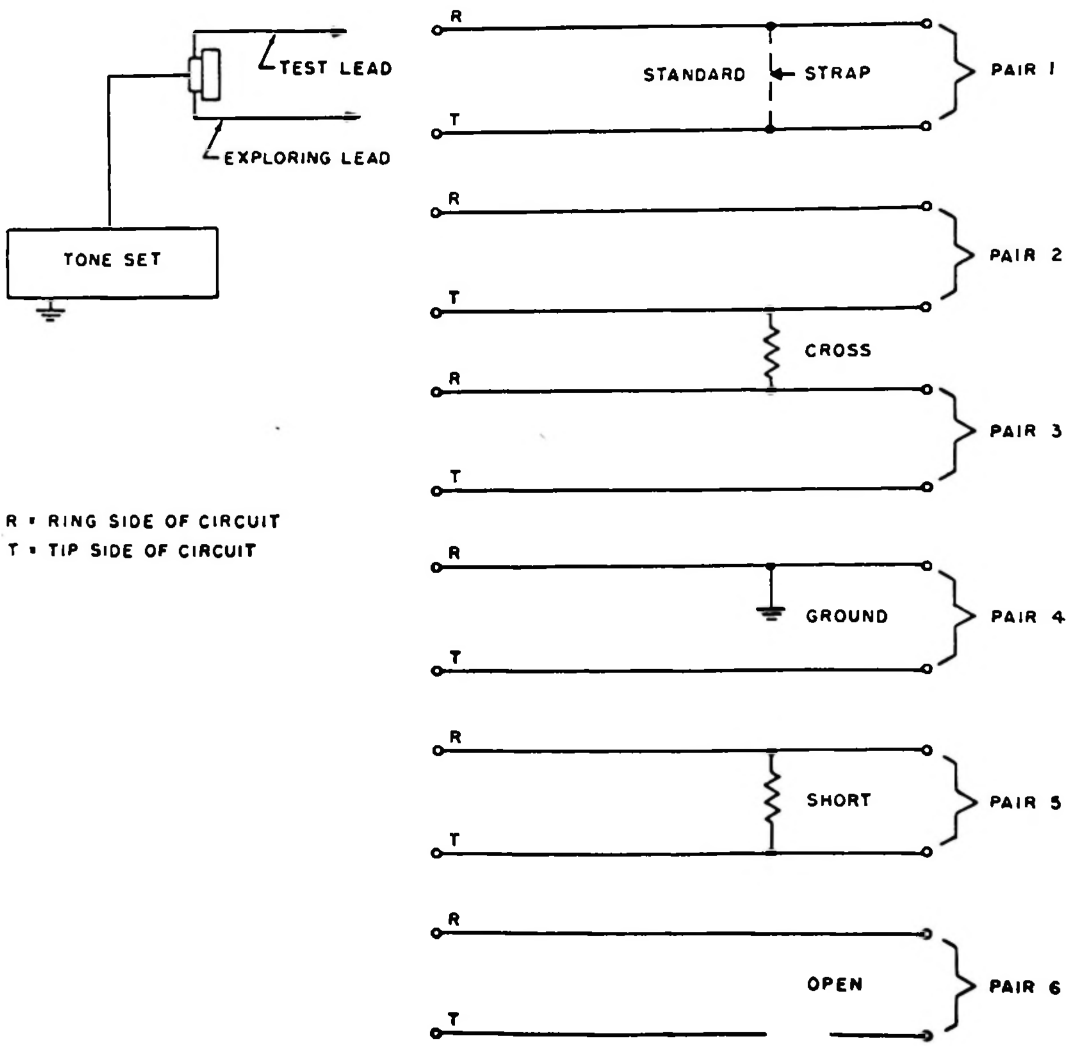


Figure 11-303. Tone Comparison or Balance Test Method of Fault Detection

tone on all of the pairs in the group. In such a case, a new standard pair within the group to be tested must be selected.

11-1976. After selecting the standard pair, the test lead is connected to the ring side of the standard pair. The exploring lead is connected to the conductor which is to be tested. If no tone is heard in the receiver for any of the tests, the pairs tested are free of trouble. If a tone is heard, it is possible that the conductors on which the tone is heard are defective or are not of the same gauge as the standard pair.

11-1977. An indication of no trouble in the circuit should be found when the two leads are connected to:

a. Ring or tip of pair 1 with the other side of pair 1.

b. Ring or tip of pair 1 with the ring side of pair 2.

c. Ring or tip of pair 1 with the tip side of pair 3.

d. Ring or tip of pair 1 with the tip side of pair 4.

e. Ring or tip of pair 1 with the ring side of pair 6.

f. Any conductor listed with any other conductor in the list.

NOTE

There may be a slight tone in the receiver for connection d above. This is because the ring side of pair 4 is grounded and some current may be induced in the tip side of the circuit.

11-1978. If trouble is indicated in the circuit, it can be localized by the procedure in which the test lead is connected to the tip or ring side of the standard pair and the exploring coil is connected to: the tip side of pair 2 or the ring side of pair 3, the ring side of pair 4, the ring or tip side of pair 5, or the tip side of pair 6. The trouble in the circuit is a cross, ground, short, or open, respectively. When testing for a short or cross, the standard pair must be strapped. After the standard pair is strapped, a standard tone is heard in the receiver. Then, if a pair of conductors balance with this standard tone, the conductors are shorted. If one conductor of a pair balances with the standard tone, the conductor is crossed with another conductor in the group.

11-1979. A grounded conductor will be indicated when a very loud tone is heard in the receiver for the first test. An open conductor is probable when a conductor cannot be balanced against the conductor of the strapped standard pair. This test method is not very accurate; therefore, when testing for an open it is advisable to use one of the other methods discussed in this section.

11-1980. EXPLORING COIL METHOD OF CABLE AND FAULT LOCATION. To locate a short in a cable, connect the tone unit in series with the faulty conductors as shown in figure 11-304. Then place the receiver plug in the proper jack of the exploring coil. Figure 11-305 shows the exploring coil in the proper position on the cable sheath. For a proper understanding of the method of lo-

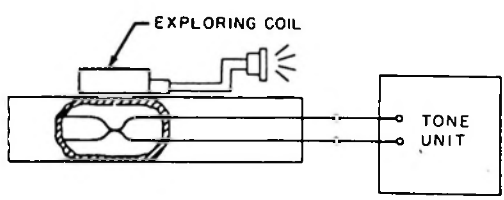


Figure 11-304. Exploring Coil Method of Locating a Short

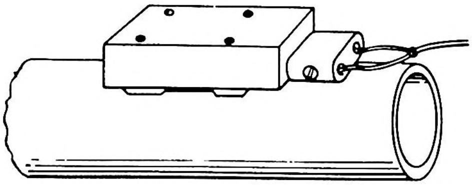


Figure 11-305. Correct Position of Exploring Coil on Cable Sheath for Location of a Short

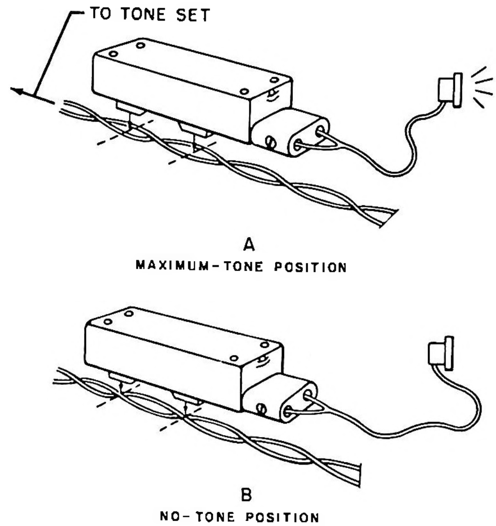


Figure 11-306. Short-Circuit Effects on Exploring-Coil Output

cating shorts with the exploring coil, it is necessary to consider the effect on the exploring coil of the currents flowing in the short-circuited conductors, as shown in figure 11-306. When the current is flowing in short-circuited conductors, maximum tone will be heard in the receiver if the exploring coil is in position A, with its pole pieces centered over two adjacent loops of the conductors. If the exploring coil is moved to position B, in which the pole pieces of the exploring coil are centered over the crossover points in a short-circuited pair, no tone will be heard. At intermediate points, the tone will be less than in position A, but more than in position B. If the exploring coil is moved steadily along the cable over the short-circuited pair, the volume of tone heard in the receiver will alternately rise and fall; this is called the short-circuit effect. Therefore, to locate the short circuit, move the exploring coil along the cable sheath from the source of tone toward the fault. The short-circuit effect will be heard up to the fault. Beyond the fault, no tone will be heard or the volume will be greatly reduced.

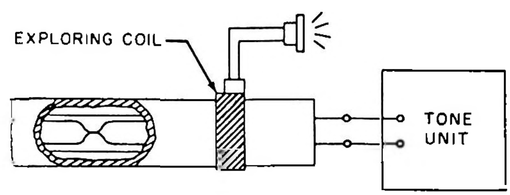


Figure 11-307. Exploring-Coil Method of Cross Fault Location

11-1981. To locate a cross in a cable, connect the tone unit in series with the faulty conductors as shown in figure 11-307. Then place the receiver plug in the proper jack of the exploring coil. Figure 11-308 shows the exploring coil in the proper position on the

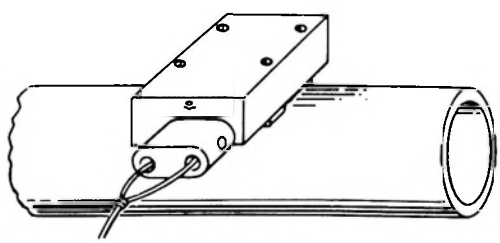


Figure 11-308. Correct Position of Exploring Coil on Cable Sheath for Location of Cross Fault

cable sheath. When current is flowing through two crossed conductors, the tone heard in the receiver will be approximately constant between the tone source and the fault. Beyond the fault, either no tone will be heard or the volume will be greatly reduced. The steady volume heard in connection with crossed conductors is called the cross-conductor effect.

11-1982. A ground in a cable can be located by connecting the tone unit in series with the faulty conductor and ground, as shown in figure 11-309. Place the receiver plug in the proper jack of the exploring coil, and hold the exploring coil in the proper position on the cable sheath as was illustrated in figure 11-305. Move the exploring coil along the cable sheath until the tone, which has been heard in the receiver, disappears or suddenly decreases in volume. This indicates the exact point of the fault. In some cases, the change in volume is so slight that the location of the fault is uncertain. In such cases, connect the tone unit first at one end of the cable and then at the other end of the cable, and check both location points to see whether they coincide. When testing aerial cable, hold the exploring coil on the underside of the cable to minimize interference due to the flow of current in the supporting strand.

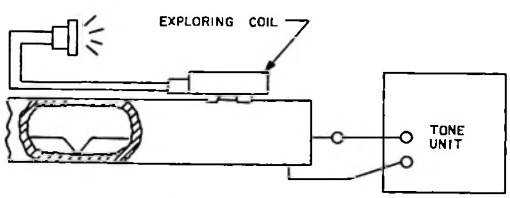


Figure 11-309. Exploring-Coil Method of Ground Fault Location

11-1983. To locate the point where conductors are split, strap the four conductors at the far end of the cable and connect the tone unit in series with one pair. Place the receiver plug in the proper jack of the exploring coil. In figure 11-310 the cross-conductor effect will be obtained up to the fault point, and thereafter the short-circuit effect will be obtained. An alternate method, shown in figure 11-311, may also be used. In this method, the short-circuit effect will be obtained up to the fault, and the cross-conductor effect thereafter. A change in the volume of the tone will be heard at the fault location.

11-1984. To locate wet spots in a cable, select a number of pairs of conductors having the lowest insulation resistance, divide these pairs into two equal groups, bunch each group, and connect the tone unit in series with the bunched groups as shown in figure 11-312. Place the receiver plug in the proper jack of the exploring coil, and hold the exploring coil on the cable sheath in the same position as when locating a

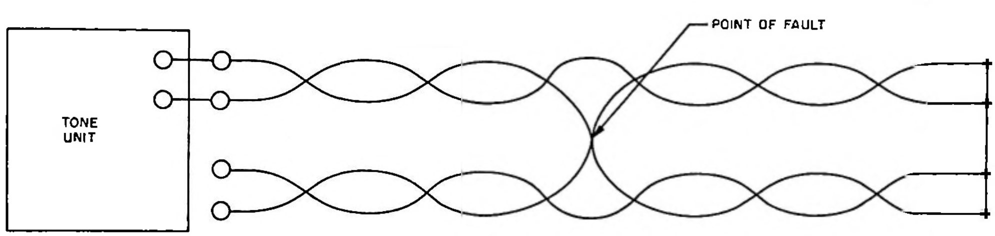


Figure 11-310. Location of Split Pairs by Exploring-Coil Method

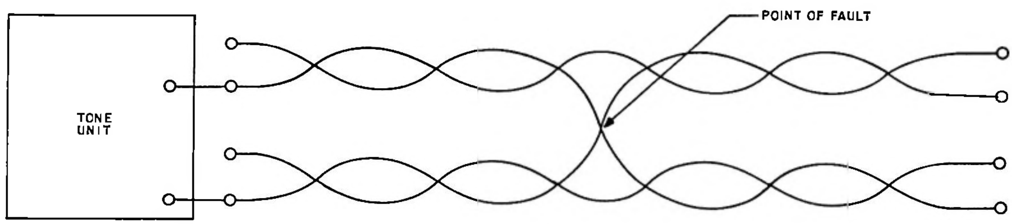


Figure 11-311. Alternate Circuit Connections for Locating Split Pairs with an Exploring Coil

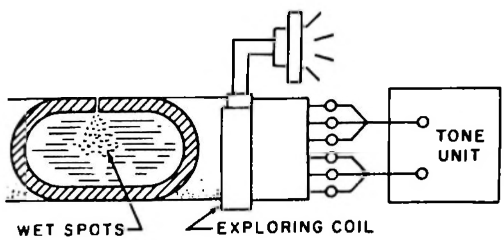


Figure 11-312. Location of Wet Spots with an Exploring Coil

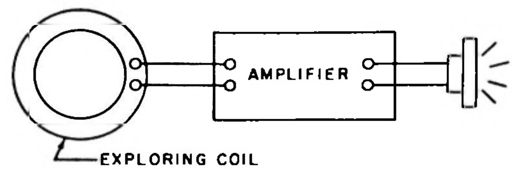


Figure 11-313. Exploring Coil and Associated Equipment for Locating Buried Cable

cross. A tone will be heard in the receiver up to the point of the fault. No tone should be heard in the receiver beyond the location of the fault.

11-1985. When locating buried cable, the tester must set up a circuit similar to the one for locating a fault in a cable. In this case, you will then walk, with an exploring coil in your hand, in the approximate area of the buried cable. You will be able to follow the path of the buried cable by listening for a tone in the receiver because a tone in the receiver will indicate the presence of buried cable at that point. The exploring coil used in this procedure is different from that used in locating faults in a cable. The exploring coil and associated equipment used for locating buried cable are illustrated in figure 11-313. This exploring coil is not a standard item issued with a test set; you must make it in the field. The information necessary for making this exploring coil may be found in the applicable handbook for operation of the test set.

11-1986. VOLTAGE BREAKDOWN TEST METHOD OF FAULT LOCATION. When the voltage breakdown test method is used, it is important that you perform the following procedures prior to the test:

a. Isolate each pair or pairs of conductors to be tested from the central office equipment, either by removing the cross-connections at a cross-connecting terminal or by removing the heat coils and protector blocks at the central office.

b. If the heat coils and protector blocks are removed at the central office, dummy blocks must be substituted. A warning sign, as illustrated in figure 11-314, shall be placed at the protector blocks of the pair or pairs to be tested, to warn airmen that a breakdown test is in progress. The type illustrated has the dummy block feature incorporated in it. Do not remove the warning sign until the breakdown test is completed.

c. Notify all personnel whose telephones are connected to the circuit of the pair or pairs to be tested, before interrupting service. Obtain the location of the cable terminals at which the pair or pairs to be tested appear, in accordance with the detail plans, and remove all drops from these terminals. Disconnect all dead cables associated with the pair or pairs to be tested.

d. Rearrange any conductors which are in contact with the binding posts of the pair or pairs to be tested. Determine whether the pair or pairs to be tested terminate in any textile-insulated conductor forms. If this is the case, determine whether the fault

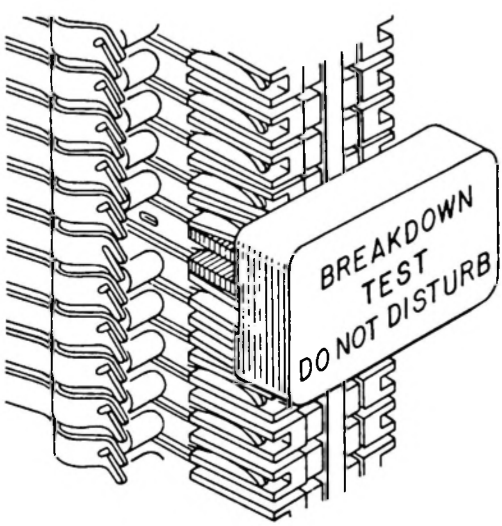


Figure 11-314. Warning Sign for Breakdown Test

is in the form. If you cannot fully determine that the fault is in the form, apply a breakdown voltage at the point of the fault and observe for evidence of breakdown. Be sure that no airmen are working on the cable to which the breakdown voltage is applied.

e. Operate the test set from a car or truck. If this is impracticable, place the test set on a rubber blanket on the ground. Notify all airmen who are involved in the breakdown test when the breakdown test has been completed.

11-1987. Following the breakdown test, connect the test clips of the line cord to the pair or pairs under test as shown in figure 11-315, and insert the line plug into the line jack on the breakdown test set. Measure the resistance of the fault, and apply voltage to the cable circuit. If the fault does not break down, check the resistance of the fault. If the fault resistance has decreased, resume operations on the same pair of conductors. If the fault resistance has increased, transfer your operations to another pair. If operations on a second pair are not successful, try a third pair. Check the resistance of the conductor when a fault breaks down. The approximate distance to the fault may then be calculated by multiplying the resistance (in ohms) by the feet/ohm ratio of the loop. Table 11-18 lists the feet/ohm ratios for various cable conductor gauges. The tone may now be applied and the exact fault location determined by the exploring coil method. Turn the set off and disconnect the test set from the pair or pairs which were tested, after the fault location has been determined.

11-1988. When making tests on coaxial cable by the voltage breakdown method, the procedure is similar to that used on paper-insulated cable. The main difference between the tests is that a higher voltage is required for coaxial cable.

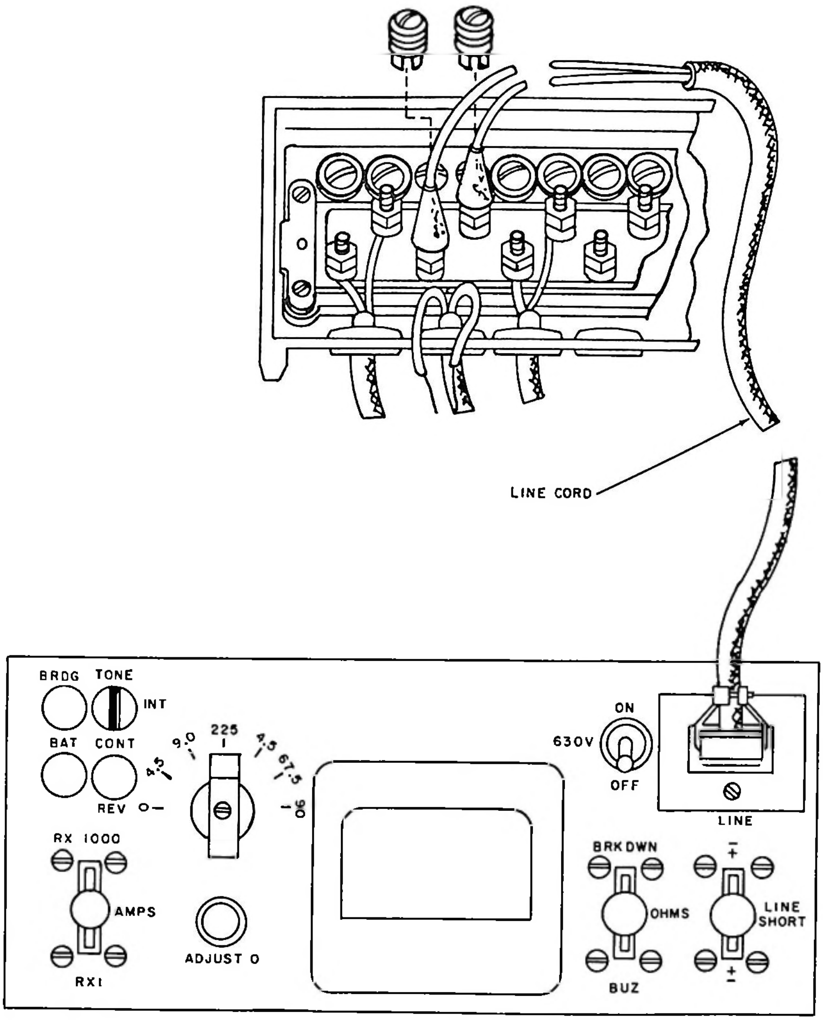


Figure 11-315. Connections to Test Pair in Voltage Breakdown Testing Method

11-1989. After the breakdown tests have been completed and the fault corrected, the pair or pairs which were tested should be restored to service. The heat coils should be replaced in the central office equipment, and the warning sign should be removed.

11-1990. MEASUREMENT OF INSULATION RESISTANCE. Insulation resistance measurements can be made by either a tester or a splicer. You should use a megger or a voltmeter to perform the test. These tests should be made periodically on existing cable circuits and on a new cable before and after the cable is installed.

11-1991. Splices that have been boiled with paraffin should be allowed to cool before insulation tests are made on the conductors. If desiccant was used in the splice, you

Table 11-18. Cable Conductors—Loop Feet per Ohm

CABLE CONDUCTORS (ga)	LOOP FEET PER OHM
13	245
16	125
19	60
22	30
24	20
26	12

should allow for error in the insulation test if the test is made immediately after application of the desiccant. This is due to the fact that the insulation does not dry thoroughly until two or three days after application of the desiccant.

11-1992. NUMBER AND LOCATION OF TESTS. In general, the length, size, and type of splice made in a cable will determine the number of insulation tests to be made. However, when a splice is made in a wet manhole or under bad weather conditions, it will be necessary to perform tests more frequently than when splices are made under normal conditions. When splicing a new cable to an existing cable, the new cable should be tested for insulation resistance. If low insulation resistance is found in the new cable, the fault should be cleared before the connecting splice is made. The cable should also be tested before and after it is joined to a terminating cable.

11-1993. When working with a distribution cable, an insulation resistance test should be made at least once every eight or ten sections. When working with a trunk cable, an insulation resistance test should be made at least once in every loading section. In addition, each group of four loading sections should be tested after completion. After each group of four loading sections is added, the entire cable should be tested. To detect any damage that might occur during the process of splicing, it is advisable to test the trunk cable daily, at the end from which the splicing work was started. These daily tests will be made on groups of conductors rather than on individual conductors. The tests are made between groups, and from these individual groups to ground. An insulation resistance test should be made on a trunk cable after the cable has been completely spliced. This test should be made between each conductor and the remainder of the conductors, both bunched and grounded.

11-1994. PREPARATION OF CONDUCTORS. When making an insulation resistance test on a distribution or trunk cable, you must clear the ends of the conductors at the far end of the cable; dry and wrap the conductors. At the testing end of the cable, strip off about 12 inches of sheath, as shown in figure 11-316. Then, remove the insulation from the ends of the conductors for approximately 3 inches. For distribution cable, bind the bare ends of the conductors into bunches of 100 pairs according to color groups. For a trunk cable, bind the bare ends of the conductors into bunches of about 20 quads.

11-1995. When making an insulation resistance test on a coaxial cable, in connection

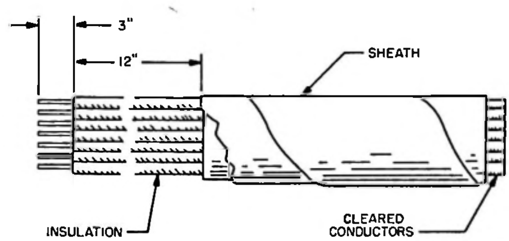


Figure 11-316. Preparation of Trunk or Distribution Cable Conductors for Insulation Resistance Tests

with the splicing operation, proceed as follows:

a. If the end of the cable is sealed with solder, cut off approximately 10 inches of the cable; then remove about 12 inches of the sheath, as shown in figure 11-317. Protect the edge of the sheath by placing a few layers of freshly boiled 1/2-inch cotton tape on the end edge. This tape will prevent the edge of the sheath from cutting into the conductors.

b. Bend the paper-insulated conductors over the edge of the sheath. This will expose the coaxials. Place two or three turns of 1/2-inch electrical tape on each coaxial, 6 inches from the end edge of the sheath. These layers of electrical tape are called tape collars. Remove the steel tape on each coaxial up to the tape collar. This procedure is shown in figure 11-318.

c. Remove the outer conductor, 1 inch from the tape collar. Push the disk nearest the outer conductor under it so that the disk is flush with the end. Cut off the inner conductor approximately 1-5/8 inches from the tape collar. Remove about 6 inches of insulation from the ends of the paper-insulated

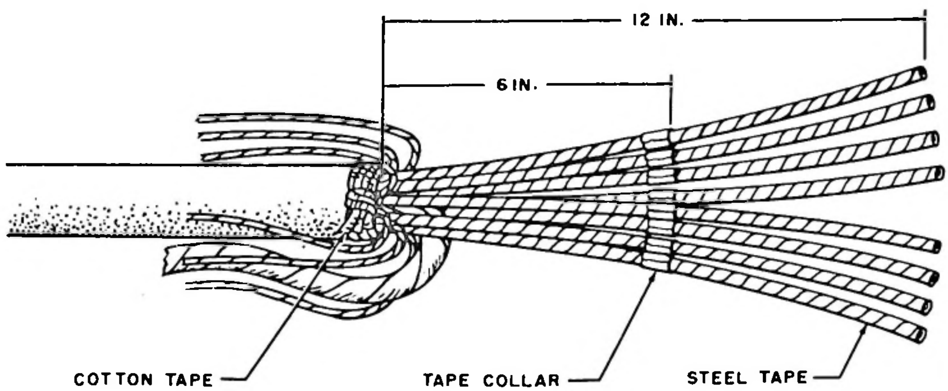


Figure 11-317. Preparation of Coaxial Cable Conductors for Insulation Resistance Tests

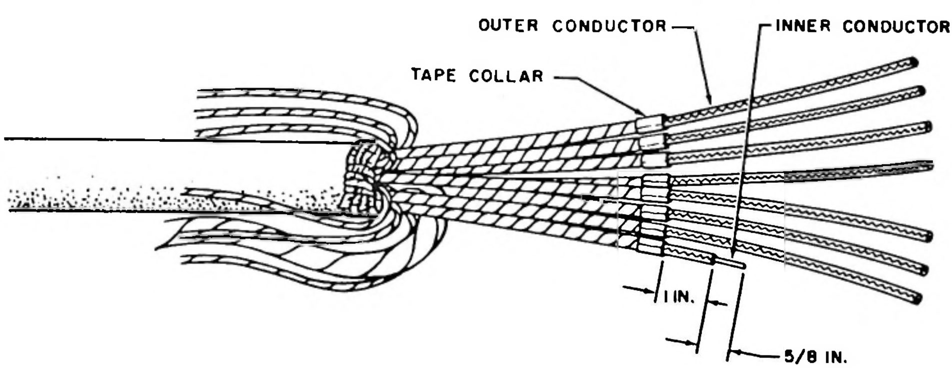


Figure 11-318. Preparation of Coaxial Cable Conductors with Steel Tape Removed

conductors, and bunch the conductors. Wrap a bare conductor around the conductors of the paper-insulated conductors and the outer conductors of the coaxials near the sheath. Ground the bare conductor to the sheath. If the cable has steel tapes, armor conductors, or a copper jacket, the bunched conductors should also be connected to this outer metallic covering.

11-1996. If the insulation resistance tests are to be made considerably in advance of the splicing operation, the two ends of the cable should be prepared. Keep the ends of the coaxials clear at the distant end by placing an 8-inch length of tubing under the end of the outer conductor of one coaxial for a distance of 1 inch, as shown in figure 11-319. Number 4 size tubing is used for 0.270 coaxials, and 5/16-inch tubing is used for 0.375 coaxials. Place the free end of the tubing in a second coaxial. Prepare the other coaxials in a similar manner. Wrap the coaxials with polyethylene tape, and prepare the ends of the paper-insulated conductors. Then wrap the paper-insulated conductors with polyethylene tape, after applying the required amount of desiccant. Cap the end of the cable with a lead sleeve or its equivalent. A talking circuit should be established between the testing end and other points when the tests are made with the cable open at the far end or intermediate point, or if the work is to be performed at a location where it is not possible for the tester to see the splicer handling the leads.

11-1997. TESTING PROCEDURES. When measuring the insulation resistance of a cable with a megger, proceed as follows:

a. Prepare the conductors and adjust the megger. Connect the line side of the megger to the bare end of the conductor or bunch of conductors under test. Connect the ground side of the megger to the sheath of the cable, as shown in figure 11-320.

b. Turn the crank of the megger until the indicator on the dial of the megger remains fairly constant. While still turning the crank, take a reading on the dial. If the insulation resistance is low, the indicator reaches a constant value quickly and does not fluctuate with continued turning of the crank. If the insulation resistance is high, it takes a little time for the indicator on the dial to reach a constant value. Before removing the megger lead after the test is completed, make sure that the conductors are discharged by allowing the generator to come to a stop and then momentarily connecting the exposed end of the conductor to the line terminal. After completing the test,

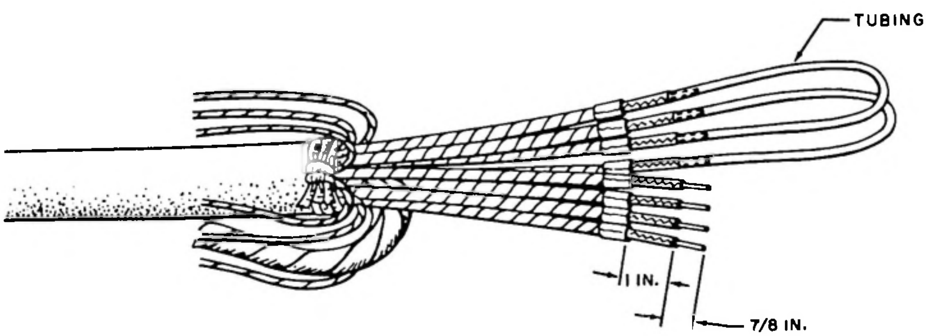


Figure 11-319. Use of Tubing to Protect Coaxial Cable Conductors

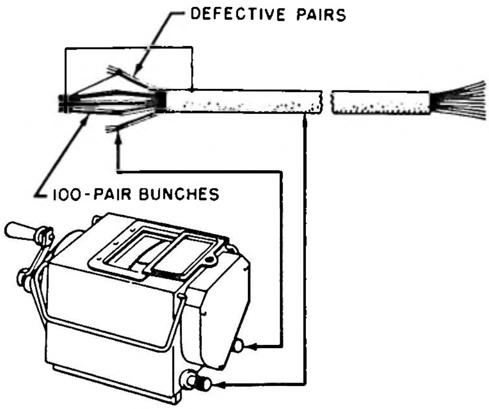


Figure 11-320. Measurement of Insulation Resistance, Using a Megger

disconnect the megger lead and connect the lead to another conductor or group of conductors which must be tested.

11-1998. For short sections the reading on the dial should be above 1000 megohms, and for long sections it should be above the resistance listed in table 11-19. If the reading on the dial is less than that listed in the table, separate the bunch of conductors which is low and test individual conductors of this bunch against all other conductors grounded to the sheath. Each conductor should then give a reading at least as high as that listed in the table. If the reading for an individual conductor is less than that listed in the table, test the conductor against the other bunches (after disconnecting the bunches from the sheath) and then against the sheath alone. This will indicate whether the fault is between the conductors, between the conductor and sheath, or both. Test each of the remaining bunches. After testing the bunches, test a few conductors of each bunch against the remaining conductors of the same bunch to determine whether there is any low insulation resistance within the bunches.

Table 11-19. Insulation Resistance Requirements

LENGTH OF CABLE (ft)	REQUIRED INSULATION RESISTANCE (megohms)	LENGTH OF CABLE (miles)	REQUIRED INSULATION RESISTANCE (megohms)
500	5,280	1	500
1,000	2,640	2	250
1,500	1,760	3	167
2,000	1,320	4	125
3,000	880	5	100
4,000	660	6	88.3
5,000	528	7	71.3
6,000	440	8	62.5
7,000	377	9	55.5
8,000	330	10	50
9,000	293	11	45.4
10,000	264	12	41.7

11-1999. If a low insulation resistance fault cannot be located and cleared by the resistance measurement or exploring coil method, the megger method may be used as follows:

a. Remove the sleeve and muslin wrappings from the splice nearest the middle of the section, and pick out and open several of the parts in which the low insulation resistance has been isolated.

b. Test to determine on which side of the splice the fault is located. After determining this, open the splice half-way between the middle splice and the end of the section in which the fault lies. Continue opening splices until the faulty splice is located. Then dry and wrap the faulty splice.

c. If the insulation resistance test at the end of the section indicates that the low insulation is confined to the conductors at the center of the cable, the fault may be caused by insufficient boiling-out at one or more of the splices. If the low insulation is found in pairs in the outside layer of the cable, the fault may be caused by the use of muslin which has not been sufficiently boiled out, or it may be in the sheath.

11-2000. The voltmeter method of measuring insulation resistance is similar to the method of measuring the unknown resistance of a communication circuit. If R_X is substituted for R_L, the insulation resistance may be calculated by using the following equation:

$$R_X = \frac{R_M (V_B - V_M)}{V_M}$$

where R_M, V_B, and V_M represent the same quantities previously defined.

11-2001. INSULATION RESISTANCE COMPUTATION. The insulation resistance required for a new cable is 500 megohms for 1 mile of conductor (500 megohm-miles). This is the minimum insulation resistance between individual conductors, and between the sheath and all other conductors at 16 degrees C (60 F). The insulation resistances for cables of various lengths are listed in table 11-19. This table is based on the requirements necessary for a new cable. When more than one conductor is tested at a time, divide the resistance (megohms) listed in the table by the number of conductors in the group to find the insulation resistance required for the group.

11-2002. When a cable is equipped with carrier loading having relatively short spacing, the insulation of the carrier-loaded conductors may be somewhat lower than the values listed in table 11-19. Since quads equipped with carrier loading are relatively few, the insulation resistances of the remaining quads will give a good indication of the insulation resistance of the cable. When testing the insulation resistance of an individual loading coil or of a loading-coil-case stub cable, a reading of 1000 megohms is satisfactory.

11-2003. The minimum insulation resistance for any length or size of cable may be computed in the following manner: The minimum insulation resistance between one conductor (1 foot long) and the cable sheath is 2,640,000 megohms. Therefore, to find the minimum insulation resistance required for one insulated conductor of any length, it is only necessary to divide 2,640,000 by the length of the conductor expressed in feet. If the insulation resistance of an entire cable is desired, it is necessary to divide the insulation resistance for one conductor by the total number of conductors in the cable.

11-2004. TESTING IN TELEPHONE CENTERS.

11-2005. The tests which can be performed with the test equipment already installed in the telephone centers are as follows:

a. Test calls can be originated and answered on magneto and command battery loops.

b. Tests can be made with a voltmeter or ohmmeter to check the continuity of a circuit and detect accidental grounds, short circuits, crosses, or opens.

c. Ringing voltage can be applied to a circuit to detect grounds, short circuits, or opens.

d. The capacitance from a line to ground or to another line can be indicated by the voltmeter needle swing in conjunction with the initial application of a potential through the voltmeter to the line. The approximate location of an open circuit on an open-wire line can be estimated from the capacitance test results.

e. Conductor loop resistance and insulation resistance can be measured with a multimeter or a Wheatstone bridge.

11-2006. When a loop or trunk is out of order, the following procedures may be adopted to clear the trouble. The trouble is analyzed by making over-all ringing and talking tests. Then specific tests are performed to localize the trouble. In a small center the tests can be performed with a voltmeter at the switchboard; in a large center the tests can be performed with the test unit or the test board. The line can be sectionalized and analysis tests can be performed on each section. After you have determined the section containing the trouble, use the testing equipment nearest that section to isolate the trouble.

11-2007. The test and control board equipment provides jacks and patching cords for removing circuits from service, such as may be required for maintenance tests, routine tests, or transmission measurements. The test and control board equipment also provides arrangements for monitoring and talking on the lines. The monitoring jacks may be arranged on some test boards to send a busy signal to the associated switchboards, so as to avoid interference between regular traffic and test calls. In centers where automatic make-busy features are not provided, it will be necessary for you to ask the operator to make the circuit busy at the switchboard.

11-2008. TESTING OF RELAY SWITCHING CIRCUITS.

11-2009. The manual type of test set used for unit testing is described below. The test set has a pin jack field into which the numbered wires of the connecting cable can be individually plugged, in order to connect the test set terminals to the proper terminals of the relay under test. The other end of the connecting cable is equipped with a contact fixture arranged to give quick electrical connections to the terminals of the wired relay unit. The plugging of the pins into the proper pin jacks is a feature needed to provide flexibility in a test set arranged to test many types of circuits, and is a part of the setup operation for any one circuit.

11-2010. The test set is equipped with signal lamps for visual-response indications and manually operated keys for the use of the tester in performing the test operations. Separate power cords are plugged into power distribution jacks which supply the various potentials commonly used in telephone offices. After the initial setup, operate the numbered keys and observe the lamp signal responses in accordance with the chart supplied with the set. Failure to obtain a particular lamp indication requires that you analyze the circuit conditions and locate the cause of the trouble. Figure 11-321 shows a small portion of a simplified circuit test arrangement for a manual test set. In this illustration a single key, when operated,

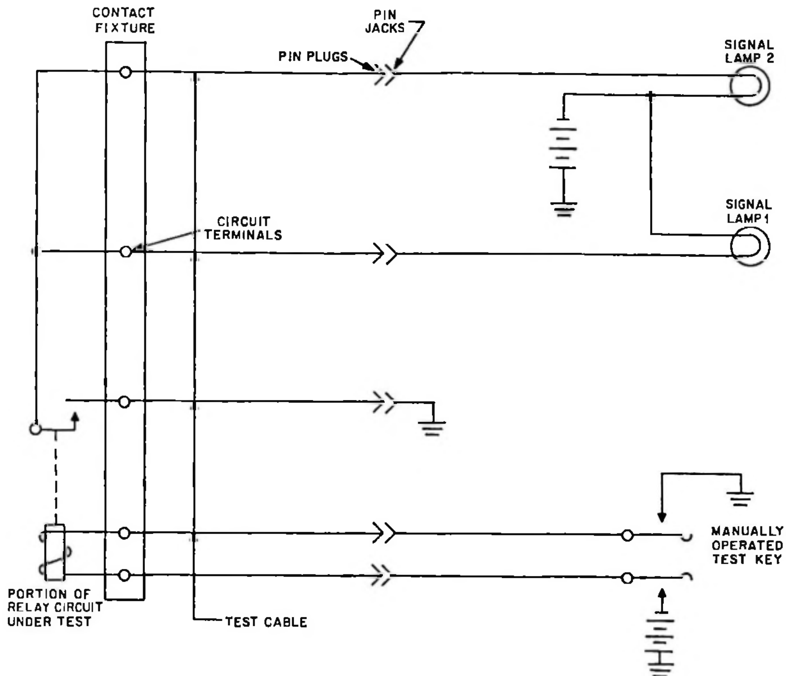


Figure 11-321. Simplified Circuit Diagram for Testing Manual Relay Switching

supplies battery and ground potentials to the inductive winding of a relay in the circuit under test. Assume that the three relay contact terminals are wired directly to the relay unit terminal strip so that they can be connected to ground and to the battery through lamps for circuit closure indications. The switching functions of the relay can then be checked by operating the test key and observing that signal lamp 1 extinguishes and that signal lamp 2 lights.

11-2011. The key and visual lamp indicating functions of the manual test set can be replaced by relays in an automatic test set, which perform these operations if they are under the control of suitable programming and advancing circuits, as shown in figure 11-322. The signal relay operates through the contacts of the relay under test, and their operating positions are checked by the watching relays, whose contact closures must match those of the signal relays. The series path through the contacts of all signal and watching relays is called a chain lead. The program circuit establishes the positions of the watching relays to meet the expected conditions prior to operating the key relay. Any lack of continuity through the chain lead caused by failure to satisfy test conditions halts the progress of the test. Additional contacts on the signal and watching relays may be used to light the signal lamps, in order to convey information to the tester concerning which portion of the circuit fails to operate properly.

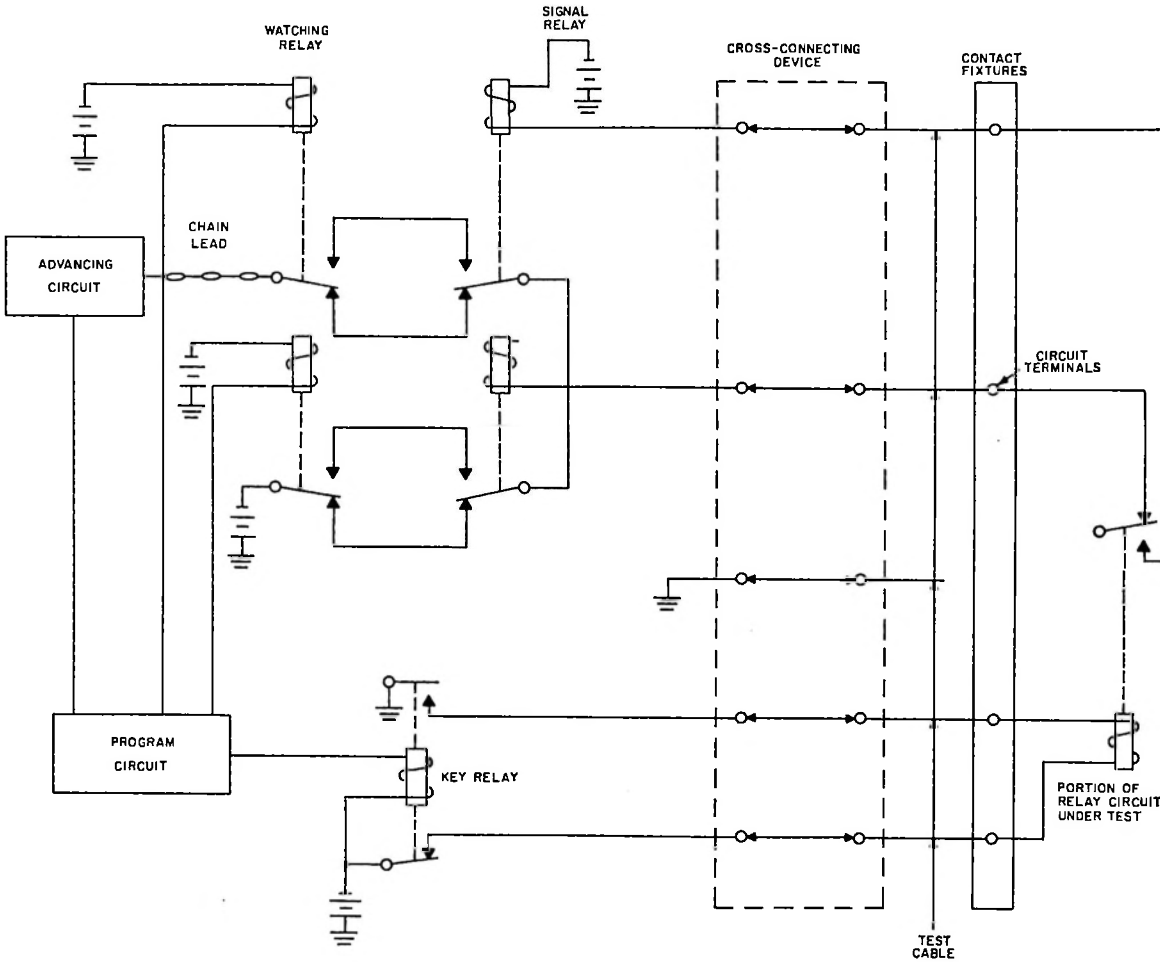


Figure 11-322. Simplified Circuit Diagram for Testing Automatic Relay Switching

11-2012. TELEGRAPH TESTING.

11-2013. LINE RELAY TESTING.

11-2014. Clean the line relay by removing the relay cover and blowing out any accumulated dust. Wipe the relay and the cover with a clean, soft cloth. Next, take out the armature contact screws to permit the entrance of a contact file, and use the file to remove pits and any build-up of material on the contacts. While you are cleaning the armature contacts, support the armature at its midposition by means of the opposite contact screw. This is to avoid bending of the armature. Also, care should be taken in filing the armature contacts. After you have removed all pits and build-ups on the contacts, blow any loose particles away from the equipment, and polish the contacts with a burnisher. Remove any particle adhering to the armature or pole-piece screws by wrapping a fresh piece of electrical tape around a thin, stiff, nonmagnetic metal, and pressing it against the particles. Do not rub the tape against the armature or the pole-piece screws, because such rubbing will leave a residue which will collect more particles.

11-2015. Loosen the screws that hold the spool heads to the relay form, and position the spool so that the armature does not touch the inside of the spool. Then loosen the armature clamping screws and position the armature so that the center of the contact will not be out of alignment by more than 25 percent of the contact diameter. When you have the contacts aligned, tighten the clamping screws. Then bend the outer armature spring contacts either toward or away from the inner armature spring contact, so that the outer contact is parallel to the armature and as close as practicable to the point where the contact spring is riveted to the armature. Back off the pole-piece screws and position the contact screw between 0.003 inch and 0.005 inch, to provide clearance between the line relay armature and the outer contact screw until it pushes the armature far enough to just touch the left-hand contact screw. Then back off the right pole-piece screw about 1/4 turn and advance the left pole piece until the armature is centered in the magnetic field. A cross section of a line relay is shown in figure 11-323.

11-2016. DISTORTION TESTING.

11-2017. Two types of signals, used to determine the performance characteristics of the digital data/telegraph equipment, are provided by the distortion generator; one is a continuous train of signals,' and the other is a standard test message. In either case the signal may be distorted with either marking bias or spacing bias.

11-2018. The typical distortion generator functions in the test message reversal mode and also in the normal test message mode as follows:

a. The standard frequency output of the distortion generator is 128 times the center frequency of the selected band. This output is applied to the timing circuit, which produces one complete square-wave cycle for each bit of the generator output signal. The time-A output of the timing circuit is applied to the reversals circuitry, where it produces alternate marks and spaces to the test-message circuits at the selected frequency, thus producing a character test message at the frequency of the selected band. The time-B signal applied to the distorting circuits provides a 1/2 bit delay in the output signal when the output from the generator is being distorted. A block diagram of a distortion generator is shown in figure 11-324.

b. The triggering circuits accept the character-advance signal and use it to reset the timing circuits to begin each character.

c. The output from either the reversals or the test-message circuits is selected by switch S1 and applied to the distorting circuits. The output from the distorting circuits is coupled to the output circuits. The output circuits provide a transmitter-keyed output polar signal at a certain voltage.

d. When operating in the switch mode, the character-advance signal applied to the distorting circuits alternates the type of applied bias.

11-2019. The distortion testing oscilloscope is used to check for distorted waveforms. This permits you to perform a visual check of not only the distorted current magnitude, but also of other telegraph set output waveform characteristics.

11-2020. Distortion current amplitude and other wave characteristics of the presentation can be analyzed by using the calibrated

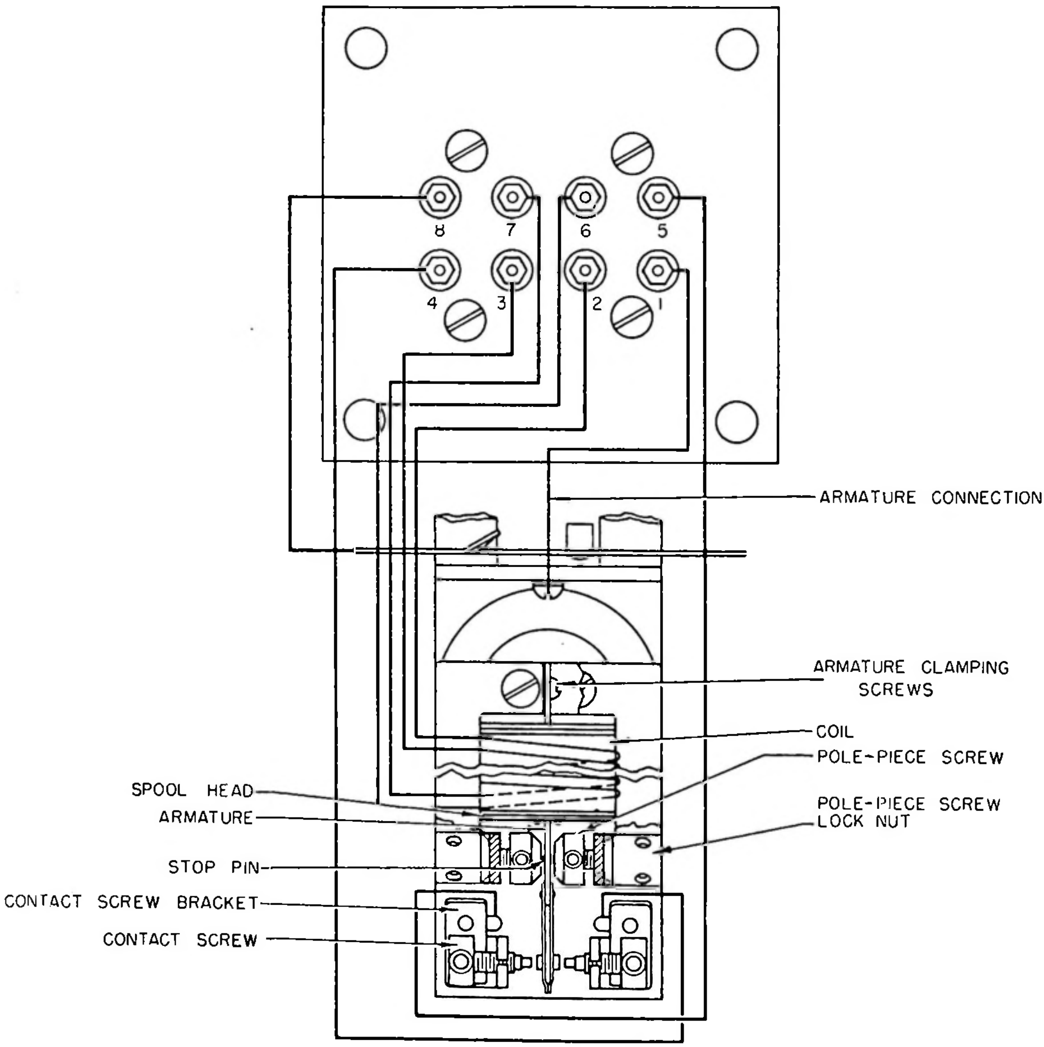


Figure 11-323. Cross Section of a Line Relay

scale overlaid on the screen of the cathode-ray tube. This scale is calibrated in both time and current magnitude; also, current indications are provided for both negative and positive signals. With the presentation centered vertically on the scale, and the vertical gain control at its extreme clockwise position, the current amplitude can be determined. A block diagram of a distortion test oscilloscope is shown in figure 11-325.

11-2021. The digital distortion tester analyzes the distortion present in the start-stop data/telegraph signals and provides an output indication in either percent distortion, as registered on a front-panel meter, or a

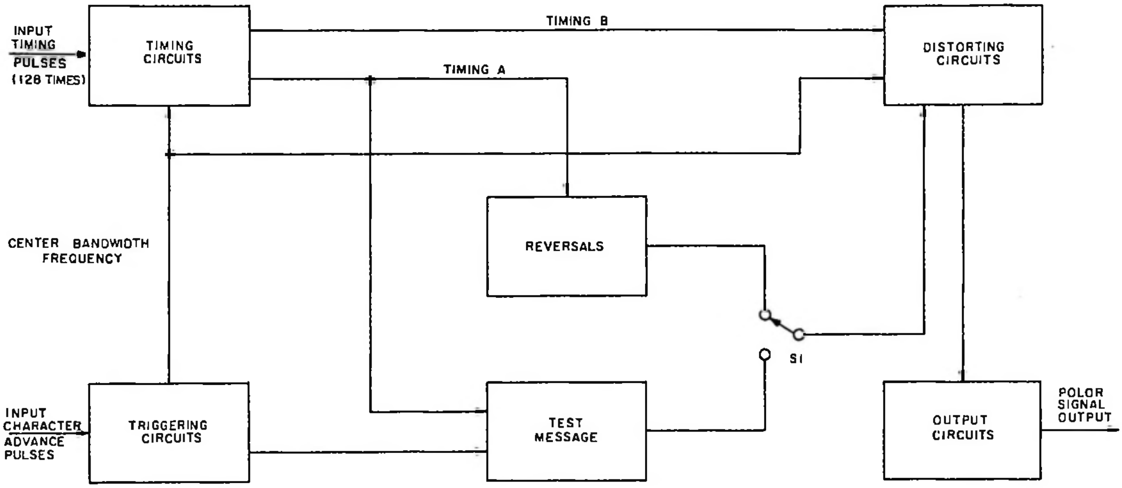


Figure 11-324. Block Diagram of a Distortion Generator

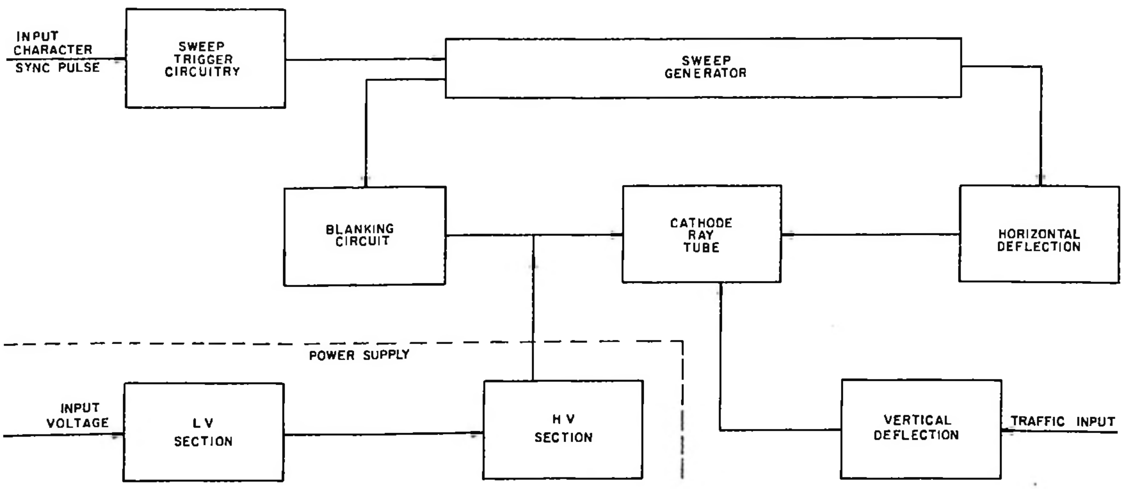


Figure 11-325. Block Diagram of a Distortion Test Oscilloscope

type of average distortion, as identified by the lighting of either of two front-panel lights. The unit may also be used to measure average distortion occurring at each transition of a character, or at the first, second, etc, transition of successive characters.

11-2022. Figure 11-326 is a functional block diagram of the digital distortion tester. The signal to be tested is applied to the loop input; a time base signal at a multiple of the center frequency of the band of the signal under test is applied to the ideal timer. This unit will provide a percent distortion output indication to the meter, depending on the mode of operation. By comparing a data/telegraph signal with an internally generated ideal signal, the digital distortion tester will permit you to determine the type and degree of distortion present in the data/telegraph signal. To attain a high degree of accuracy, each bit of the ideal signal is divided into segments. The meter then indicates a distortion proportional to the displacement of the center of the segment from the ideal transition time. During peak distortion measurements, the meter permits only distortion which is greater than that already indicated on the distortion tester to be read into the unit.

11-2023. CARRIER EQUIPMENT TESTING.

11-2024. The voice signal enters the carrier transmitter arm through the back ter-

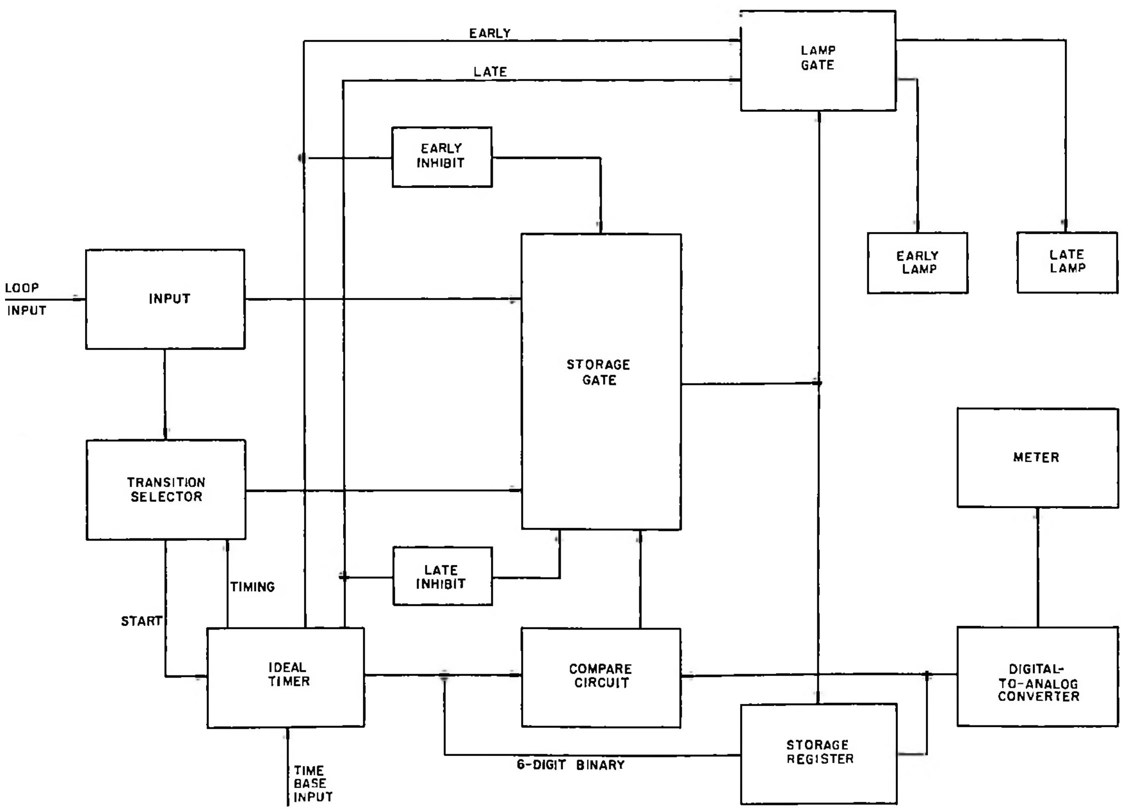


Figure 11-326. Block Diagram of a Digital Distortion Tester

minals on most sets, and may be observed at the hybrid jack. You can also inject a voice frequency at the same hybrid jack. After the voice signal passes through the hybrid assembly, you may observe the signal at the A attenuator line jack shown in figure 11-327. If the signal injected at the hybrid jack is correct for your equipment, and if the A attenuator control is adjusted properly, you should have correct attenuation at this point. The voice signal now passes through the A attenuator and into the low-pass filter, at the output of which the signal is corrected. The voice signal then passes into the modulator, where it is modulated with the carrier signal from the carrier oscillator. This modulated signal is passed through the B attenuator and on to the bandpass filter. The modulated signal then passes through the transmitter amplifier. At this point the attenuation should be zero for most types of carrier equipment.

11-2025. The received carrier signal is applied through the back of the set, and can be measured at the equalizer gain control and through the receiver amplifier, where the gain and amplitude are corrected. The test signal is then applied to the bandpass filter and the C attenuator, where the undesired frequencies are removed. The test signal is then applied to the demodulator, where the voice signal is removed from the carrier. The voice signal is then sent through the low-pass filter and the voice-frequency (vf) amplifier. At the output of the vf amplifier the voice-frequency signal can be observed.

11-2026. CARRIER-LEAK TESTING.

11-2027. In an ideal situation there is no carrier leak, but in a practical situation each channel of the carrier equipment does have a small output at the carrier frequency. Any output at the carrier frequency is known as a carrier leak. When making the carrier-leak check with the decibel meter for any one channel, the other carrier frequencies are shorted out. The reading on the decibel meter should not exceed the maximum reading given for your carrier equipment. A carrier-leak test arrangement is shown in figure 11-328.

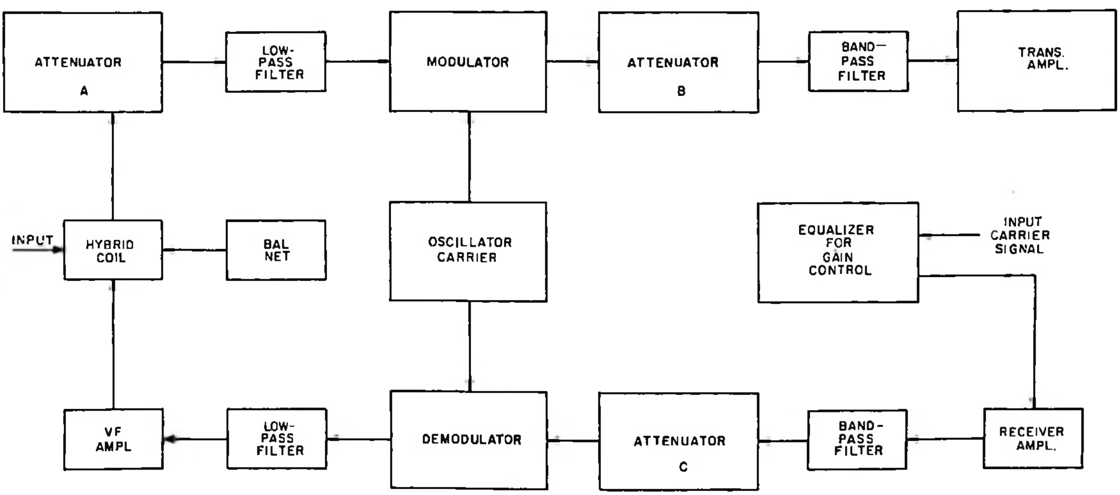


Figure 11-327. Block Diagram of a Carrier Equipment Transmitter and Receiver

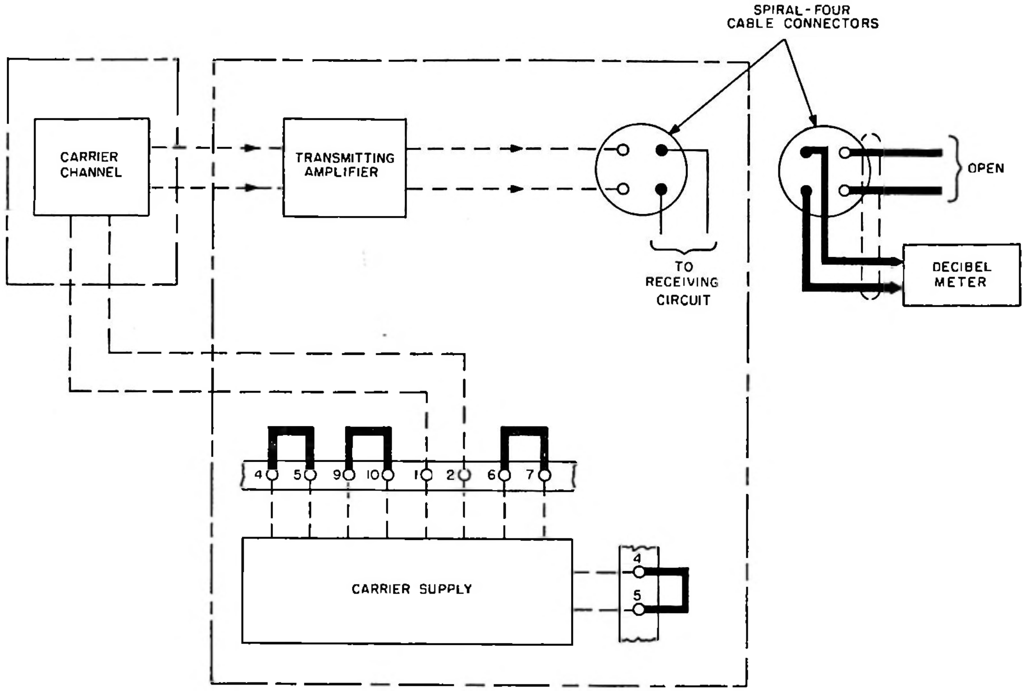


Figure 11-328. Carrier-Leak Test Arrangement

11-2028. LOOP TESTING.

11-2029. When trouble has been isolated to a particular terminal of the carrier equipment, a useful procedure in trouble shooting is to loop the transmitting branch to the receiving branch on the line side of the equipment. The carrier equipment can then be tested without dependence on, or interference from, the other terminals in the equipment. This procedure is subject to the considerations and limitations discussed in the following paragraphs.

11-2030. The transmitting and receiving branches must be opened at the point where they are cross-connected. This can be accomplished by removing a plug-in unit from the succeeding equipment, by removing a test plug, by opening a normal-through jack, or by disconnecting wires at the terminal block, whichever can be accomplished most easily. The levels in the transmitting and receiving branches must meet the requirements specified in the equipment alignment procedures. This may require adjustment of gain controls to provide attenuation in the equipment, or external attenuation may be used if available.

11-2031. MULTIPLEXING EQUIPMENT TESTING.

11-2032. The primary method used to test multiplexing equipment is through the use of preventive maintenance routines for the individual equipment. Visual inspection for charred or blistered components, broken wires, or other obvious mechanical damage is also used. Multiplexing equipment is, in most respects, maintenance-free. For this reason, the few tests performed on multiplexers are designed for the individual set. Figure 11-329 illustrates a block diagram of a typical multiplexing set.

11-2033. Some tests that are appropriate for multiplexing equipment are voltage checks, continuity checks, and signal tracing techniques.

11-2034. MICROWAVE TESTING.

11-2035. In a microwave transmitter, such as the one diagrammed in figure 11-330, the input modulating voltage is applied directly to the video modulator. Within the video modulator the signal is amplified to the level necessary to modulate the klystron. The tone oscillator output signal is applied to the input of the video modulator. This signal is necessary for proper operation of the combiner in the diversity control unit.

11-2036. The klystron produces the carrier output. The equipment uses an automatic frequency controlling device to maintain a stable center frequency. The local oscillator produces a highly stable frequency, which is multiplied in the waveguide to produce a frequency higher than the transmitter signal. Within the mixer in the waveguide assembly a portion of the transmitter output signal and the local oscillator output signal heterodyne together. The output i-f frequency is applied to the i-f amplifier, where the signal level is increased sufficiently to operate the automatic frequency control unit and the video monitor equipment.

11-2037. The automatic frequency control and video monitor equipment use a delay line discriminator to permit you to determine from the i-f signal whether or not the klystron is operating at the correct frequency. If the klystron frequency is correct, the output of the automatic frequency control and video monitor will not change. However, if the output frequency of the klystron increases or decreases, the automatic frequency control and video monitor output error voltage will reflect an output change, which can be used to correct the dc klystron voltage so that the output signal from the

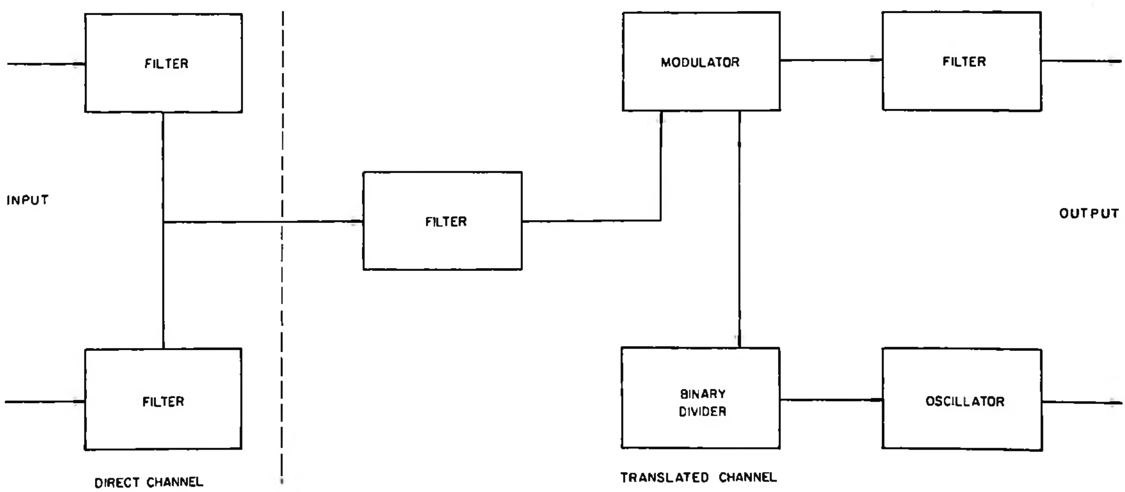


Figure 11-329. Block Diagram of a Typical Multiplexing Set

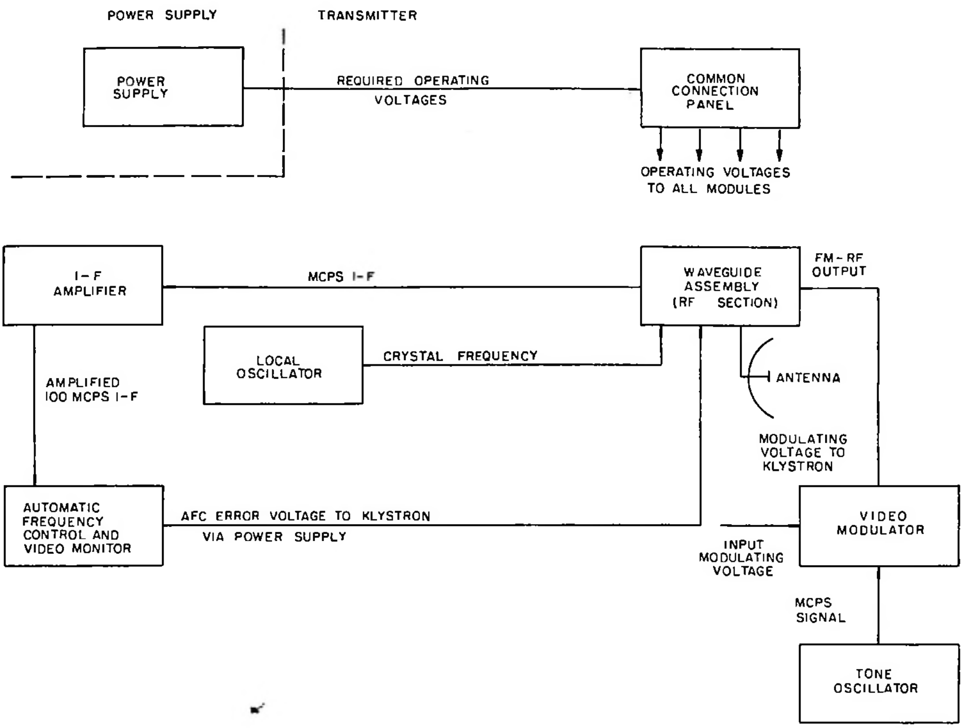


Figure 11-330. Block Diagram of a Microwave Transmitter

klystron will return to the correct value. The signal is directed into the waveguide assembly, where it is heterodyned with the output frequency of the local oscillator. The difference frequency is fed through a cable to the preamplifier. The preamplifier, with a low-noise front end, amplifies and applies the i-f signal through the bandwidth reduction filter and phase equalizer to the i-f amplifier. At the same time, additional circuits in the amplifier provide an automatic gain control function. Controlling the i-f gain of the receiver shown in figure 11-331 minimizes the effect on the output for a given change in the input. The same basic circuits in the i-f amplifier which develop the automatic gain control voltage are used to operate the squelch circuit. The squelch circuit removes the plate voltage from the video amplifier when the input signal is below the desired squelch level.

11-2038. MICROWAVE IMPEDANCE TESTING.

11-2039. Figure 11-332 shows a typical slotted-line measurement setup. The addition and cancellation of the incident and reflected waves within the transmission mode will create a mismatch with the output load. Errors may occur at several places in the slotted-line measurements. The sources of these errors are probe loading, harmonics, frequency modulation, detector characteristics, and other spurious signals. There will naturally be some power pick-up by the sampling probe supplying the indicating devices, and this will also set up reflections

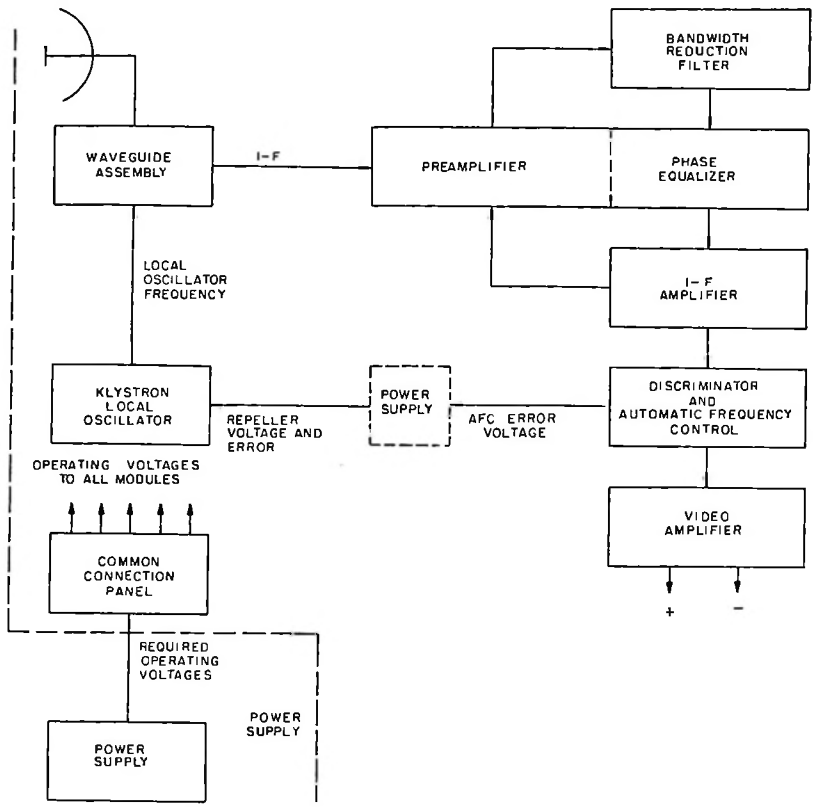


Figure 11-331. Block Diagram of a Typical Microwave Receiver

on the line. These errors will increase as the insertion of the probe is increased. This makes it necessary to maintain probe penetration at a minimum.

11-2040. Because the ratios of different voltage levels are being measured with the aid of slotted lines, it is essential that the detection follow the same law for all levels. If barretter type detection units are operated at levels less than 200 microwatts and the associated crystals are operated at a power level of less than 20 microwatts, the characteristics are essentially those of a square law device. Spurious signals and undesirable harmonics can be reduced to a minimum by the use of low-pass filters.

11-2041. A reflectometer is another instrument used for measuring impedance. The reflectometer indicates the impedance magnitude, but it does not provide phase angular information as does the slotted-line method. The reflectometer method of measurement is useful for fast-swept frequency production measurements.

11-2042. Figure 11-333 shows a typical reflectometer setup. The directional couplers sample both the input and the reflected wave.

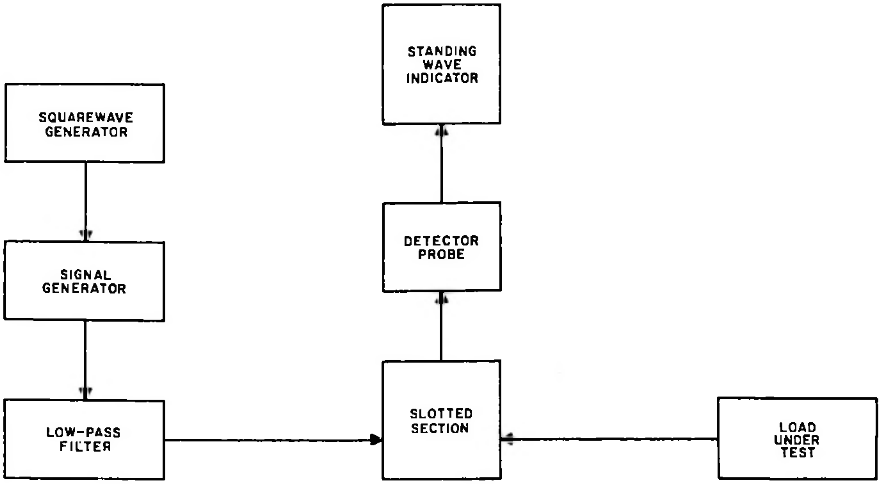


Figure 11-332. Typical Slotted-Line Testing Setup

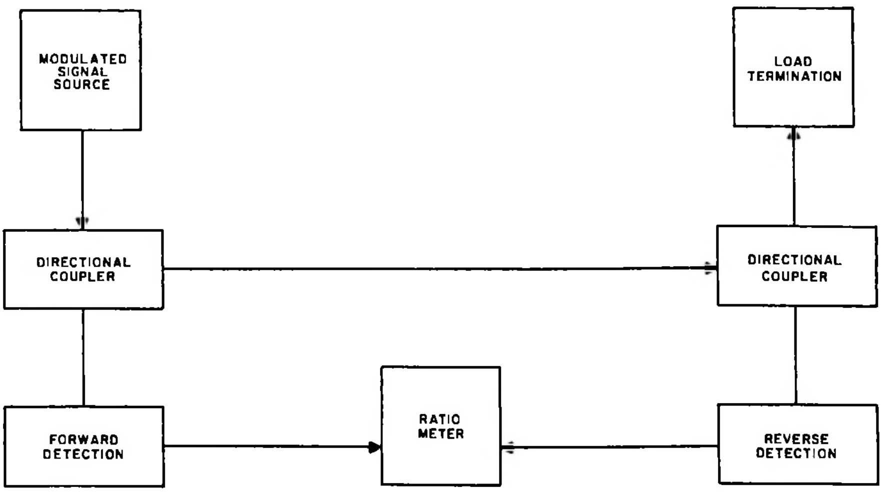


Figure 11-333. Typical Reflectometer Setup with Two Directional Couplers

The outputs from the couplers are fed to the detectors and then to a ratio meter. The resultant ratio of the two sampled signals is then read directly on the meter. The input power should be maintained at a level of about −20 dbm for forward detection and at a level of about −10 dbm for reverse detection. By maintaining the input at these stipulated levels, you will have assured close approximation to square-law operation of the crystal.

11-2043. Since slotted sections below 500 mc are excessively long, other methods of impedance measurement are more desirable. One such method uses a vhf bridge. You can adjust the two available controls until a sharp null is obtained. At this null, the unknown impedance in ohms can be read from one dial, and the phase angle can be read from the other. The vhf bridge measurement setup is shown in figure 11-334.

11-2044. Since the voltages measured during the null are very small, the lines connected to the bridge should be well shielded to protect them from extraneous voltages. The signal to the bridge should produce a sharp null. As the bridge is basically an unbalanced device, the measurements in the balanced equipment can be made with a balun. A half-wave balun is shown in figure 11-335. The balun structure operates as a

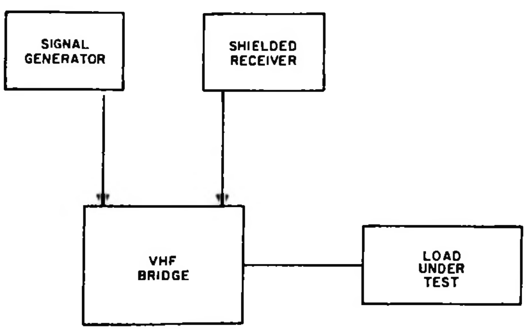


Figure 11-334. Test Setup of VHF Bridge

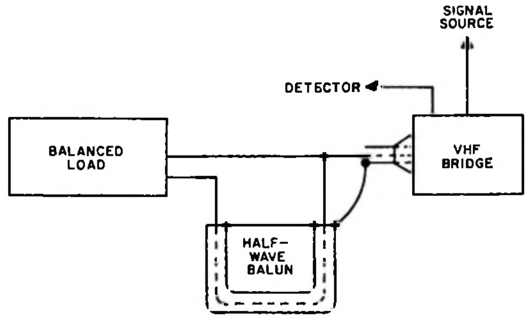


Figure 11-335. Half-Wave Balun with VHF Bridge

4-to-1 impedance transformer, thus, an impedance measured at the balun must be multiplied by 4 to obtain the actual impedance.

11-2045. MICROWAVE TEST SETUPS.

11-2046. Figure 11-336 shows the test setup used to measure the Q of a microwave cavity. The frequency-modulated klystron input to the cavity is monitored on the oscilloscope so that you can plot the response curve of the cavity. The output of the precision cavity wavemeter is applied to the z-axis input of the oscilloscope. This provides a small dot on the screen of the oscilloscope to indicate the frequency of the precision cavity. The 3-db power points, representing the effective voltage of the response, are then determined as follows: The frequency difference in the micrometer reading of the precision cavity required to move the dot on the oscilloscope from one 3-db point to the other is converted into the frequency difference by the following equation:

$$\text{Q of the resonators} = F/\Delta F$$

where F is the frequency of either 3-db point and Δ F is the change of frequency from one 3-db point to the other.

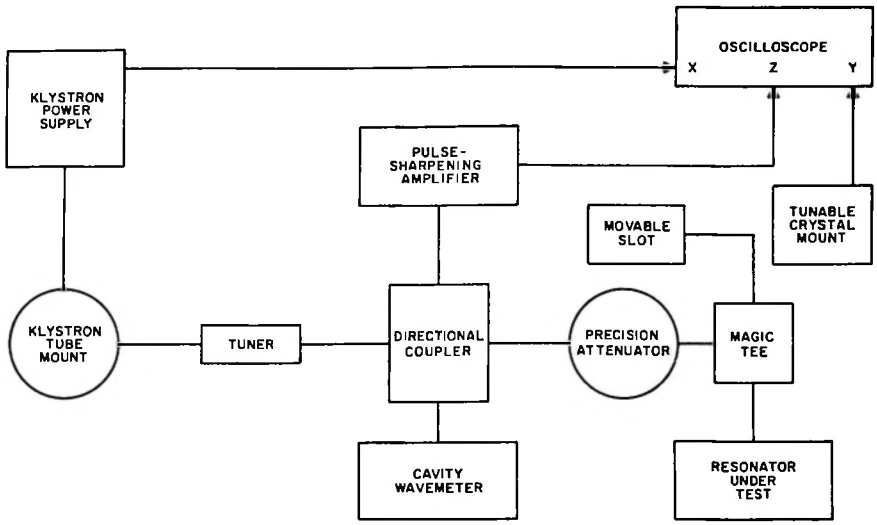


Figure 11-336. Test Setup to Determine the Q of the Cavity

11-2047. Figure 11-337 shows the test equipment setup used to perform a spectrum analysis. The sawtooth repeller voltage from the power supply frequency-modulates the klystron. The sharply tuned i-f amplifier will amplify the mixer output only at an instant when the klystron frequency differs from that of the input signal frequency.

11-2048. Figure 11-338 shows the test setup required to determine the antenna pattern of an individual microwave set. The automatic plotting of an antenna radiation pattern uses a linear attenuator geared to a recording pen. The setting of the linear attenuator is continuously adjusted through the action of a servomotor so that it maintains a constant i-f amplifier output. The drive mechanism is arranged to move the recording paper a distance which is proportional to the angular rotation of the antenna under test. This arrangement provides a plot of the antenna pattern as a function of angle.

11-2049. Figure 11-339 illustrates the test setup required to measure noise in the output of the microwave mixer and i-f amplifier. The noise power is measured by the thermocouple microwattmeter or the square-law detector followed by the milliammeter calibrated at the i-f frequency. The figure representing the amount of noise from the mixer and the i-f amplifier is obtained by measuring the exact amount of power required to double the reading of the power indicator.

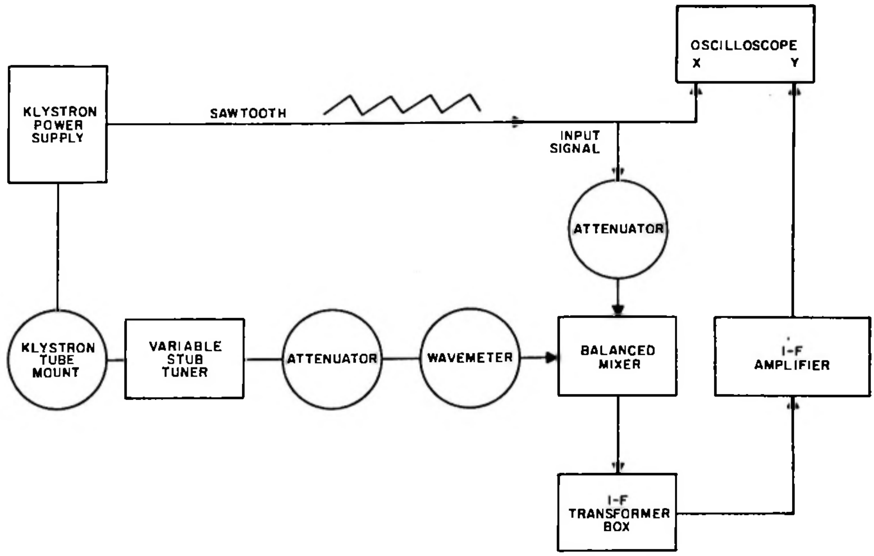


Figure 11-337. Test Setup for Spectrum Analysis

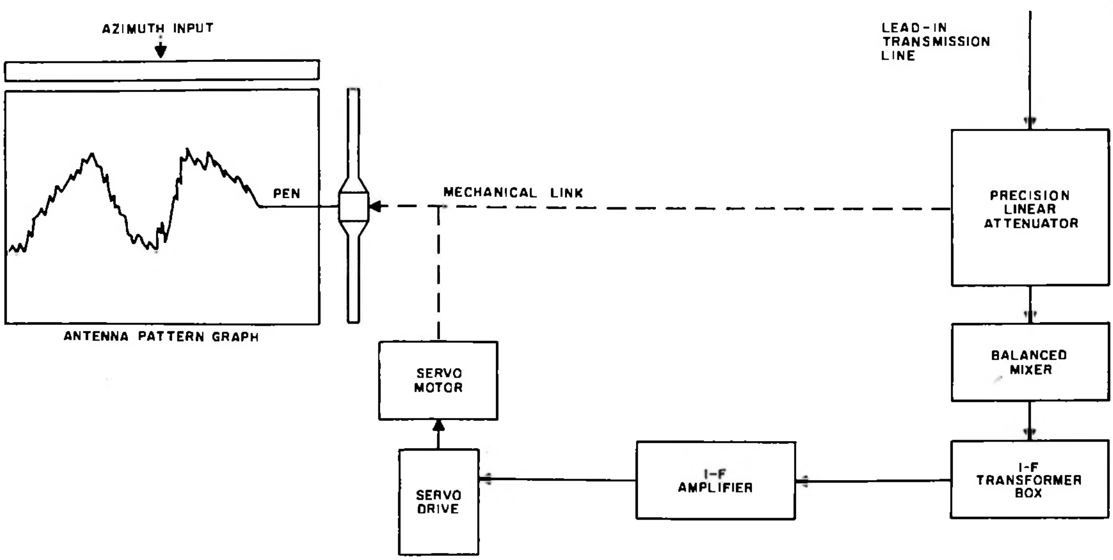


Figure 11-338. Antenna Pattern Test Setup

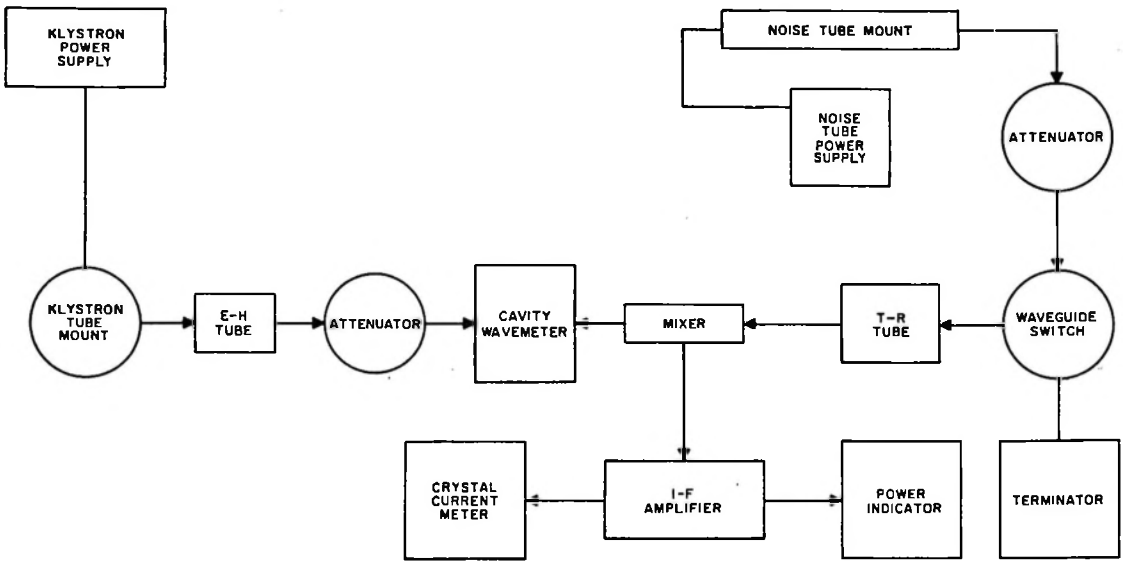


Figure 11-339. Microwave Mixer and I-F Amplifier Noise Measurement Test Setup

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