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CHAPTER 12 MERIT MEASUREMENTS

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CHAPTER 12

MERIT MEASUREMENTS

CHAPTER 12

MERIT MEASUREMENTS

12-1. INTRODUCTION. This chapter explains and discusses the factors contributing to the quality of communication/electronic/meteorological equipment operation. The information is divided into 11 sections. Complex waveform and phase development is covered in Section I. Factors contributing to distortion are explained in Section II. Stability parameters are discussed in Section III. Transient response is discussed in Section IV. Spectrum and spectrum analysis techniques are given in Section V. Noise figure measurement is covered in Section VI. The determination of impedance and distributed constants at microwave frequencies is explained in Section VII. Gain measurement information is given in Section VIII. Attenuation is discussed in Section IX. Time interval measurement is covered in Section X. External performance factors are discussed in Section XI.

SECTION I

COMPLEX WAVEFORM AND PHASE DEVELOPMENT

12-2. GENERAL.

12-3. PURPOSE.

12-4. INTERPRETATION. The primary objective of this section is the interpretation of the common voltage and current waveforms as observed on an oscilloscope or other test equipment. A complete understanding of the appearance and reason for the normal waveform will prepare you to recognize an abnormal waveform, and may possibly help you understand the reason for the abnormality. Terms such as "loss of high frequencies" and "loss of response" will have special meanings when referenced to a particular waveform. The causes of deteriorated voltage and current waveforms will be discussed.

12-5. PRESENTATION. This section about complex waveforms and phase development illustrates and explains the normal waveform and abnormal waveform, as well as their causes. It does not discuss or describe trouble-shooting procedures because it is assumed that you will inaugurate your own trouble-shooting procedure, using the knowledge of waveforms contained in this section. This section does not permit a mathematical analysis of waveforms; it is presented as an aid to those individuals having a vital interest in waveforms, and who cannot, or do not, desire to interpret mathematical explanations.

12-6. DESCRIPTION.

12-7. GRAPHICAL. A voltage or current waveform, as encountered in the communications and electronics field, may be graphically represented in both the height and width dimensions. The height of a graphically displayed waveform represents quantity, or the amplitude of voltage or current; the width of the displayed waveform represents the elapsed time, or waveform duration.

12-8. PARAMETERS. The voltage or current waveform is normally represented in a two-dimensional (horizontal and vertical) plane with no depth involved. The horizontal ("X") axis on a graph will represent time measured, in either whole or parts of a second; the vertical ("Y") axis on a graph will represent amplitude, quantity, or intensity of the subject waveform measured, in either whole or parts of volts or amperes. Any portion of the waveform extending above the horizontal (zero amplitude) reference line is considered positive, while any portion of the waveform extending below the horizontal reference line is considered negative (opposite to the positive portion of the subject waveform).

12-9. COMMON WAVEFORMS.

12-10. SINUSOIDAL WAVEFORMS.

12-11. DESCRIPTION. The sine wave is the basis of all other waveforms. It represents the simple action of a swinging pendulum, a bouncing or vibrating spring, a free-running self-excited oscillator output, etc. When any outside source changes the shape of the sine wave, the wave is said to be distorted. However, you will find that the original sine wave is still present in combination with other sine waves introduced by the distorting agency to produce a single resultant waveform. Therefore, any waveform, no matter how complex, may be reduced to its individual sine wave components. The original sine wave components cannot be reduced further, because they are the final remaining single-frequency, basic components. No waveform that is composed of more than one frequency is a true sine wave.

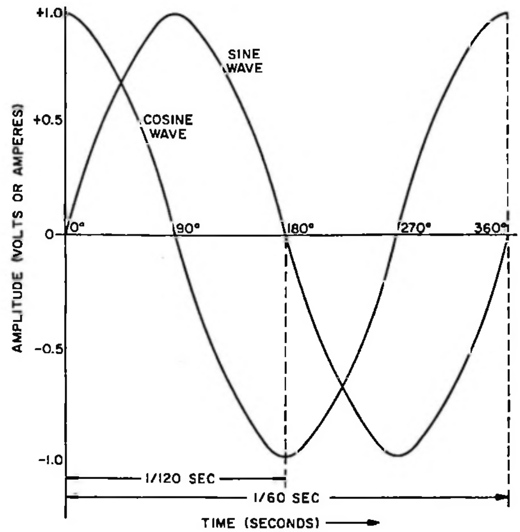


Figure 12-1. Sine and Cosine Waveforms

12-12. GRAPHICAL. Since the primary purpose of this portion of the section is to familiarize you with the basic structure of a sine wave, the sine wave is graphically presented in figure 12-1. A 60-cycles-per-second wave is represented by one complete cycle in figure 12-1. Therefore, the total time duration for this cycle is 1/60 of one second. Half of this cycle time is above the horizontal reference (zero amplitude) line and is considered positive, while the other half is below the horizontal reference line and is considered negative. These two halves of the sine wave do not cancel out or nullify each other because each half-cycle occurs during a different time. During the positive half-cycle, the negative half-cycle has not occurred, and, therefore, does not exist. During the time the negative half-cycle exists, the positive half-cycle does not. The single cycle illustrated in figure 12-1 must not have any bumps or kinks on either its increasing or decreasing side. The top (positive peak) and the bottom (negative peak) must be smoothly curved, with no appearance of either a point or a flat spot in this region. The positive and negative half-cycles of the sine wave must be exactly equal in both amplitude and time duration. In figure 12-1, the maximum positive peak amplitude is represented by positive 1 volt, while the maximum negative peak amplitude is represented by negative 1 volt. The time duration illustrated is exactly 1/120 of a second for each half-wave. Considering that a sine wave represents a complete mechanical revolution, or circle, it requires 90 degrees to complete 1/4 of the circle, 180 degrees to complete 1/2 of the circle, 270 degrees to complete 3/4 of the circle and 360 degrees (or back to 0 degrees) to complete the entire circle. There are an infinite number of points represented on a sine wave. For example, there are 360 different points, each representing an advance of 1 degree, on each sine wave. However, we have shown only four points to illustrate the shape of a basic sine wave. This was done because it is only necessary to familiarize you with the general features of the sine wave curve rather than provide a point-by-point analysis. Amplitude will not affect the general outline of a sine wave, provided that the positive and negative portions of the waveform contain equal amplitudes. If you view the sine wave on some form of oscilloscope, you may position the instrument controls incorrectly and present the wave as either too narrow or too wide for its height. This will not affect the actual waveform, but may affect your perspective; the peaks of the normal waveform may appear sharp or peaked for the narrow vertical version, and flat or broad on the wide or stretched horizontal version. Refer to paragraphs 8-567 through 8-569. A point-by-point examination of the wave will prove it to be a true sine wave, disregarding its misproportional appearance. A cosine waveform is also shown in figure 12-1. The cosine wave is the same in all respects as the sine wave except for one difference; it leads, or begins 90 degrees (1/240 of a second in this case) before, the sine wave time begins. The cosine wave is superimposed on the same graph as the sine wave to illustrate the cosine lead. These two waveforms are not used to provide a resultant waveform.

12-13. HALF-SINE WAVEFORM. The half-sine waveform consists of a series of unidirectional pulses, each of which resembles a half-cycle of a sine waveform. The half-sine wave may exist either above (positive) or below (negative) the horizontal reference line, as shown in figure 12-2. The half-sine waveform is produced by removing from the complete sine wave any amplitude variations in one direction for a period of one-half the time duration of the complete cycle. As illustrated in figure 12-2, there are two types of half-sine waves. In part A of the figure, the negative portion of the sine waveform has been removed and its 1/2-cycle time interval remains as a zero dc reference level. In this type of half-waveform, the frequency remains the same as the original full sine wave frequency. In part B of the figure, the negative half of the original full sine wave has been inverted over the horizontal reference line; consequently, the average dc voltage level is increased. Since each alternation occurs in 1/2 the time interval of the original full sine waveform, inverting the negative alternations to occupy the empty spaces between the positive alternations causes the frequency to double. Half-sine waveforms are composed of the original fundamental frequency in conjunction with a dc component and an infinite series of even-numbered harmonics of progressively decreasing amplitude.

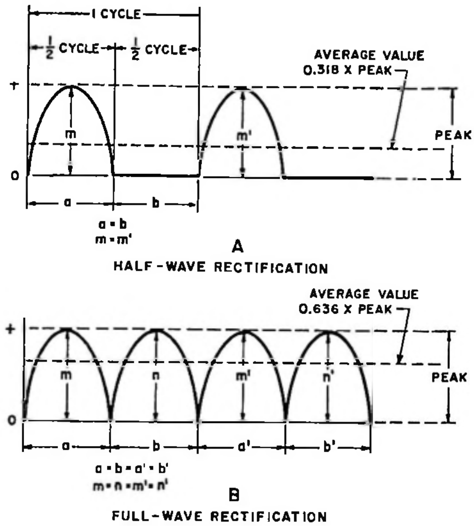


Figure 12-2. Half-Sine Waveforms

12-14. RESULTANT WAVEFORMS.

12-15. DESCRIPTION. The sine wave is the basic or standard alternating current (or voltage) waveform used in combinations with phase or time differences and amplitudes to algebraically form all other waveforms. The sine wave is the wave most commonly used as an input to circuits under test because it does not introduce distortions commonly associated with nonsinusoidal waveforms. All nonsinusoidal waveforms can be reduced into their individual component sine waves. A nonsinusoidal waveform is composed of more than one sine wave; other frequencies, usually harmonically related, are algebraically added to the fundamental frequencies to produce the resultant nonsinusoidal waveform. In this case, the sine wave of lowest frequency is normally considered to be the fundamental frequency, and higher frequencies which are exact multiples of the fundamental frequency are considered as harmonics of the fundamental. However, in some cases, the nonsinusoidal waveform being considered may be composed of only harmonic frequencies because the fundamental sine wave may have been intentionally removed. The algebraic addition of the fundamental sine wave (f) and the second harmonic of the fundamental (h = 2f) will provide a resultant nonsinusoidal waveform (R), as shown in figures 12-3, 12-4, and 12-5. However, only the resultant would be shown on an oscilloscope or equivalent test instrument. The second harmonic of figure 12-3 is shown in phase with the fundamental because its amplitude increases in the same direction as the fundamental from the horizontal and vertical zero reference level. The second harmonic of figure 12-4 is shown 180 degrees out of phase with the fundamental because it proceeds from the horizontal and vertical zero reference level in a direction exactly opposite (negative direction) from the fundamental. The second harmonic shown in figure 12-5 is shifted 90 degrees behind (lagging) the fundamental.

12-16. The resultant waveforms contained in figures 12-3, 12-4, and 12-5 are the only waves you will see on an oscilloscope or other equivalent test instrument. The amplitude of the second harmonic, relative to the fundamental sine wave, will either increase or decrease the amount of dip at points A and B of figure 12-4, represented by two heavy arrows. However, if the phase of the second harmonic is changed, with respect to the fundamental sine wave, the appearance of the resultant waveform will change completely. As shown in figure 12-5, the positive half of the resultant wave-

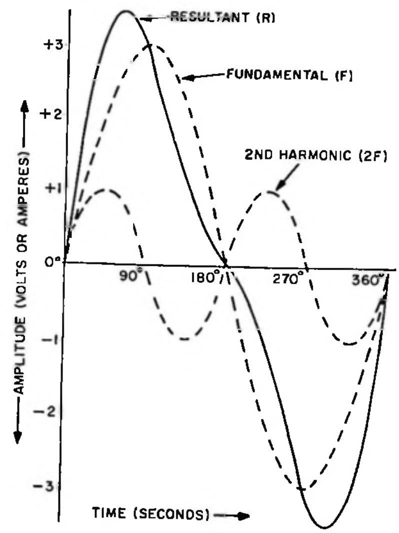


Figure 12-3. Waveform Resulting from Algebraic Addition of a Fundamental Sine Wave with Its Second Harmonic in Phase

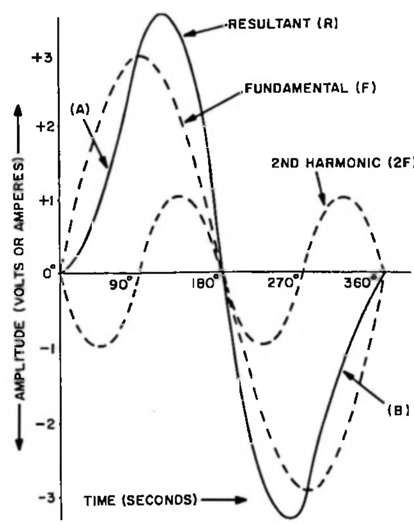


Figure 12-4. Waveform Resulting from Algebraic Addition of a Fundamental Sine Wave with Its Second Harmonic Delayed 180 Degrees

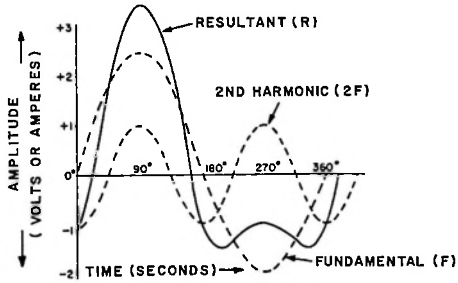


Figure 12-5. Waveform Resulting from Algebraic Addition of a Fundamental Sine Wave with Its Second Harmonic Delayed 90 Degrees

form looks like part of a sine wave, but the negative half does not. Therefore, this resultant is definitely not a true sine wave. At any point along the horizontal zero reference line you can add the vertical amplitude of the fundamental directly to the vertical amplitude of the harmonic to obtain the final amplitude of the resultant waveform at that particular point (that is, if f = 3 units and h = —1 unit, R = 2 units; if f = —6 units and h = —3 units, R = —9 units).

12-17. The resultant waveforms, illustrated in figures 12-3, 12-4, and 12-5, show that the algebraic addition of a harmonic waveform to the original fundamental sine wave produces a new waveform that is no longer a sine wave. These new waveforms, in many cases, are created deliberately to perform functions beyond the capabilities of the original signal. For example, the new waveform resulting from the addition of a fundamental and its harmonics may be used as timing pulses.

12-18. BASIC PHASE DISTORTION. Figure 12-5 represents the results of feeding the fundamental and its second harmonic through a network which delays the second harmonic by 90 degrees. This normal resultant waveform, obtained by the algebraic addition of the second harmonic (without phase shift) to the fundamental sine wave, is known as phase distortion. You can recognize this type of time delay only by becoming familiar with the proper waveforms. If the phase of a fundamental sine wave is shifted, its shape will not change. Therefore, special methods must be employed to recognize any changes in phase.

12-19. BASIC HARMONIC DISTORTION. The addition of harmonics to the fundamental wave shape creates a new resultant waveform. This resultant is a distortion of the original waveform, and, if undesirable, is termed harmonic distortion. You can probably recognize the resultant waveform created by the addition of only one harmonic to the original waveform. However, the addition of several harmonics to the fundamental sine wave, in and out of phase, will create a resultant waveform of pure confusion. As was mentioned previously, you can separate or remove any waveform from the resultant with the aid of suitable filters. Therefore, by removal of all except one frequency component, you can extract a pure sine wave of some frequency which was not evident in the original wave or its harmonics. This newly extracted sine wave can then become the fundamental sine wave input to a circuit under test.

12-20. COMPLEX WAVEFORMS. The resultant waveforms shown in this section are created either by adding harmonics to the fundamental waveform, by changing the phase of the harmonic with respect to the fundamental, or by a combination of harmonic addition and phase change. Therefore, all resultants are termed complex waveforms, no matter how simply or easily they are recognized. Actually, other effects are realized in the resultant waveform; depending on the addition of even (that is, 2nd, 4th, 6th, 8th, etc) harmonics or odd (that is, 3rd, 5th, 7th, 9th, etc) harmonics, on the percentage of harmonic waveform amplitude injected, and on the phase of the introduced harmonic with respect to the fundamental sine wave. All so-called distorted waveforms are classified as complex waveforms. However, complex waveforms are grouped into types of complex waveforms. To familiarize you with some of the primary features of nonsinusoidal waveforms consisting of a single harmonic addition to the fundamental sine wave, a series of illustrations are provided in figures 12-6 through 12-8.

12-21. MIRROR SYMMETRY. The term mirror symmetry refers to the fact that if the positive part of the resultant wave is inverted over the horizontal reference line, it will exactly match the negative part of the resultant waveform; or, conversely, if the negative part of the resultant wave is inverted over the horizontal reference line, it will have the same shape, outline, and appearance of the positive part of the resultant waveform. By viewing the resultant wave-

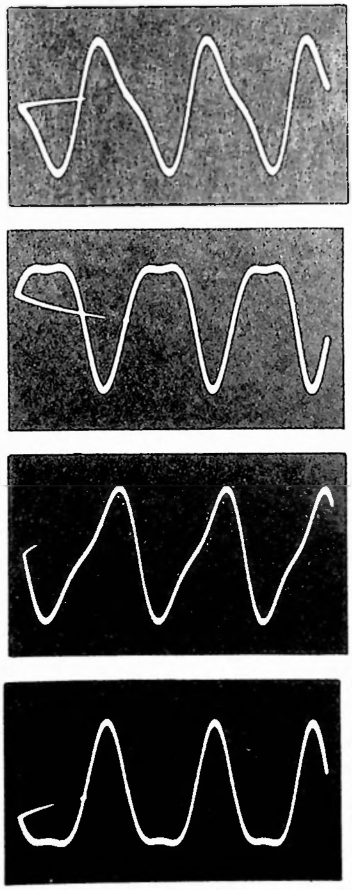


Figure 12-6. Resultant Waveforms Created by Algebraic Addition of Second Harmonic to Fundamental Sine Wave When Second Harmonic Amplitude is 30 Percent of Fundamental

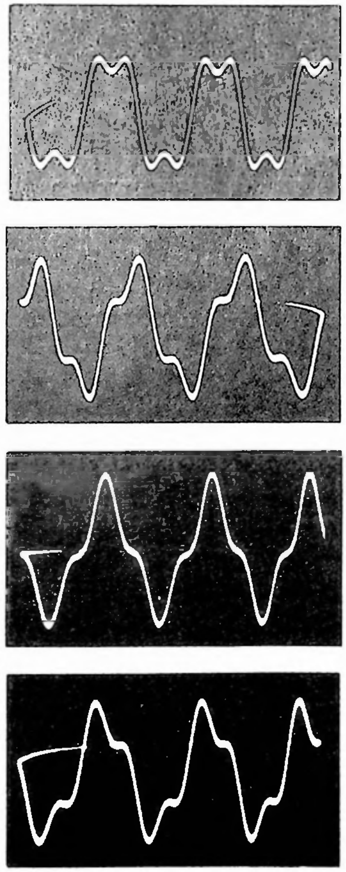


Figure 12-7. Resultant Waveforms Created by Algebraic Addition of Third Harmonic to Fundamental Sine Wave When Third Harmonic Amplitude is 30 Percent of Fundamental

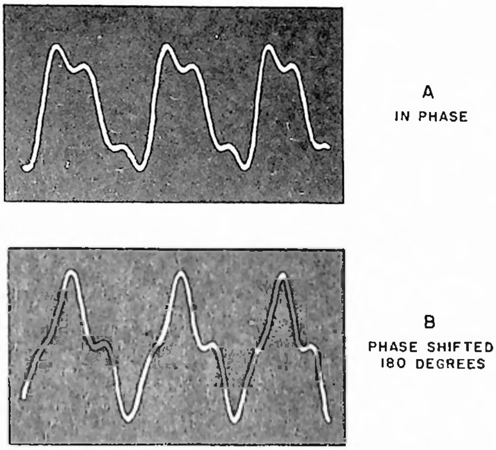


Figure 12-8. Resultant Waveforms Created by Algebraic Addition of Third Harmonic to Fundamental Sine Wave When Third Harmonic Amplitude is 15 Percent of Fundamental

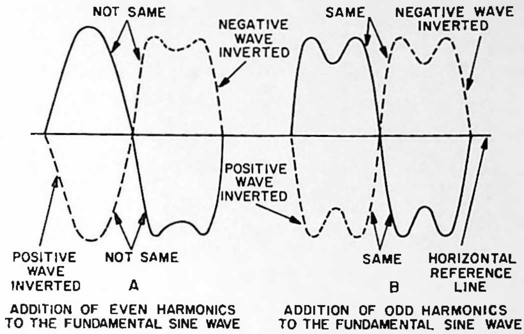


Figure 12-9. Presence or Absence of Mirror Symmetry Due to Harmonic Addition to Fundamental Sine Wave

form, you can definitely determine whether even harmonics (that is, 2nd, 4th, 6th, etc) were added to the fundamental sine wave to create the resultant, or whether odd harmonics were used for this purpose. If even harmonics (that is, 2nd, 4th, 6th, etc) were algebraically combined with the fundamental sine wave, there will be a lack of mirror symmetry, as illustrated in part A of figure 12-9. If odd harmonics (that is, 3rd, 5th, 7th, 9th, etc) were algebraically combined with the fundamental sine wave, there will be mirror symmetry, as illustrated in part B of figure 12-9.

12-22. SQUARE WAVEFORMS.

12-23. DESCRIPTION. The square waveform is a resultant waveform type composed of a sine waveform in conjunction with odd harmonics. Unlike the original sine waveform, the application of a square waveform to either a capacitive or inductive circuit will result in an output of a completely different waveform shape. As is illustrated in figure 12-10, the leading edge of a square wave rises from some zero reference value to its maximum value, where it remains as a constant-amplitude wave over a set period of time; it then drops back toward its original zero reference level, or beyond, until it reaches a minimum value, where it remains as a constant-amplitude wave over an exact period of time which matches its positive excursion. The rise and fall times are negligible in an ideal square wave. Part A of figure 12-10 illustrates the ac square waveform; it is so called because the waveform extends in a negative direction below the horizontal reference line as well as above. Parts B and C of figure 12-10 illustrate the pulsating dc square waveforms; they are so called because they contain a dc component which prevents the waveforms from crossing the horizontal zero dc reference level. However, all three forms are identical except for amplitude. All corners must be square, the sides perpendicular, and the extremities flat. Unfortunately, this idealized square waveform cannot be attained because the waveforming equipment is not perfect.

12-24. COMPOSITION. The square wave is formed by the algebraic addition of the

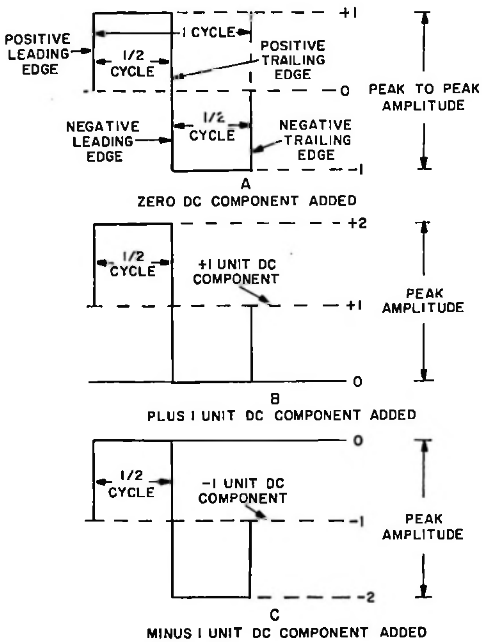


Figure 12-10. Square Waveforms

fundamental sine wave and an infinite number of odd harmonics of the fundamental sine wave. However, as is basically illustrated in figure 12-11, as few as three added odd harmonics will produce a reasonable facsimile of a square wave. However, a minimum of 10 added harmonics are required to produce a usable square wave. The fundamental sine wave may start at any phase; for illustrative purposes figure 12-11 shows the sine wave as beginning at the horizontal and vertical zero reference level. Figure 12-11 shows the resultant waveforms as you progressively add one additional harmonic to the fundamental sine wave. This illustration shows the algebraic addition of only three odd harmonics; however, it conveys the true impression that as each additional harmonic is added, the leading and trailing edges of the resultant waveform become steeper while the top and bottom become flatter. The frequency (f) of any order odd harmonic can be determined by calculating the value with the aid of the formula:

$$f_N = (2N + 1)\,f_1$$

where f_1 is the fundamental frequency and N is the order of the harmonic. For example, if your fundamental frequency is 100 cps and you wish to know the frequency of the 6th odd harmonic, you would proceed as follows:

$$f_6 = \left[(2 \times 6) + 1\right] \times 100$$

$$f_6 = \left[(12) + 1\right] \times 100$$

$$f_6 = (13) \times 100$$

$$f_6 = 1300 \text{ cps}$$

It also follows that each odd harmonic shown in figure 12-11 has been added in phase (zero phase difference) with the original fundamental sine wave. In addition, another factor of figure 12-11 should be inspected. The amplitude of each harmonic is in direct proportion to the harmonic order; that is, the third harmonic contains 1/3 the amplitude of the fundamental, the fifth harmonic contains 1/5 the amplitude of the fundamental, etc. If the harmonic is not in phase, or has incorrect amplitude, etc, the resultant square wave is said to be distorted. However, the type of distortion observed may indicate the kind of trouble, and even the source of trouble, within a circuit.

12-25. RECTANGULAR WAVEFORMS.

12-26. DESCRIPTION. The rectangular waveform contains all but one feature of the square waveform discussed in paragraphs 12-23 through 12-24 and illustrated in figures 12-10 and 12-11. This one different feature is that the former wave has identical periods of positive and negative pulsations, whereas the latter wave has unlike periods (time duration) of the positive pulse with respect to the negative pulse. The rectangu-

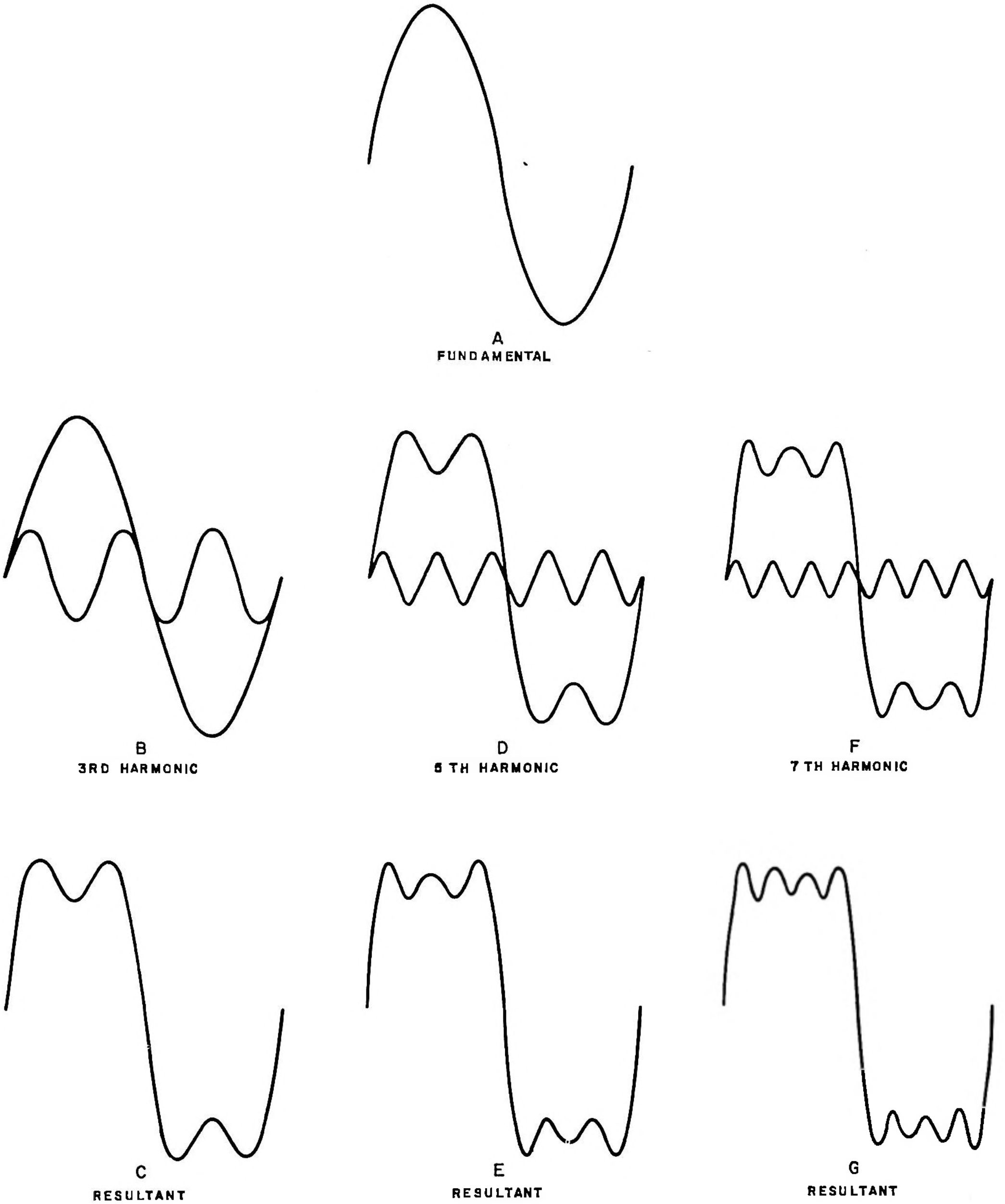


Figure 12-11. Formation of a Square Wave

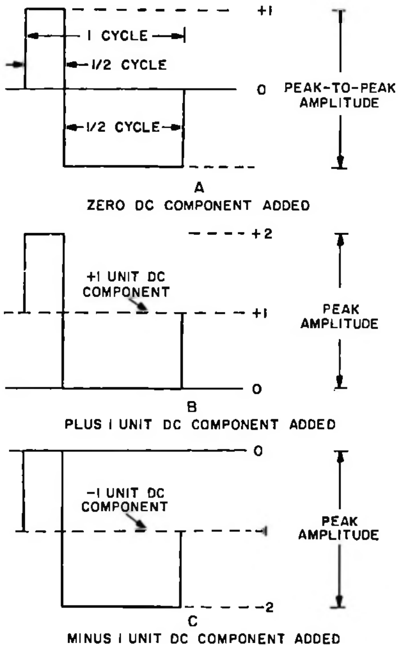


Figure 12-12. Rectangular Waveforms

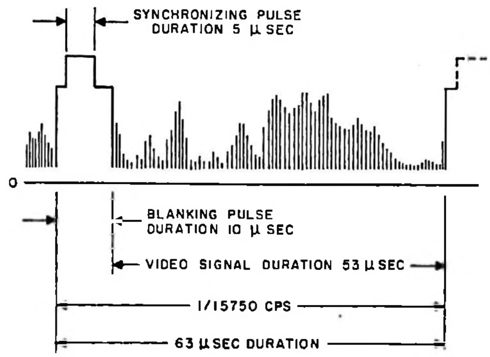


Figure 12-13. Rectangular Waves Used in Television

lar pulse can be bidirectional or it can be unidirectional, in that the waveform may be entirely above or entirely below the horizontal zero reference level, as shown in figure 12-12.

12-27. The period of the rectangular waveform, like that of the square waveform, is the total time required to complete both half-cycles together as one unit. For example, in part A of figure 12-12, if the positive half-cycle duration is 50 microseconds and the negative half-cycle duration is 150 microseconds, the total period for one cycle of this frequency is 200 microseconds. Considering that the frequency of a cycle is the reciprocal of the time required for that cycle, you will find that the frequency of this example waveform is 5000 cps. This means that this particular cycle will repeat itself 5000 times every second; it is said to have a frequency repetition rate of 5000 cps. The shorter pulse durations require the presence of higher-frequency components, whereas the longer pulse durations require the presence of lower-frequency components.

12-28. RECTANGULAR WAVEFORM APPLICATION. The rectangular waveform is rarely used as a test voltage. However, it may be used in many special applications to perform a specific function. Figure 12-13 illustrates such a function. The rectangular wave in this case has a rectangular wave riding atop the first wave. This is a practical situation in the transmission and reception of a television signal cycle. The video information is shown riding on the minimum amplitude portion of the first rectangular wave between the blanking pulses, representing the positive excursions of this rectangular pulse. However, rectangular synchronization pulses are shown riding atop the blanking pulse. This rectangular pulse represents the so-called front porch, the synchronizing pulse proper, and the so-called back porch.

12-29. SAWTOOTH WAVEFORMS.

12-30. DESCRIPTION. The sawtooth waveform, like all other waveforms except the fundamental sine wave, is composed of

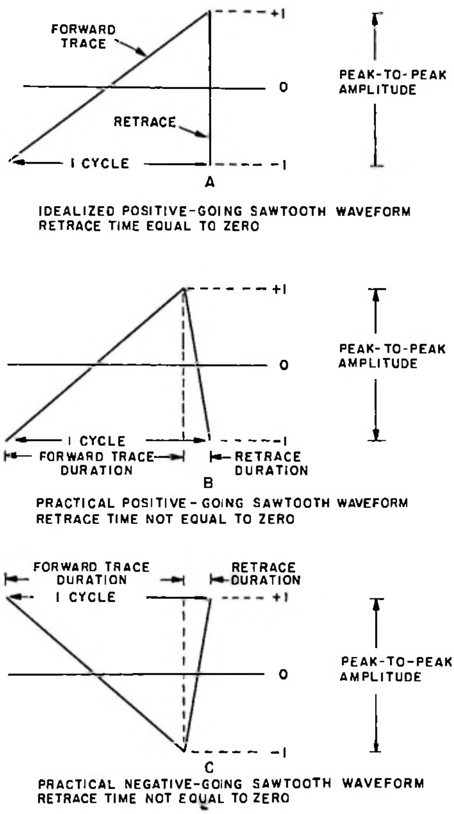


Figure 12-14. Sawtooth Waveforms

sine-wave components. As illustrated in figure 12-14, the wave consists of a gradual linear change from a maximum negative-going peak to its maximum positive-going peak, and then a rapid drop to its original amplitude. Considering that this waveform is composed of many sine-wave components which may differ in both frequency and phase, you cannot apply the sawtooth waveform to any inductive or capacitive device to cause different lead or lag times between the sine wave components that compose the sawtooth waveform. The output from any component device, other than a pure resistance, would not be the same as the original sawtooth input.

12-31. In the ideal waveform illustrated in part A of figure 12-14, the retrace time is shown as zero seconds. This is not the true practical case because you know that any action or reaction requires a definite time for accomplishment. Parts B and C of figure 12-14 reflect the practical case where the retrace time is some finite time, rather than zero. However, the retrace time is normally assigned the smallest practical duration consistent with the design of the equipment with which it is to be used. If the voltage amplitude increases at a constant rate during the forward trace, the waveform is said to be a linear sawtooth. The fact that half of the waveform, shown in figure 12-14 is above the horizontal zero reference level and the other half is below this reference level will normally not be seen on an oscilloscope, because the reference level line (time base line) is absent from the display.

12-32. COMPOSITION. Unlike the square waveforms which were produced by the algebraic addition of odd frequency, in-phase harmonic waves with the fundamental sine wave, the sawtooth waveform is the resultant wave produced by the algebraic addition of both even and odd frequency harmonics to the fundamental sine waveform.

12-33. A positive-going sawtooth waveform is produced by the algebraic addition of all harmonics to the fundamental sine wave, but the fundamental harmonic components must begin in-phase and start in a negative direction, as shown in figure 12-15. A negative-going sawtooth is the resultant of the same sine-wave components, but the fundamental and the in-phase harmonics must start in a positive direction. Figure 12-15 illustrates the method of progressive algebraic addition of each higher-frequency harmonic to the fundamental sine wave in order to gradually obtain the ultimate sawtooth waveform. However, only the first

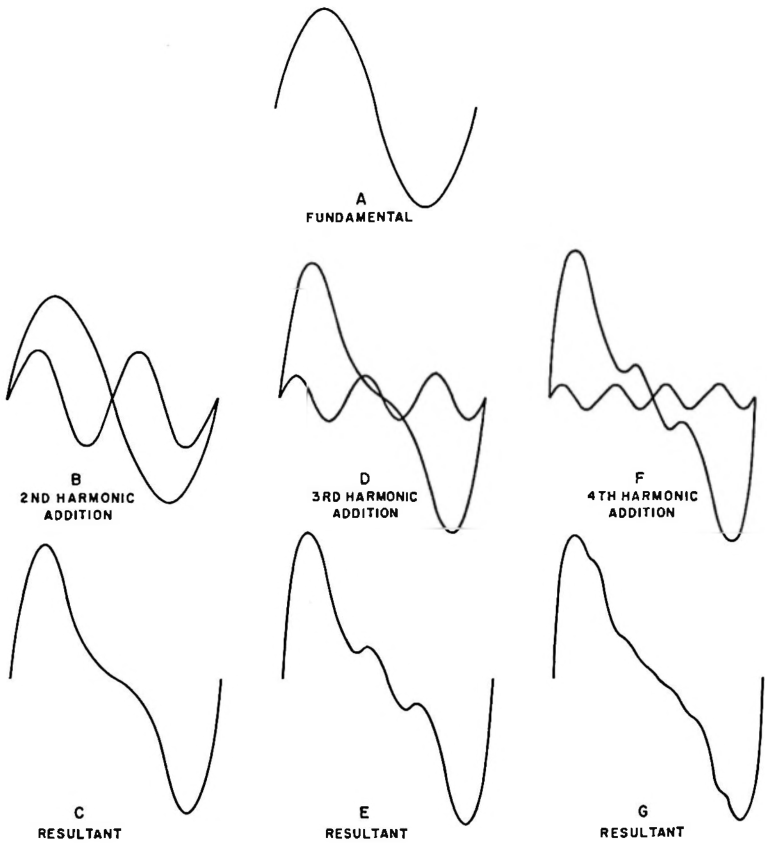


Figure 12-15. Formation of Sawtooth Waveform

two harmonics (2nd and 3rd) have been combined with the fundamental sine wave to form the resultant shown in part D of figure 12-15. As more harmonics are progressively added, the resultant wave will approach more and more closely the required sawtooth form.

12-34. TRAPEZOIDAL WAVEFORMS.

12-35. DESCRIPTION. The trapezoidal waveform is the resultant of the algebraic addition of sine waves, but it is more easily understood if explained in terms of a sawtooth rectangular waveform, both of which are composed of basic sine waves. As previously stated, a sawtooth of voltage applied to the input of either an inductive or capacitive device will not appear at the output of the reactive component or device as a sawtooth waveform. Therefore, those applications which require a sawtooth current waveform must obtain the sawtooth current from a device which has a sawtooth current output produced from a trapezoidal voltage input. The trapezoidal waveform has the necessary characteristics to cause a linear change in the amplitude of the current, with respect to time, as it passes through the resistive and inductive components of a coil. Figure 12-16 illustrates the resultant output of a resistor versus a coil with an applied input sawtooth waveform. The sawtooth wave passing through the pure resistive element produces no change in the output waveform. However, the output from the coil with an applied sawtooth waveform is essentially a

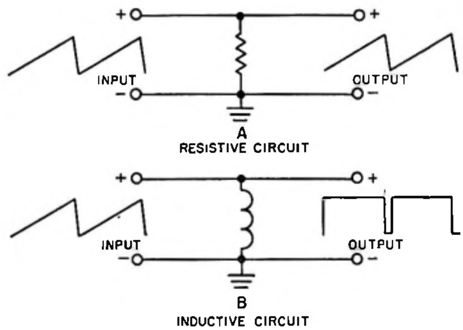


Figure 12-16. Output Current Waveforms for Resistive and Inductive Circuits Resulting from a Sawtooth Voltage Input

rectangular waveform. Figure 12-17 illustrates the algebraic addition of a sawtooth waveform to a rectangular waveform in order to produce a resultant trapezoidal waveform for application to a series resistive-inductive circuit, the output of which is a sawtooth current waveform.

12-36. FEATURES. The trapezoidal waveform occurs in numerous varieties because of the amplitude differences of the sawtooth voltages and rectangular voltages prior to algebraic addition. A comparison of two varieties of trapezoidal waveforms is illustrated in figure 12-18. The resultant waveform in part C is not the same as the resultant waveform in part F. Part C is the resultant trapezoidal waveform most commonly used in electronic technology. Sawtooth current waveforms are generated by deflection circuits for cathode-ray-tube deflection coils. When a deflection coil has

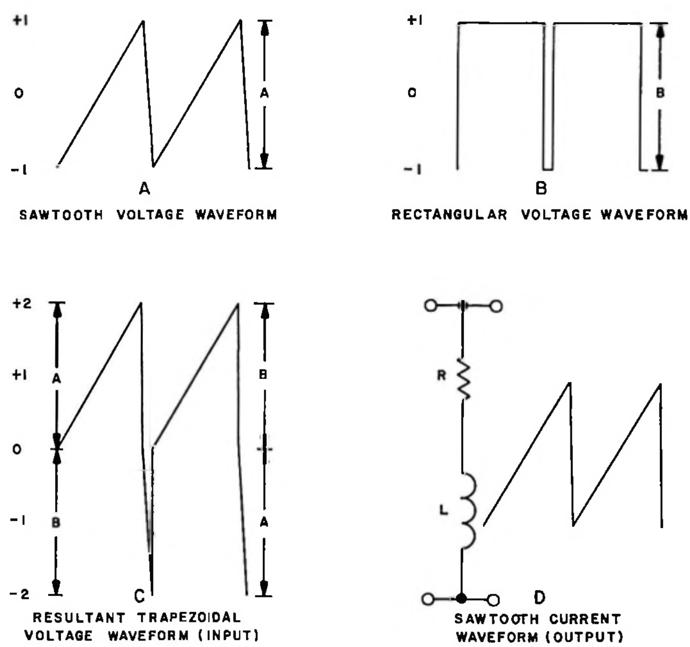


Figure 12-17. Output Current Sawtooth Waveform Resulting from Application of a Trapezoidal Input Voltage Waveform to an Inductor

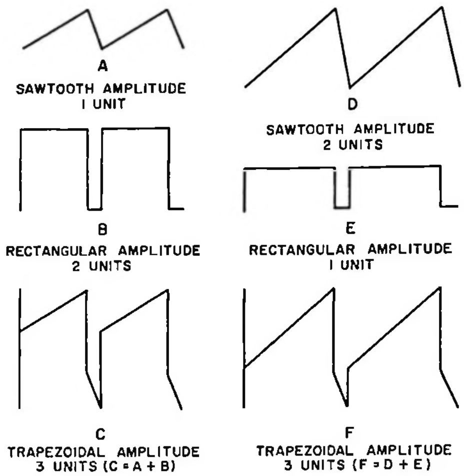


Figure 12-18. Trapezoidal Voltage Waveform Varieties

a small internal resistance as compared with its inductive reactance, the sawtooth current waveform is produced by a sweep voltage that is a combination of a small sawtooth waveform and a large rectangular waveform.

12-37. DIFFERENTIATED VOLTAGE WAVEFORMS.

12-38. DESCRIPTION. Various complex waves can be resolved into their component sine-wave frequencies, and any group of frequencies can be extracted from a complex wave by means of a filter. In the case of differentiation, the differentiated waveform extracts the high-frequency sine-wave components, while the integrator extracts the low-frequency sine-wave components. A differentiated waveform is obtained by the process of differentiation. This process is simply the procedure whereby a waveform is passed through inductive or capacitive components to provide a voltage output which is proportional to the rate of change of the input voltage waveform. The most popular method of differentiation is the use of a capacitive-resistance (RC) network. The time constant of the circuit, in microseconds, is the product of the resistance and capacitance in ohms and farads, respectively. A rapid change occurring in the input voltage waveform will produce a narrow, sharp-peak pulse (spike) in the output. The peak amplitude of the output pulse is therefore directly proportional to the rate of change in the input waveform. Figure 12-19 illustrates a square-wave input to each of the two most common differentiating circuits, and also the output obtained from each circuit. A square wave was used because of its rapid amplitude change and high harmonic frequency content on both the leading and trailing edges of the applied pulse. The flat horizontal portions of the square wave will produce zero output because they contain zero slope (change), and

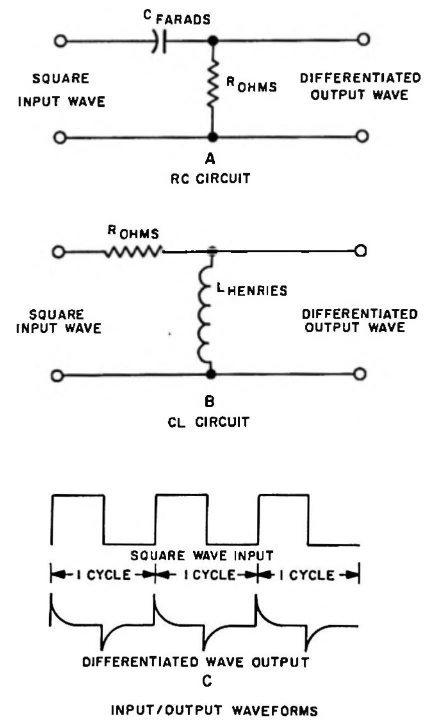


Figure 12-19. Input to and Output from Differentiating Circuits

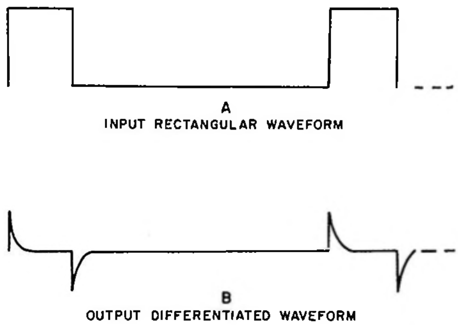


Figure 12-20. Rectangular Input and Resultant Differentiated Output Waveform

also because the time constant of the differentiator will not pass the lower frequencies contained in the square wave. A sine wave is not used as an input voltage to a differentiation circuit because these circuits accomplish their function by shifting the phase of the input waveform. In the case of a sine wave, the output will be shifted in phase and will have a smaller amplitude, but it will still be a sine wave. This output may be particularly useful as a phase-shifted wave formed from a continuously variable time constant function.

12-39. RECTANGULAR VOLTAGE WAVEFORMS. The rectangular waveform has the same characteristics as the square waveform with respect to differentiation. It is important to note that the output voltage has a sharp peak only when the differentiating circuit contains a short time constant. Also, you should realize that the sharp peak is produced only during the rapid rise or rapid fall of the input voltage. Therefore, the differentiator circuit is known as a "peaker" circuit. In a circuit containing a time constant of less than 1/10 the time required for one cycle of input voltage, the time constant is said to be short; it is said to be long if the circuit components permit a time of 10 times the duration of one cycle of input voltage. A square waveform produces an output differentiated wave with evenly spaced positive and negative excursions, whereas a rectangular waveform produces a positive and negative peak spaced close together (paired), with a distance separation from the next pair of peaks, as shown in figure 12-20.

12-40. SAWTOOTH VOLTAGES. When you apply a sawtooth voltage to a differentiating circuit containing a short time constant, the output from that circuit will be a rectangular waveform. If the applied sawtooth is positive-going, the negative spike of the output rectangular waveform will increase in amplitude as the retrace duration time is made smaller. This condition is illustrated in figure 12-21. However, as the time constant of the differentiated circuit is increased, the output progressively takes on the appearance of the input sawtooth waveform, as shown in figure 12-22.

12-41. RESISTOR-INDUCTOR DIFFERENTIATION. The RL differentiator, consisting of a resistor and an inductor in series, is used for the same purpose as the RC differentiator. The output of the RC circuit is taken from across the resistor, whereas the output of the RL circuit is taken from across the inductive element. However, using either form of differentiator, the time constant of the circuit represents the actual time required for the voltage to charge the capacitor in the RC network, or for the current to charge the coil in the RL network. The actual time constant of a differentiator in microseconds can be obtained by use of the applicable formula:

$$T = RC$$

or

$$T = L/R$$

where R is in ohms, C is in microfarads, and L is in microhenries. The shape of the voltage waveform across the capacitor and

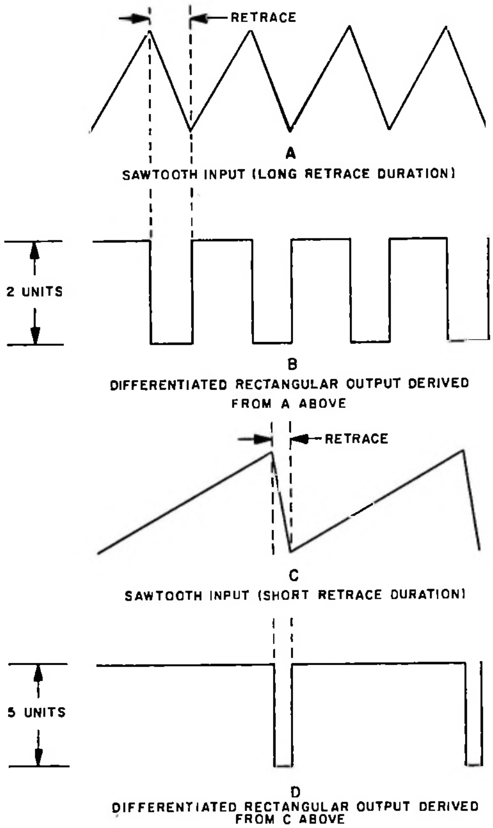


Figure 12-21. Differentiated Wave Amplitude Changes Resulting from Sawtooth Input Rate-of-Change

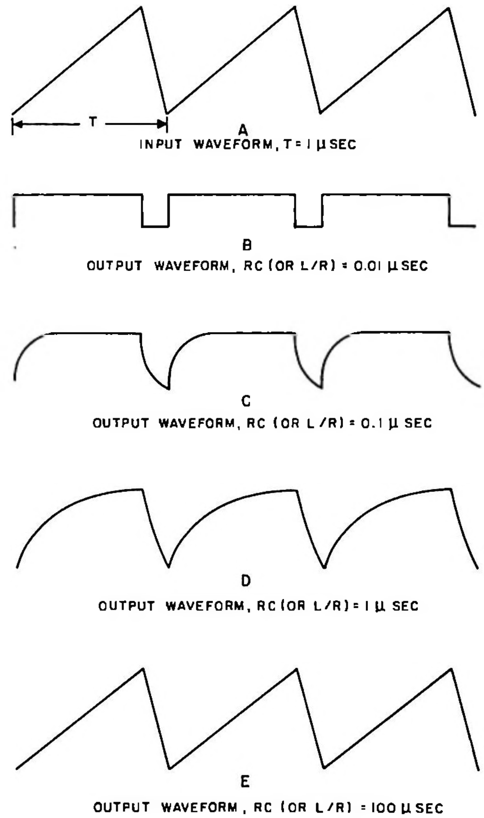


Figure 12-22. Differentiated Output Waveforms for Sawtooth Input Waveform Progressively Illustrating an Increasing RC or RL Circuit Time Constant

the waveform of the current through the coil are both identical. Therefore, technical data pertaining to the output voltage waveform of the RC network is the same for the output current waveform of the RL network.

12-42. INTEGRATED VOLTAGE WAVEFORMS.

12-43. DESCRIPTION. In contrast to the differentiator circuit, which is actually a high-pass filter, the integrator sums up the applied voltages and discriminates against high frequencies; it is thus a low-pass filter. The integrator can use exactly the same components as the differentiator circuit. However, in the case of an RC integrator circuit the output is taken from across the capacitor, whereas in the differentiator circuit the output was taken from across the resistor. The reverse is also true of the RL integrator circuit. The output is taken from across the resistor, whereas in the differentiator circuit the output was taken from across the coil. Figure 12-23 illustrates the two common forms of integrator circuits.

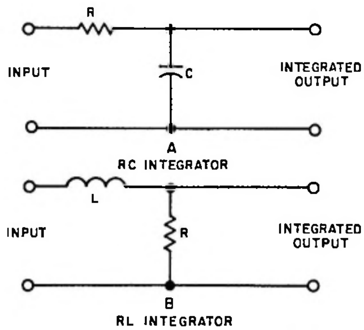


Figure 12-23. Typical Integrator Circuits

12-44. When a square wave is applied to an integrator circuit and the integrator circuit time constant is increased, the output waveform will gradually take on the appearance of a sawtooth waveform and will decrease in amplitude. However, the shorter the circuit time constant, the more closely the shape of the output will resemble the shape of the input. Figure 12-24 shows the various forms and representative amplitudes for an input square or rectangular waveform.

12-45. When a square waveform is applied to the input of an integrating network having a long time constant, the output waveform will approximate a sawtooth wave with the charge or trace portion equal to the discharge or retrace portion. However, if a rectangular waveform is used as the input to the same integrating circuit, the output will not have equal charge and discharge times; therefore, the resultant waveform will build up in either a positive or negative direction. A positive build-up of the output waveform as a result of an input rectangular waveform with longer positive pulse durations than negative pulse durations is illustrated in figure 12-25. As you can see, the shorter time duration provided by the negative portion of the input rectangular pulse will provide less time for the output discharge, and as a consequence, the output waveform will charge to its maximum value more quickly.

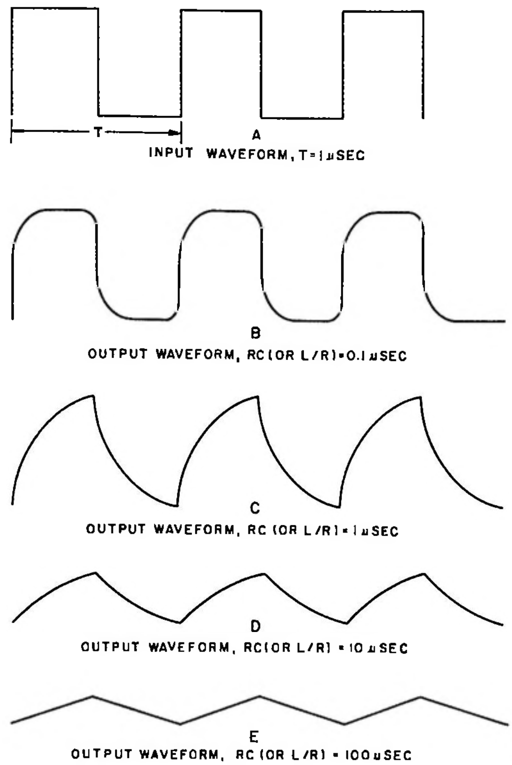


Figure 12-24. Integrated Output Waveforms Progressively Illustrating an Increasing RC or RL Circuit Time Constant

12-46. MODULATED WAVEFORMS.

12-47. GENERAL. The art of superimposing on, combining with, or changing the original carrier frequency, by the addition of intelligence in the form of electrical energy, is termed modulation. The three primary types of modulation are amplitude modulation, frequency modulation, and phase modulation.

12-48. AMPLITUDE MODULATION. The radio-frequency (rf) carrier is normally generated with the characteristics of a constant frequency and a constant amplitude. However, you can vary the amplitude of

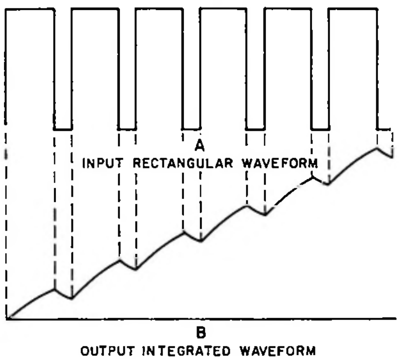


Figure 12-25. Cumulative Wide Integrated Pulse Obtained from Narrow Pulse Rectangular Waveform

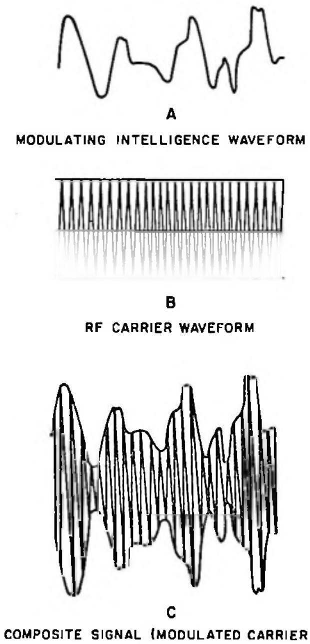


Figure 12-26. RF Carrier Amplitude Modulated by a Complex Waveform

this carrier in direct accordance with the intelligence to be transferred (that is, the spoken word, music, etc), by simply adding the amplitude of the intelligence algebraically to the amplitude of the rf carrier. This is accomplished by means of some form of mixing circuit. Figure 12-26 illustrates a hypothetical waveform representing the intelligence, a hypothetical rf carrier, and the resulting composite signal obtained by algebraic addition of the carrier and intelligence frequencies. This resultant amplitude-modulated waveform is transmitted and received, and the amplitude is then detected (separated from the carrier) and converted back to a facsimile of the original intelligence information.

12-49. The amplitude-modulated carrier illustrated in part C of figure 12-26 is actually composed of one carrier frequency component in addition to two other frequency components (sidebands) for each individual frequency component originally contained in the modulating voltage waveform. For example, if the carrier frequency shown in part B of the figure were 1000 kilocycles and the modulation voltage shown in part A were 10 kilocycles, the resultant modulated rf carrier shown in part C would contain the original carrier frequency component of 1000 kilocycles plus a lower sideband component obtained by algebraically subtracting the modulating frequency from the carrier frequency (1000 kc − 10 kc = 990 kc) and an upper sideband component obtained by algebraically adding the modulating frequency to the carrier frequency (1000 kc + 10 kc = 1010 kc). If you modulated the same carrier with two modulating frequencies, such as 10 kc and 20 kc, the resultant modulated carrier would be composed of five frequency components (that is, 1000 kc, 990 kc, 1010 kc, 980 kc, and 1020 kc). As you can see, the original frequencies of the modulating voltages are not apparent, and the intelligence is now contained within the parameters of the sidebands created. Do not assume that the original carrier now contains intelligence, because the carrier will be completely eliminated after it has served its purpose.

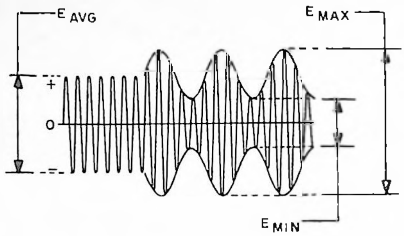


Figure 12-27. RF Amplitude Percentage Modulation

12-50. In calculating the percentage of amplitude modulation, the difference between the maximum and minimum carrier amplitudes is compared with the average carrier amplitude. Figure 12-27 and the formula below illustrate this principle.

$$\text{Percentage modulation} = \frac{E_{max} - E_{min}}{2E_{avg}} \times 100$$

The modulating voltage can increase the amplitude of the carrier any amount above and below the horizontal zero reference level without creating any difficulty. However, if it decreases the amplitude of the carrier to the zero reference level, it will remove the existing carrier frequency and thus create distortion. This type of distortion is termed overmodulation, and is illustrated in figure 12-28. Superimposed modulation is normally undesirable. This type of modulation is readily recognized because the negative portion of the modulated carrier is not inverted. Superimposed modulation is normally encountered as a result of hum or noise modulation. Figure 12-29 provides an example of superimposed modulation without the aid of actual modulator equipment. In addition, no sidebands are evident. Thus, only the modulation frequency and the carrier frequency are present. The superimposed distortion signal illustrated in figure 12-29 has limited application in that this type of signal is sometimes applied to the input of an amplifier, and if the output from

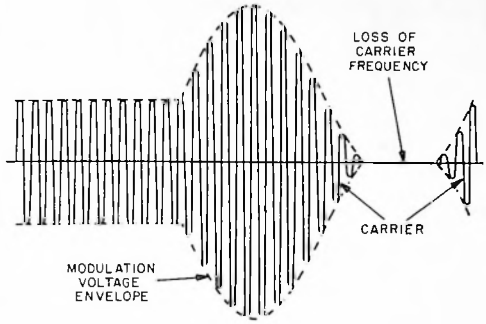


Figure 12-28. Overmodulated RF Carrier

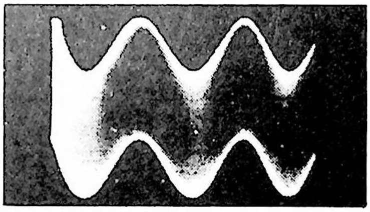


Figure 12-29. Superimposed Modulation

the amplifier is reasonably close to being identical with the input, you can say that the amplifier stage is distortionless and therefore linear. However, if the amplifier stage is not distortionless, or linear (amplifies all applied frequencies within its range equally), the output will be amplitude-modulated by the amplifier stage, and the output will therefore contain the upper and lower sideband-frequency harmonic components not contained within the original waveform. These frequency components will then modulate each other. This effect is termed intermodulation distortion, and is illustrated in figure 12-30.

12-51. FREQUENCY MODULATION. This method of modulating the constant-frequency, constant-amplitude carrier will also permit the transference of intelligence. However, considering that the majority of man-made

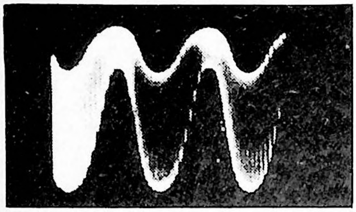


Figure 12-30. Intermodulation Distortion

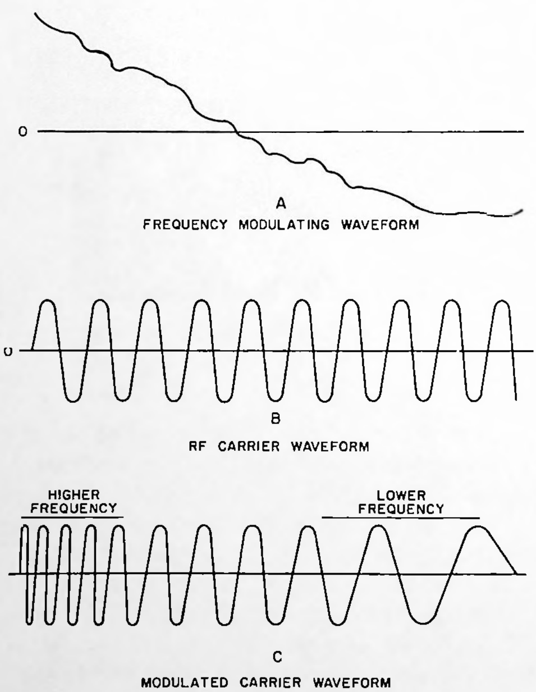


Figure 12-31. RF Carrier Frequency Modulation by a Complex Waveform

or natural noise interference is amplitude modulation, the frequency modulation method is relatively free from noise or other interference during the process of intelligence transference. In the frequency modulation technique, the modulation signal represents the intelligence by frequency changes rather than by amplitude changes. Therefore, when the modulation signal is used to modulate the constant-frequency carrier signal, no amplitude change occurs. The resultant rf modulated carrier signal will contain no amplitude variation, but its frequency will vary in direct accordance with the modulation intelligence (that is, spoken word, music, etc). The resultant composite signal is illustrated in figure 12-31.

12-52. PHASE MODULATION. The art of phase-modulating a constant-amplitude, constant-frequency carrier will result in basically the same type of transference characteristics. To reiterate, amplitude modulation involves changing the carrier amplitude, and frequency modulation involves changing the carrier frequency; logically, then, phase modulation involves changing the carrier phase in direct accordance with the intelligence. The primary difference between frequency modulation and phase modulation is that in the former the modulation index is inversely proportional to the modulation frequency, and in the latter the modulation index is completely independent of the modulation frequency. Therefore, with a fixed modulating voltage, the frequency band required to accommodate a phase-modulated signal is directly proportional to the modulating frequency, whereas with frequency modulation the band occupied is completely independent of the modulating frequency unless the modulation index is less than 0.5. Figure 12-32 illustrates the progression of phase-modulating a carrier frequency to produce a resultant waveform capable of transference.

12-53. RESPONSE AND DISCRIMINATOR WAVEFORMS.

12-54. RESPONSE CURVE. A response curve is a graph showing the relationship between output voltage and frequency. The

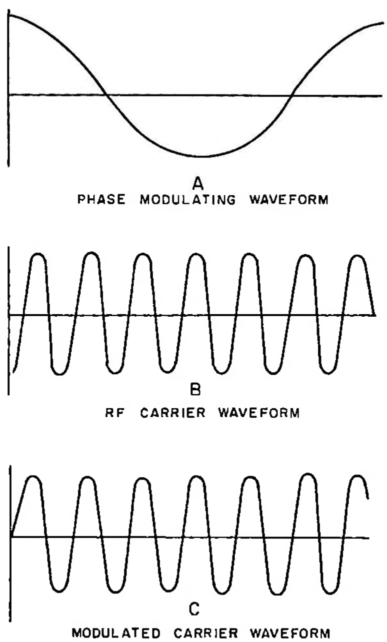


Figure 12-32. RF Carrier Phase Modulation by a Sinusoidal Waveform

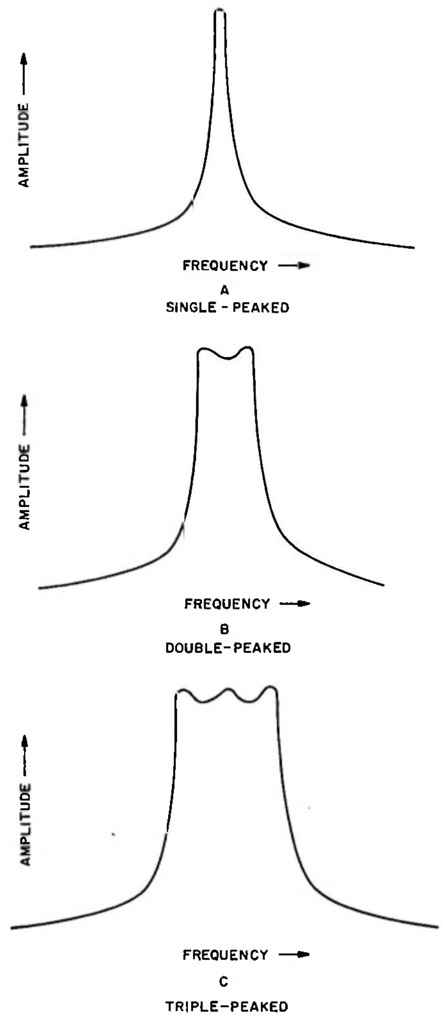


Figure 12-33. Primary Types of Response Curves

response curve can indicate the degree of acceptance, amplification, or rejection, by either a component or a circuit, as the signal frequency is varied over a desired range. There are three primary types of response curves: single-peaked, double-peaked, and triple-peaked, as shown in figure 12-33.

12-55. As you can see, frequency is plotted along the horizontal axis, while amplitude of the output current or voltage is plotted along the vertical axis. The circuit response for a given input frequency is the measured amplitude separation between that point on the response curve representing the frequency and the horizontal zero reference line. The amplitude of the response curve may be shown as either above or below the horizontal zero reference line, as illustrated in figure 12-34.

12-56. The half-power points shown in figure 12-34 are 3 db down from the peak or maximum amplitude point on the curve. The term 0.707 is an amplitude value above or below the horizontal zero reference level, and is obtained from the reciprocal of the square root of 2 ($1/\sqrt{2}$). A single-peaked response curve indicates that you are tuned to a single frequency, and you will naturally obtain a very narrow frequency

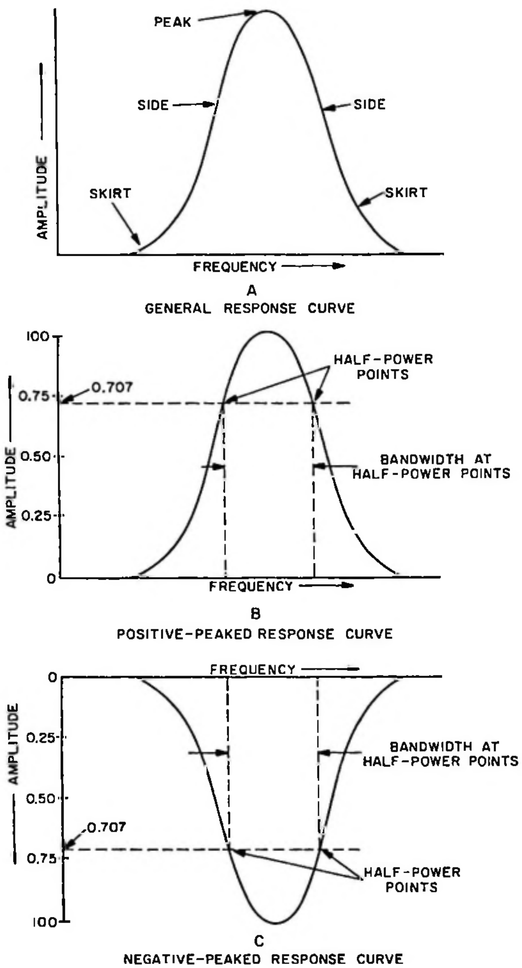


Figure 12-34. Positive and Negative Single-Peaked Response Curves

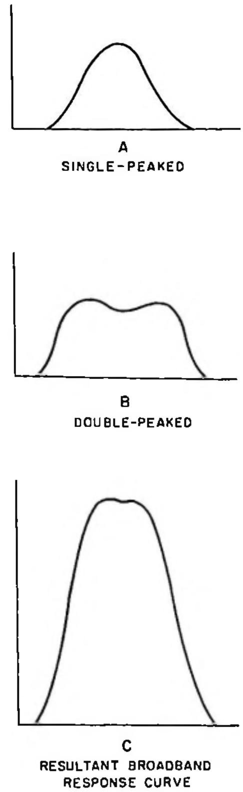


Figure 12-35. Response Curve Combinations to Produce a Required Resultant Wide-Band Response Curve

pass band. A double-peaked waveform is the result of the deliberate design of transformer-type circuits which, when tuned to a single frequency, will provide a voltage maximum peak above the resonant frequency and a voltage maximum peak below the resonant frequency. The resonant frequency will be represented by the dip between the two peaks, as illustrated by part B of figure 12-33. The purpose of this type of waveform is to increase the frequency pass band by increasing the amplitude of a greater number of frequencies adjacent to the center frequency. The greater the dip between the two peaks, the greater the coupling between the primary and secondary windings of the transformer. Too great a dip is undesirable. A flat-topped curve is the ideal because all frequencies within the pass band will be the same amplitude. Several response curves may be algebraically added through a mixing circuit to produce a flat-topped, broad resultant response curve, as illustrated in figure 12-35.

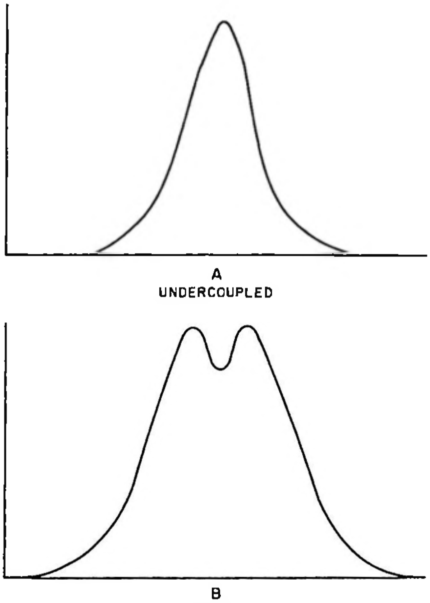


Figure 12-36. Response Curve Coupling

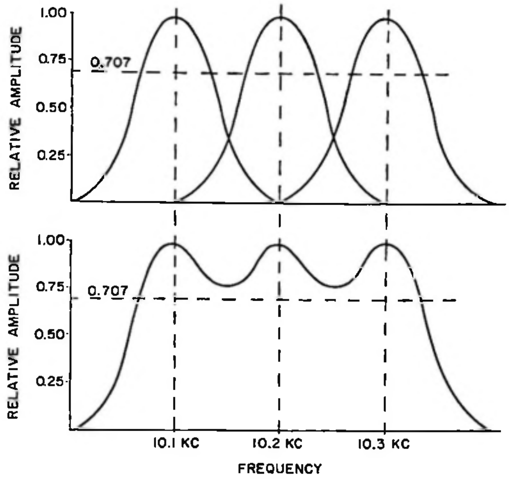


Figure 12-37. Response Curve Resulting from Stagger-Tuned Stages

12-57. The terms overcoupled (close coupling) and undercoupled (loose coupling) refer to the spacing between the primary and secondary windings of the transformer. For example, if the primary is brought closer to the secondary (overcoupled), all frequencies within the bandwidth will be transferred from the primary to the secondary with approximately the same amplitude; this provides a wider pass band, less frequency selectivity, and greater over-all amplitude. However, if the primary and secondary windings are moved farther apart, more impedance is effectively placed between the two windings, and only the frequencies containing the greatest amplitude will have sufficient energy to bridge the gap. This will create a sharply peaked waveform in the output, representing a very narrow bandwidth, high frequency selectivity, and less over-all amplitude, even though the waveform peak is more pronounced. These effects are illustrated in figure 12-36.

12-58. Broad-banding, or the art of increasing the bandwidth to permit a greater number of frequencies to pass, is accomplished by two primary methods: overcoupling, as was discussed in paragraph 12-57, and stagger tuning. The term stagger tuning refers to the tuning of a series of circuit stages to slightly different frequencies. For example, three stages could be tuned 1000 cycles apart from one another, as illustrated in figure 12-37, to combine as a resultant waveform. This resultant waveform is considered as a triple-peaked response curve.

12-59. DISCRIMINATOR CURVES. The output from a discriminator circuit is sometimes referred to as an "S" curve. Figure 12-38 illustrates the ideal form of the "S" curve used as a reference standard. Any deviation from this shape represents incorrect tuning of the primary or secondary transformer windings, or other improper

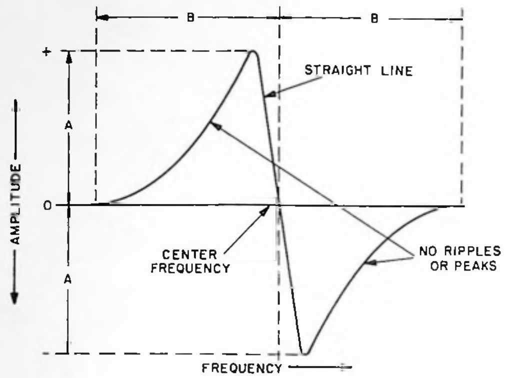


Figure 12-38. Discriminator "S" Curve

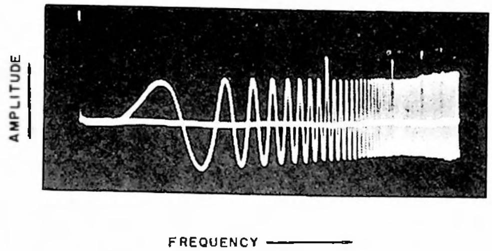


Figure 12-39. Ideal Audio-Frequency Response Curve

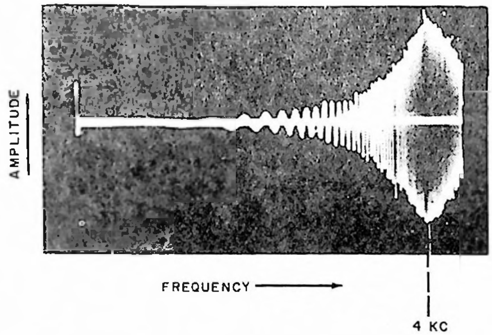


Figure 12-40. Resonant Circuit Audio-Frequency Response Curve

circuit adjustment. The "S" curve is linear, and it always crosses the horizontal zero reference axis at the point on the curve representing the center frequency. Many times a marker pulse is electronically added so that it appears at some point on the curve. However, this marker will disappear at the center frequency because this point occurs at zero voltage amplitude. The positive amplitude and low-frequency components on one side of the center frequency should equal the negative amplitude and high-frequency components, respectively, on the opposite side of the center frequency. In other words, A = A (amplitude) and B = B (frequency separation) in figure 12-38.

12-60. The audio-frequency response curve shown in figure 12-39 is ideal. The constant height of this response curve proves that the circuit under test has a flat response from its lowest to its highest frequency. The horizontal zero reference base line is useful for measuring relative amplitudes. Considering that the portion of the wave below the reference or base line is exactly the same as the wave above the base line, you need only observe the top half of any "S" curve for full information. Any peaks extending above the average amplitude of the waveform represent accentuation of the frequencies within that region of the pass band, and valleys or dips reflect attenuation of the frequencies at those points. Therefore, this waveform as an input not only shows you the circuit behavior as a whole, but instantly reflects any unusual frequency characteristic of any recently added components, filters, or circuits. For example, figure 12-40 represents an undesirable voltage-frequency characteristic within an LC filter circuit which is resonant at 4000 cps.

12-61. Video and other high-frequency response curves are similar to low-frequency (audio) response curves. However, in high-frequency curves the frequency bandpass is wider (broader), with an extremely low-frequency limit (60 cps) and an extremely high-frequency limit (in the mega-

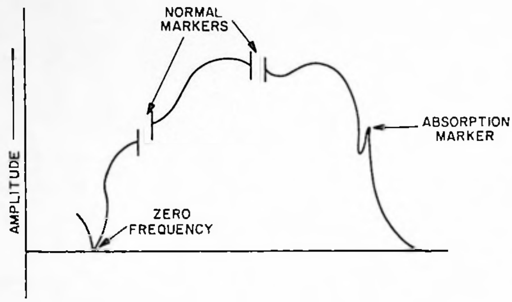


Figure 12-41. High-Frequency Response Curve

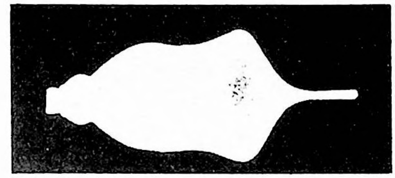


Figure 12-42. Non-demodulated High-Frequency Response Curve

cycle range). Two different types of markers may be used to designate exact frequencies; the first is a disturbance along the response curve at a particular frequency, whereas the second is produced by a tuned circuit which removes or absorbs energy from the response curve at a particular frequency. Both types of markers are illustrated on the typical high-frequency response curve shown in figure 12-41. Figure 12-42 illustrates another type of high-frequency response curve. However, the curve shown in figure 12-42 has not been demodulated and is not popular because frequency markers are very hard to discern on this type waveform.

SECTION II

FACTORS CONTRIBUTING TO WAVEFORM DISTORTION

12-62. GENERAL.

12-63. Distortion is normally considered as a deviation from the desired waveform. The undesirable waveform in one application may be the desired waveform in some other application. Therefore, the term distortion refers to a particular waveform application, and is meaningless if no application is being considered.

12-64. A normal high-frequency current is characterized by its amplitude, frequency, and phase relationships, and can be altered by changing any one of these characteristics. Actually, any two or possibly all three characteristics may be altered by a circuit change. If the circuit change produces the desired signal, this new signal is termed an undistorted, or pure, waveform; if the circuit change produces an undesired signal, it is termed a distorted waveform.

12-65. The factors contributing to waveform distortion in one application may be the same factors required to produce a desired waveform in some other application. This section considers and explains only those undesirable factors which contribute to distortion.

12-66. CIRCUIT.

12-67. GENERAL.

12-68. The primary cause of distortion created within an actual circuit can be traced to overloading of the tubes in that circuit. For example, overloading an amplifier tube would cause the tube to operate over a nonlinear portion of its characteristic curve, and the output waveform would not retain the same shape as the input. The same overloading effect can occur if the tube is defective or if one or more of the applied operating potentials is incorrect. For other than tube or circuit defects and incorrect-operating potentials, you can eliminate distortion by simply decreasing the input amplitude (volume control) or intensity. The placement of components, wires, or leads may create undesirable feedback voltages in a phase relationship which results in distortion. Distortion-removing circuits, designed to eliminate feedback, may be defective. Neutralization circuits may be used to remove or balance out distortion resulting from undesirable feedback. The elimination or overemphasis of the amplitude of particular frequencies, within a desired band or range of input signal frequencies, will create distortion. The primary and secondary windings of frequency-sensitive transformers may be incorrectly tuned or spaced an incorrect distance from each other. Therefore, the sideband frequencies, which form an important part of the resultant desirable signal, may be missing from the output signal. Finally, defective input or output components may blank out certain pass-band frequencies or permit undesirable voltages to pass, thereby causing distortion.

12-69. AMPLITUDE DISTORTION.

12-70. Amplitude distortion may be caused by a limitation of bandwidth or by irregularities within the bandwidth. In either event, amplitude distortion is normally expressed in terms of attenuation because it is a logarithmic quantity which is algebraically added for cumulative stages. Amplitude distortion, free of phase distortion, cannot change the symmetry of a symmetrical input pulse. The response of a circuit should be the same for all frequencies present in the input signal voltage, but, if the circuit response is not the same for all input frequencies, suppression or exaggeration of the amplitude of some frequencies will create distortion. The fundamental plus harmonics will be seen or heard in the output waveform when amplitude distortion exists.

12-71. In the case of an amplifier stage, you can see whether amplitude distortion is present by applying a signal voltage of known characteristics to the amplifier input and then viewing or measuring the output signal. The output waveform should be a replica of the input waveform. Amplitude distortion caused by an amplifier is the result of the generation, by the amplifier, of frequencies which were not contained in the input. The result of generating additional frequency components is seen by the change in waveform amplitude. A part of the output voltage may be returned to the input 180 degrees out of phase to cancel the tube distortion. This process is called inverse feedback.

12-72. POWER AMPLIFIERS (DOUBLE-ENDED). In class A1 power amplifiers, the amount of amplitude (nonlinear) distortion will depend on the grid voltage swing, and this, in turn, depends on the plate load impedance, grid bias, and tube operating point. The grid bias voltage is adjusted so that a grid voltage swing, which is equal to the bias voltage, will produce an output with the maximum permissible distortion. Then, when the grid signal voltage is equal to the bias voltage, the output signal has the maximum power level that can be obtained from the amplifier without excessive distortion. The value of grid swing developed by this method to produce full power output with minimum distortion is termed full grid swing. The grid current must, at no time, be permitted to flow because of the resulting distorted output.

12-73. POWER AMPLIFIERS (SINGLE-ENDED). The output harmonic content of a single-ended triode class A amplifier is composed mostly of second harmonic content. The second harmonic distortion increases if the operating point is lowered while holding a constant load resistance. The second harmonic amplitude distortion decreases if the plate load impedance is increased while holding a fixed operating point. However, with a given value of plate voltage, you can increase the plate load resistance and lower the operating point to provide an increased grid swing without exceeding the permissible distortion allowance.

12-74. VOLTAGE AMPLIFIERS. In resistance-capacitance-coupled voltage amplifier stages, it is not practicable to use a plate load resistance over 5 times the electrical size of the plate-to-cathode resistance (r_p). The plate supply voltage is normally limited to prevent tube damage. The distortion resulting from low plate voltages can be lowered in a pentode tube by lowering the screen voltage. This will lower the plate current and thus effectively raise the plate voltage. The screen voltage should normally be approximately one-third the plate supply voltage. The transient time required for the coupling circuit to charge or discharge may be so long at high frequencies that amplitude distortion will be produced. Transformer coupling between the final amplifier stage and the load is normally required because the load impedance may be

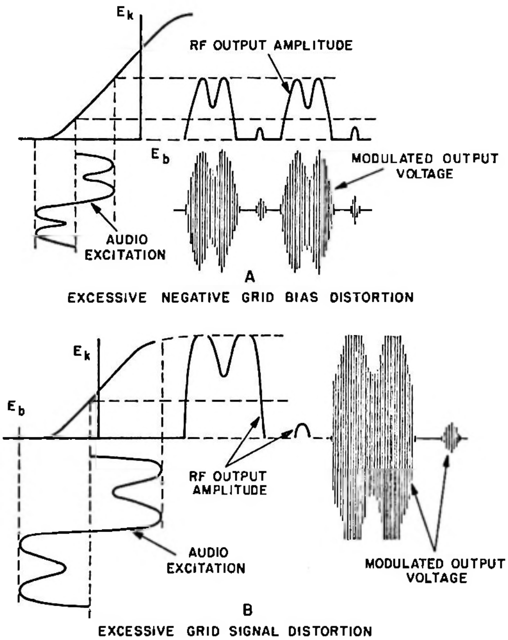


Figure 12-43. Modulation Distortion

so small that amplitude distortion would otherwise result.

12-75. RADIO-FREQUENCY AMPLIFIERS. Intermodulation between the different applied input frequencies will create amplitude distortion. This distortion becomes more apparent with higher-amplitude input signals, as shown in figure 12-43.

12-76. FREQUENCY DISTORTION.

12-77. Frequency distortion occurs when different frequency inputs are not all amplified equally. The distortion may be audible or inaudible, depending on the circuit frequency response limits. In addition, if the circuit output load is composed of reactive components, the low-frequency resonance and the increase in inductive resistance at high frequencies will increase the nonlinear (amplitude) distortion and modify the response. If a feedback network

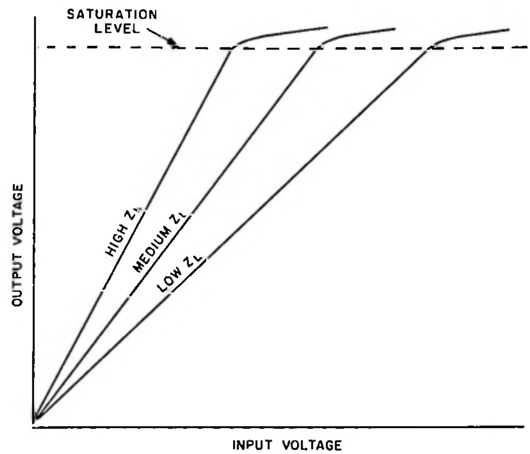


Figure 12-44. Saturation Distortion

contains reactive elements, then the overall gain of an associated stage is a function of frequency, and frequency distortion due to feedback will be obtained. However, negative feedback, even when reactive elements are present, will decrease the total circuit distortion at the expense of maximum gain.

12-78. The distortion in linear amplifiers as a result of the relationship between the input voltage and output voltage is a type of frequency distortion as well as a type of amplitude distortion. As the input voltage is increased, the plate current will increase up to a point where an input voltage increase will not produce a proportional plate current increase. At this point, the tube is said to be saturated. This is illustrated in figure 12-44.

12-79. With a square waveform applied to the input of a linear circuit, the output should also be a square waveform. However, if the circuit response is not the same for all frequencies, the output waveform will not be the same shape as the input waveform. Figure 12-45 shows output waveforms for nonlinear circuit response.

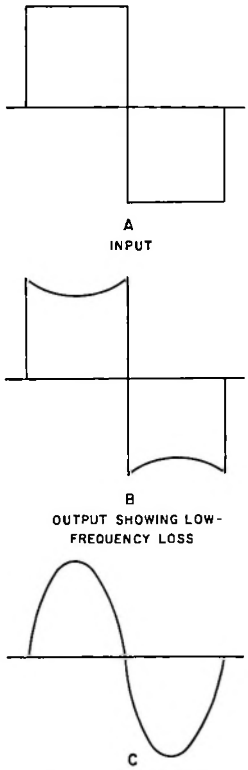


Figure 12-45. Output Waveforms Resulting from Poor Circuit Frequency Response

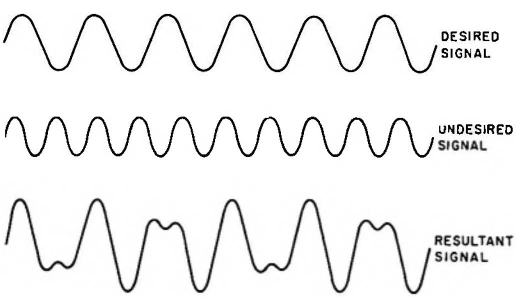


Figure 12-46. Combination of Two Signals Forming an Amplitude- and Phase-Modulated Resultant

12-80. Frequency modulation distortion is often termed flutter distortion. This type of frequency distortion is generally the result of speed fluctuations as a recording is driven by the recorder or reproducer motor. The flutter effect may also be caused by a loudspeaker when it is reproducing two frequencies. This is true because the sound pitch is a function of the relative velocities and sources with respect to the listener. Both linear and nonlinear loudspeakers produce this type of distortion.

12-81. INTERFERENCE DISTORTION.

12-82. SIGNAL CARRIER. Figure 12-46 illustrates two signals, separated slightly in frequency and differing in amplitude. The third waveform is the resultant obtained when the desired and undesired signals are combined algebraically at every point. The amplitude of the resultant varies at a rate equal to the difference in frequency between the two original signals. If both signals differ in frequency by 1000 cycles, the resultant waveform amplitude will change 1000 times per second. In an a-m receiver, this amplitude would be separated by the detector and heard as a whistle from the speaker. The resultant waveform may at times lead, lag, or be in phase with the desired signal. The resultant is therefore phase-modulated. This phase modulation (and, indirectly, frequency modulation) is directly proportional to the amplitude difference between the two signal carriers. When the amplitude ratio between the two signals is 2 to 1, the phase angle shift is slightly under 30 degrees. The rate of phase shift change is in direct proportion to the frequency difference between the two original signals.

12-83. STATIC. Static is primarily a form of amplitude distortion caused by uncontrolled electrical waves associated with thunderstorms and other natural phenomena. The strength of these waves is sometimes great enough to drown out the desired station or prevent clarity of reception. Limiter stages will limit incoming bursts of static amplitude, and, by selecting a narrow bandwidth, you can remove much of the continuous crackling variety of static through frequency selectivity. For fm reception, transmission allocations are in the higher-frequency bands, where static amplitude changes are not very effective; most of the outburst energy is limited to lower frequencies.

12-84. THERMAL AGITATION AND TUBE HISS. With no external natural disturbances or other station interference, there are still internal tube and circuit noises which will limit the weakest received signal to some minimum amplitude. Any signal lower than this minimum amplitude will not be amplified with clarity. Thermal agitation is the term applied to the noise created by the random motion of electrons in any conductor. The thermal noise produced is proportional to the amplifier bandwidth. Tube hiss is the term applied to the noise created by "shot effect," which refers to the fact that electrons moving through a tube are a congregation of separate particles which do not impinge on the anode as a continuous fluid movement, but rather as sporadic fluctuations. This tube hiss noise is normally distributed evenly throughout the frequency spectrum.

12-85. IMPULSE NOISE. Impulse noise, as distinguished from random noise, consists of external sharp bursts of energy. Normally, this noise is associated with automobile ignition systems and sparking gaps in electrical machines. A limiter stage is required to decrease the effects of this type of interference.

12-86. HUM. Hum interference is normally a result of insufficient filtering in the power supply, of heater-cathode leakage, of unshielded transformers and chokes in the dc line, or of defective coupling between circuits. This type of interference, as is true of other types of noise, will combine with the desired signal and produce distortion.

SECTION III

STABILITY PARAMETERS

12-87. GENERAL.

12-88. Stability is a term used to describe two different concepts. One is the steadiness or constancy of circuit characteristics, and the other is the property of certain circuits or parts of circuits which, when disturbed, create forces to restore the circuit characteristics back to an equilibrium condition. In connection with the steadiness concept of stability, the gain stability of an amplifier circuit and the frequency stability of an oscillator circuit are prime considerations. In connection with the restoration-of-equilibrium concept of stability, or the tendency of a circuit not to oscillate, feedback is an important consideration. This is a method of feeding a part of the output back into the input as a stabilizing factor.

12-89. STABILITY AND LINEARITY.

12-90. LINEAR, TIME-INVARIANT.

12-91. The term linear, time-invariant refers to the fact that parameters such as R, L, C, μ, G_m, r_p, etc, do not depend on the signal input applied to the circuit being considered or on time elements. The use of transient response, gain, and phase characteristics to describe a circuit is based on linear, time-invariant considerations. If all terms associated with the transient response of a circuit die out with time, the circuits under consideration are stable circuits. Step function responses are possible, and they eventually die out with time, to become stable. Figure 12-47 illustrates three typical step-function responses, all of which are stable; however, the relative stability of curve C is much better than that of curve A. The determination of the parameters or limits of the stability of a feedback network is desirable because you must know whether a circuit will oscillate when turned on, and thus damage the circuit elements, and you must know how to modify unstable circuit elements to improve circuit stability. When the amplitude of oscillation increases without limit, the circuit is not stable. The variation due to aging, temperature changes, replacements, etc, of resistors, capacitors, inductors, tubes, and semiconductors of an amplifier are reflected in the corresponding lack of amplifier gain stability. Actually,

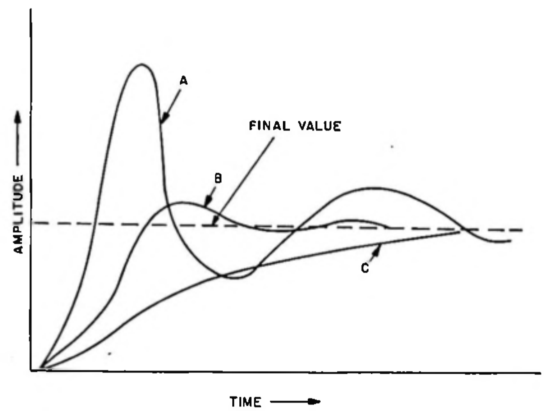


Figure 12-47. Typical Step Function Stable Response Curves

the worst enemy of circuit stability is the tube or semiconductor involved.

12-92. The criteria of general stability are based on linearly operating components and equipments. However, all circuit operation is nonlinear to some extent because common sources of nonlinearity consist of such things as poor motor characteristics, amplifier overloading noise, friction static, etc. If you exercise normal precautions, the linear theory will still apply quite well.

12-93. If the input signal is large enough to extend the operational range of a tube or semiconductor beyond its designed scope of operation, the output signal will be distorted. However, if negative feedback is now introduced, and the input signal is increased by the same amount the gain is decreased, the output will be undistorted and the amplitude will remain the same. The same effect as negative feedback can be obtained with a cathode-follower circuit because the output follows the input grid voltage. Unfortunately, a cathode follower circuit provides no signal gain unless you consider a signal loss as a negative gain.

12-94. Ultra-high-frequency oscillators will normally provide poor modulation linearity because the fully loaded self-excited oscillator will not oscillate below 20 to 40 percent of the normal dc plate voltage, the tank-circuit Q limits the percentage of modulation at higher frequencies, high positive modulation peaks may cause emission saturation (a decrease in average rf power), and the grid-to-plate capacitance may create undesirable phase relationships, which will cause nonlinearity and affect the stability of the stage.

12-95. NONLINEAR, TIME-INVARIANT.

12-96. The preceding paragraphs (12-87 to 12-94) were concerned primarily with stability within linear circuits. Stability in a linear circuit does not depend on the input variations. However, stability in a nonlinear circuit may depend on a particular input signal. This is true primarily because the feedback circuit elements may be stable for small input signals, but not for large input signals. Instability in a linear circuit will cause the response to increase without limit, whereas oscillations with a constant amplitude can, and do, exist in a nonlinear circuit. In the linear circuit, only the frequencies present at the input or generated within the circuit can be present in the response. But, in the nonlinear circuit, the output may consist of harmonics, subharmonics, or beat frequencies if the input contains more than one frequency. Therefore, various types of stability must be defined for the nonlinear circuit. For example, the definition must include the tendency of an oscillator to maintain its period and amplitude of oscillation. Normally, for study purposes, the nonlinear input must be represented by a sinusoidal input with an equivalent phase and gain, with both the phase and gain depending on the amplitude of the input signal. The only frequency output considered with respect to phase and gain is the fundamental waveform. The nonlinearity does not introduce any subharmonics, and the circuit must filter out all higher frequencies.

SECTION IV

TRANSIENT RESPONSE

12-97. GENERAL.

12-98. DESCRIPTION.

12-99. A transient is a nonperiodic surge of voltage or current. Transients may be produced by the application of discontinuous signals to the input of a circuit or device or by a sudden change in circuit conditions, as when a switch is closed. Therefore, the transient response of a circuit or device is its effect on the voltage and current when a sudden change in the input signal or circuit conditions occurs.

12-100. LINEAR VERSUS NONLINEAR CIRCUITS.

12-101. The ability of a linear circuit or device to handle discontinuous signals or pulses is determined by the transient response of the stage. The steady state (forced response) must be calculated as zero while you determine the transient response of a stage. The calculation of transient response by this method will apply only to a linear circuit because the nonlinear circuit response cannot be broken down or decomposed into different forced and transient components. However, when a linear constant parameter stage is excited by a step pulse such as shown in figure 12-48, the forced response (steady state component) remains constant with time, and may be zero. Therefore, the total response shape will depend only on the transient component. The step response of a linear con-

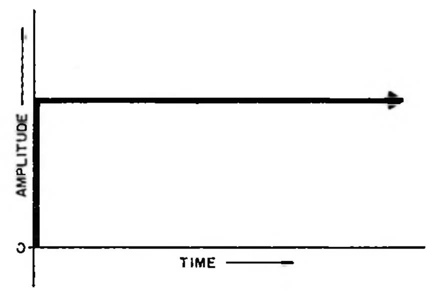


Figure 12-48. Input Step Function

stant parameter (that is, r_p, C, etc, do not change with input signal changes) can be displayed on an oscilloscope and measured directly. Unfortunately, when you are attempting to measure a step response through a linear time-varying stage (that is, r_p, C, etc, change with input signal fluctuations), you will have difficulty separating the forced component from the transient response component when the total response is displayed on an oscilloscope. This section deals only with the measurement or illustration of linear constant parameter stages, The response to a unit impulse, as shown in part B of figure 12-49, is actually the time derivative of the stage response to a unit step, as shown in part A of figure 12-49.

12-102. MEASUREMENT TECHNIQUE.

12-103. The input signal to the amplifier or other electronic device should be a rectangular pulse with the stage output being applied

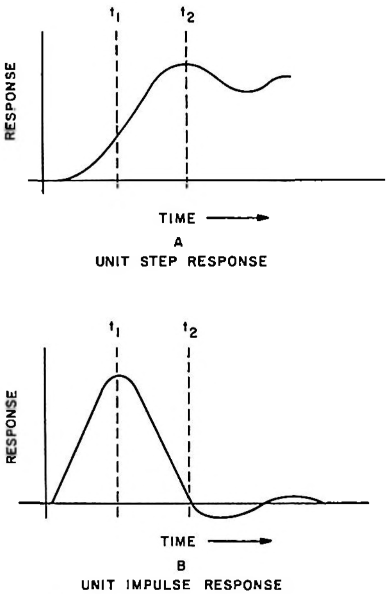


Figure 12-49. Linear Constant Parameter Amplifier Responses

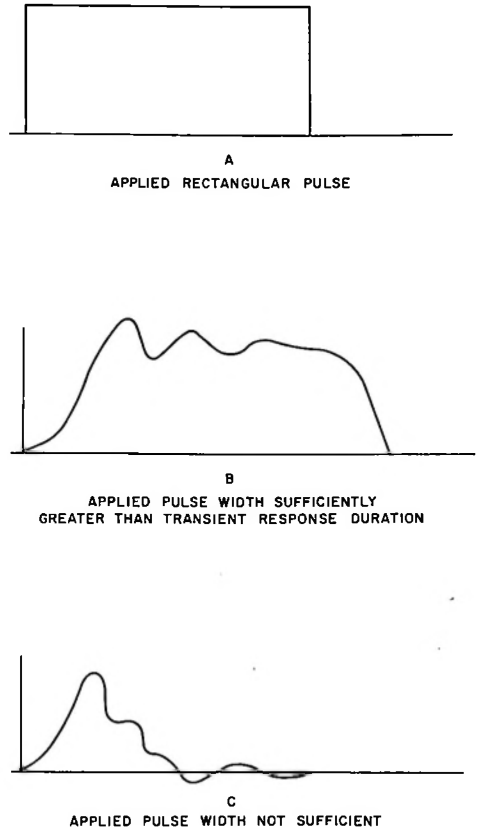


Figure 12-50. Comparison of Applied Pulse Width and Transient Response Times

to an oscilloscope to provide a visual display. The half-period of the input rectangular pulse must require a longer time than the transient response duration. The oscilloscope should be adjusted to show the width of a single complete pulse or less. The results of applying a proper pulse versus a too narrow pulse is shown in figure 12-50. The transient response of a stage, with an input square wave, is measured by viewing two separate response characteristics. The first of these is related to the leading edge of the output response curve, and is composed of the rise time, time delay, and overshoot. The second of the response curve characteristics is related to the flat-top portion of the curve, called sag. Sag is possible in a circuit only if the circuit is not capable of passing dc currents. In order to examine the leading edge of the wave, as shown in part B of figure 12-51, you should use a fast sweep rate on the oscilloscope, whereas a slow sweep rate should be used to illustrate the flat top, as shown in part C of figure 12-51. Wide input pulses normally have long rise times. Therefore, narrow input pulses with short rise times are used to obtain response pulse leading-edge measurements. Wide pulses should be applied to the input for response pulse sag measurements. Figure 12-51 illustrates a pulse width of 5 microseconds as being adequate for leading-edge measurements, but 1000-microsecond input pulses are required for the flat top sag portion of the applied pulse.

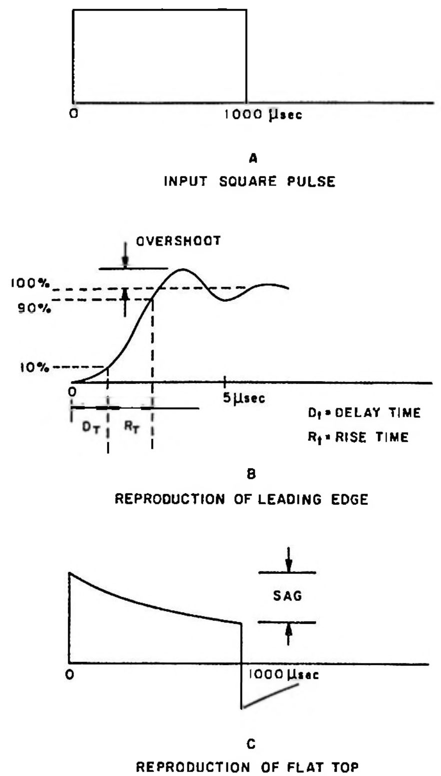


Figure 12-51. Transient Response Characteristics

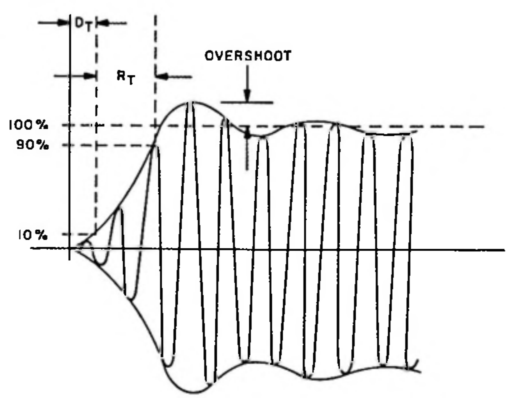


Figure 12-52. Typical Transient Response of a Tuned Stage

12-104. When a radio-frequency signal rather than a rectangular pulse is used as the input to a tuned stage or band-pass device undergoing transient response measurements, the parameters of the response are related only to the leading edge because there will be no flat top characteristics. The response of a band-pass device or tuned stage is illustrated in figure 12-52.

12-105. TRANSIENTS.

12-106. The total response in a linear circuit includes all of the individual transients due to the store of energy in each inductor, capacitor, and external energy source connected to the circuit, plus the steady state (forced response) of each external applied energy source. You can compute the response by starting at any arbitrary time (T = 0), where all of the initial energy conditions of the proposed circuit are known.

12-107. ELEMENTS.

12-108. REACTIVE. The discharge of a capacitive element through a resistor requires the time T_2 minus the original starting time (T_1) for the voltage or current to decay down to 37 percent of the original value. The same analogy applies if the capacitive or inductive component charges up to 63 percent. This type of analysis may be used for any periodic applied voltage. The steady state current and voltage for an applied voltage are determined, the periodic voltage is resolved into its individual harmonic components, and then the transient is determined. The transient waveform does not bear a relationship to the applied voltage, as the transient waveform depends only on the circuit constants and the initial current and voltage conditions.

12-109. RESISTIVE. Time is considered to be zero when no reactive elements are present. Pure resistance elements do not charge or discharge with time.

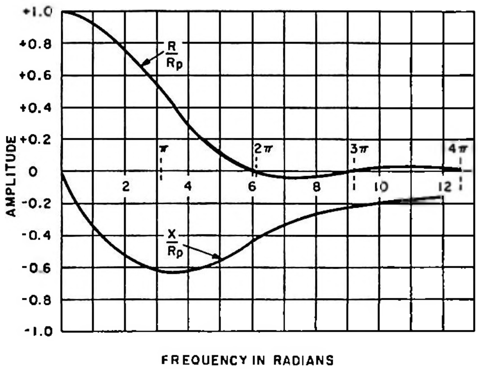


Figure 12-53. Series Resistive and Reactive Diode Components Represented as a Function of Frequency and Transit Times

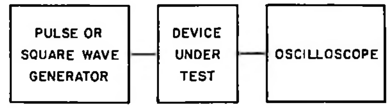


Figure 12-54. Typical Test Setup for Measurement of Transient Response in Low-Pass Equipment, Block Diagram

12-110. HIGH-FREQUENCY TUBES. At ultra-high frequencies, the transit time of an electron traveling between the cathode and the plate of a tube constitutes an appreciable part of the input cycle. This contributes a conductance element to the grid input admittance because as an electron passes the grid it introduces a grid current, even if it does not strike the grid. The grid conductance increases as the square of the applied frequency. The transit time effect on an average tube can be visualized if you realize that the input resistance to a type 57 tube is several megohms at a frequency of 5 megacycles, while at 100 megacycles the resistance drops to about 1500 ohms. In triode tubes the transconductance of the tube decreases and lags behind a greater amount as the transit time increases. The amplification factor also decreases and the phase angle increases. The transit time of a diode, at ultra-high frequencies, causes the dynamic plate resistance of the diode to decrease. Therefore, the cathode-to-plate resistance must be represented by a series resistor and capacitor. Figure 12-53 illustrates the curves of both the resistance and the capacitance, each with respect to transit time, as the frequency increases. As you can see, the resistance R/R_p eventually oscillates about a zero reference level, and is sometimes negative. Much of the tube tranit time difficulty has been solved by placing the electrodes closer together.

12-111. MEASURING EQUIPMENT.

12-112. GENERAL. The device being tested contains a definite transit or rise time to be measured. However, the equipment used to test the device may affect this rise time. In fact, if the rise times of the pulse generator and oscilloscope are each less than 10 percent of the rise time of the device being measured, an accuracy within less than 1 percent can be obtained. To partially compensate for the sag on the rectangular pulse, introduced by the testing equipment, the individual sags may be subtracted from the final measurement to obtain the correct value, that is, assuming that the sag introduced by the equipment is small. Figures 12-54 and 12-55 illustrate the required test setup for the measurement of transient response within linear equipment. The delay time measured with the aid of test setups shown in figures 12-54 or 12-55 will result in a larger delay time than is actually contained in the equipment itself because of the additional time introduced by the test equipment. However, the test equipment time can be directly subtracted from the final measured result; the measuring equipment will also tend to re-

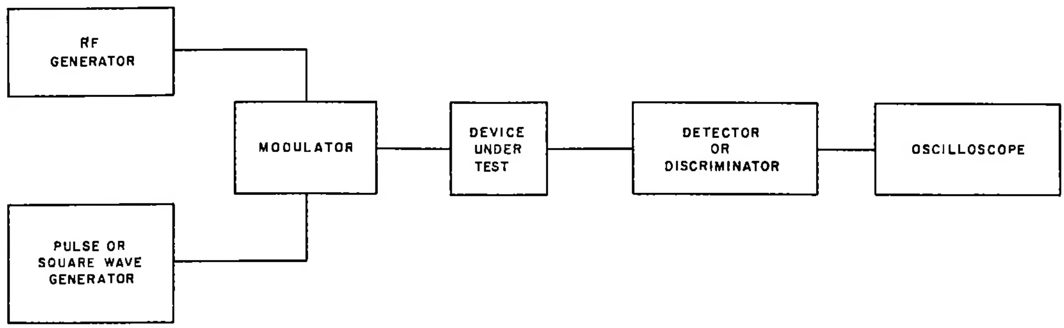


Figure 12-55. Typical Test Setup for Measurement of Transient Response in Bandpass of FM Equipment, Block Diagram

duce the leading edge overshoot of the waveform in the device being tested.

12-113. SYNCHROSCOPE. The requirements of test equipment used to measure transient response are extremely rigid, as illustrated by several special features of the oscilloscope. A triggered oscilloscope, referred to as a synchroscope, requires a variety of sweep speeds. The sweep circuit may be triggered by the trigger pulse which starts the pulse generator; the response being measured may be used to trigger the sweep so that later transient responses may be measured; or the applied signal may be used to trigger the sweep circuits of the synchroscope and then passed through a delay line to the circuit under test.

12-114. PULSE GENERATOR. The pulse or square wave generator must contain a wide range of available pulse widths and frequencies; the leading and trailing edges of the output pulse must be short as compared with the pulse width; the sag should be flat and contain no ringing or oscillation; finally, the carrier frequency pulses must be relatively free from frequency modulation during the active pulse.

12-115. TEST EQUIPMENT CONNECTION. The connection of test equipment to the device being tested is extremely critical because the internal impedance of the test equipment can load down the equipment under test and thus cause response distortion. A cathode follower stage or other isolating device should be employed between the test equipment and the device under test to minimize the loading effect. All connecting leads should be maintained as short as possible, and the connecting lines must be matched to prevent feedback reflection along the line at high frequencies, which would cause spurious ripples or ringing.

12-116. TRANSISTORS.

12-117. The transient response of a transistor used as a switch is important because of the time required to turn the transistor switch from the "off" to the "on" position, and vice versa. Normally, either a step or a pulse of current applied to the input is required to turn the transistor from "off" to "on." Referring to figure 12-56, which is the high-frequency equivalent circuit, you can make some assumptions and simple calculations which will provide you with an approximate rise time within a transistor circuit operating in an active region. First, assume that load resistor R_L is small enough to represent a short circuit; therefore R_3, (the collector resistance) and C_C (the barrier capacitance) are effectively in

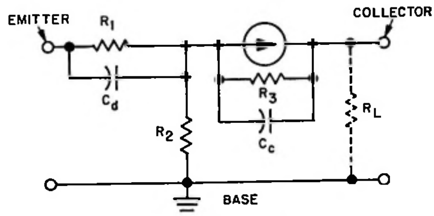


Figure 12-56. Transistor Equivalent Switching Circuit

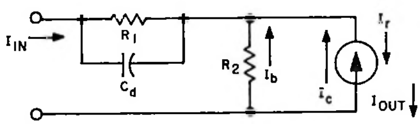


Figure 12-57. Simplified Transistor Equipment Switching Circuit with Small Output Load

parallel with R_2 (the barrier resistance). Now, since R_2 is very much smaller than R_3, we can neglect R_3 in this parallel circuit. Furthermore, as C_C (the barrier capacitance) is much larger in reactance than R_2 at the assigned frequency, $2\pi F$, you can neglect C_C also. Thus, for a grounded-base configuration, the equivalent circuit of figure 12-56 can be reduced to the more practical equivalent circuit shown in figure 12-57. The value of C_D (the diffusion capacitance) is equal to the reciprocal of the assigned frequency, $2\pi F$, and the emitter resistance, R_1

$$\left(\frac{1}{2\pi F_1 R_1} \right).$$

Next, you should include a subsidiary circuit (R and C), as shown in figure 12-58. The time of RC equals the reciprocal of $2\pi F_1$. When the time T equals zero, output I_R equals zero, and when time T equals infinity, output I_R equals the input (I_{in}). The time constant is the reciprocal of the

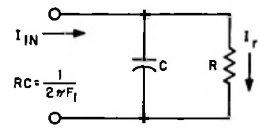


Figure 12-58. RC Circuit, Simulating Frequency Dependence, Used to Calculate I_R

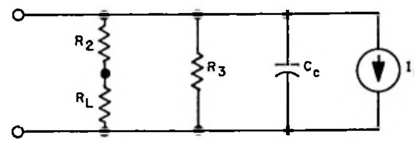


Figure 12-59. Simplified Equivalent Transistor Switch Circuit with Large Output Load

assigned frequency $\left(\frac{1}{2\pi F_1} \right)$, and the rise time is $\left(\frac{2.2}{2\pi F_1} \right)$. This is the time required for the output current to rise from 0.1 to 0.9 percent of its final value. The effect on transient response of C_c (the collector capacitor) was not taken into account in your previous calculations because load resistor R_2 was considered small enough to be an effective short circuit. Figure 12-59 illustrates a simplified equivalent circuit which recognizes that R_L does not affect the response when the input is from an infinite series-impedance current source and $R_L C_c$ is very much larger than the time constant $\frac{1}{2\pi F_1}$. In fact, if $\frac{R_2}{R_3}$ is very much smaller than 1 and $2\pi F_1 R_3 C_c$ is very much larger than 1, then the time constant, calculated for the response with a very small or shorted value of R_L, would be increased by the amount, $2\pi F_1 R_L C_c + 1$. Therefore, the rise times and turn-on times would be increased by this same factor.

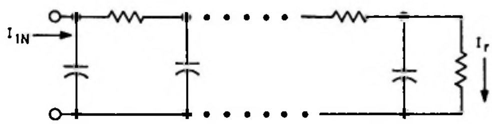


Figure 12-60. Transistor Switch Equivalent Delay Time Circuit

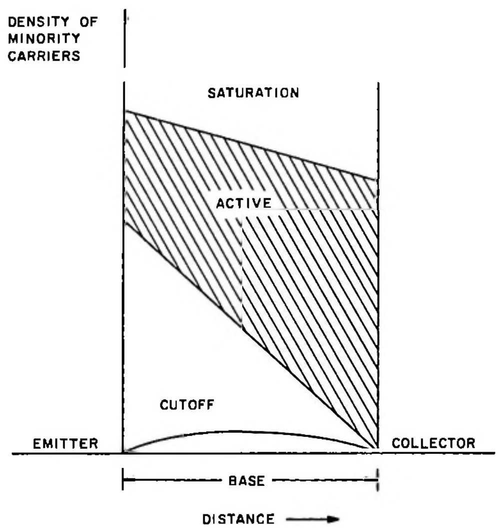


Figure 12-61. Comparative Minority-Carrier Densities in Transistor Base for Cutoff, Active, and Saturation Regions of Operation

12-118. DELAY AND STORAGE TIMES.

12-119. TRANSISTOR DELAY TIME. In our previous calculations of transistor transit response, we neglected the finite time required for the impulse signal to diffuse across the base region. Under this condition, if a current pulse is applied to the transistor emitter, a response will appear at the collector only after some delay in time. The value of I_R may be obtained for use in the equivalent circuits illustrated in figures 12-57 and 12-59 by representing this delay time with an equivalent circuit, as indicated in figure 12-60. The line attenuation will increase in direct proportion to the frequency. The resistive-capacitive delay line indicates a time interval before a response is indicated at the output.

12-120. TRANSISTOR STORAGE TIME. Our discussion up to this point has concerned the turn-on transistor process. If the transistor is in the active region, the turn-off process consists simply of applying a pulse of reverse polarity, and the required time constant is calculated in the same manner, using the equation for the turn-on process. Unfortunately, if the transistor is in the "on" condition and is operating in the saturation region, an abnormally large time delay will occur before the transistor responds to the turn-off signal. This peculiar delay, termed storage-time delay, is illustrated by figure 12-61, which shows the minority carrier density in the base region for three situations. The first situation illustrated is the cutoff condition, with both the emitter and collector junctions back-biased. The minority carrier density is therefore zero at both junctions, and very small throughout the base region. The second situation is illustrated by the active curve in figure 12-61, where the minority carrier density is high at the emitter junction, and zero at the collector junction. The change in density between the two junctions is the result of the diffusion process which accounts for current flow across the base of the transistor. However, if an input signal drives the emitter junction to a back-bias condition, the diffusion process will not continue until the minority carriers in the base region have been removed. The third situation is illustrated by the saturation curve in figure 12-61. Unlike the cutoff and active conditions, the previously discussed equivalent circuits used to determine time delays, response, etc, do not apply in this case. This is true because both the emitter and the collector are emitting carriers into the base region. In addition, as both junctions are forward-biased, the junction voltages

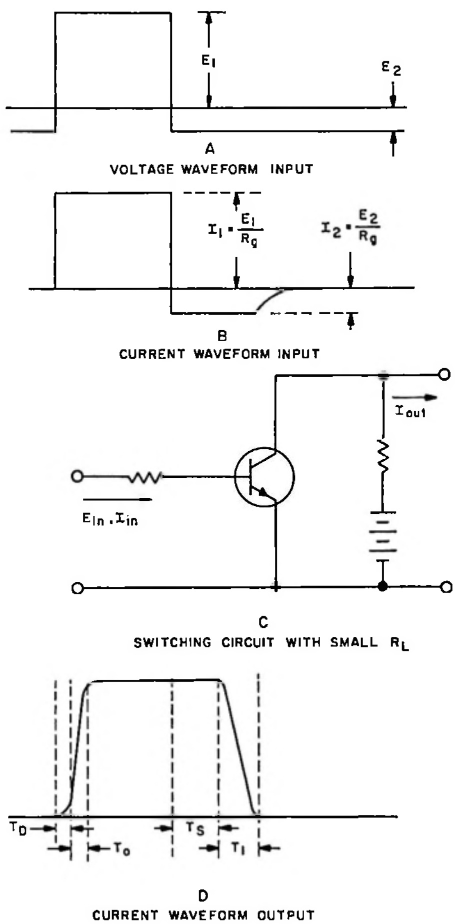


Figure 12-62. Grounded-Emitter Switch Circuit with Input Voltage and Current Waves and Output Current Response Waveform for Small R_L

will be small and the collection process at the junctions will be slow. This, in turn causes the density of minority carriers in the base region to build up to a large value. This high density level in the base region must be permitted to decrease before the turn-off process begins to take effect. This long-storage time delay may represent two or three times the normal rise or fall time in the active region. From the foregoing it is evident that when a transistor switch is used in an application requiring high switching speeds, it must be restrained from entering the saturation region.

12-121. TRANSISTOR RESPONSE.

12-122. Figure 12-62 illustrates the approximate waveforms required to represent the response from a transistor driven from cutoff to saturation and back again. A grounded-emitter switch is used because this type of configuration is the most useful and its response can represent other configurations. In figure 12-62, the delay time is represented by the symbol T_D, the rise time by T_O, the storage time by T_S, and the decay or fall time by T_1. The input current reverses at the end of the pulse rather than falling only to zero. As a result, the output current response falls toward a negative value rather than toward zero. The fall time of the pulse is thereby reduced. The voltage input waveform of figure 12-62 was terminated in a minus E_2 voltage because the storage of minority carriers in the base region does not permit the transistor input impedance to immediately attain a large value. In fact, the input impedance remains small until the minority carriers at the transistor junctions are swept away. At this point the input impedance increases and causes the input current to decrease in direct proportion to the speed with which the minority carriers throughout the base region drift to the junctions.

SECTION V

SPECTRUM AND SPECTRUM ANALYSIS MEASUREMENT

12-123. GENERAL.

12-124. ELECTROMAGNETIC FREQUENCY SPECTRUM.

12-125. A chart showing electromagnetic frequency spectrum is given in figure 12-63. This chart indicates the frequency and wavelength of the various frequency bands of interest to you. To help you with the minor mathematics involved in this section, a wavelength-frequency conversion chart has been provided in figure 12-64. This chart, in conjunction with table 12-1, which contains tabular multiplying factors, covers the major portion of the electromagnetic-wave spectrum. The analysis of a complex waveform, prepared in terms of a graphical plot of the amplitude versus frequency, is known as spectrum analysis. Spectrum analysis recognizes the fact that any waveform is composed of the summation of a group of sinusoidal waves, each of an exact frequency and all existing together simultaneously.

12-126. ACOUSTIC SPECTRUM.

12-127. The acoustic spectrum chart illustrated in figure 12-65 is provided to show the limits of the sound spectrum as set by human ear sensitivity. This chart is based on a musical pitch of 440, a physical pitch of 426.667, and an international pitch of 435. Complex waveforms are divided into two separate groups - periodic waves and non-periodic waves. Periodic waves are those which contain the fundamental frequency and its related harmonics. Non-periodic waves are those waves which contain a continuous band of frequencies, resulting from the repetition period of the fundamental frequency approaching infinity

Table 12-1. Auxiliary Wavelength-Frequency Conversion Table*

FOR FREQUENCIES FROM	MULTIPLY F BY	MULTIPLY λ BY
0.03 -- 0.3 mc	0.01	100.000
0.30 -- 3.0 mc	0.10	10.000
3.00 -- 30.0 mc	1.00	1.000
30.00 -- 300.0 mc	10.00	0.100
300.00 -- 3000.0 mc	100.00	0.010
3000.00 -- 30000.0 mc	1000.00	0.001

* Refer to figure 12-64.

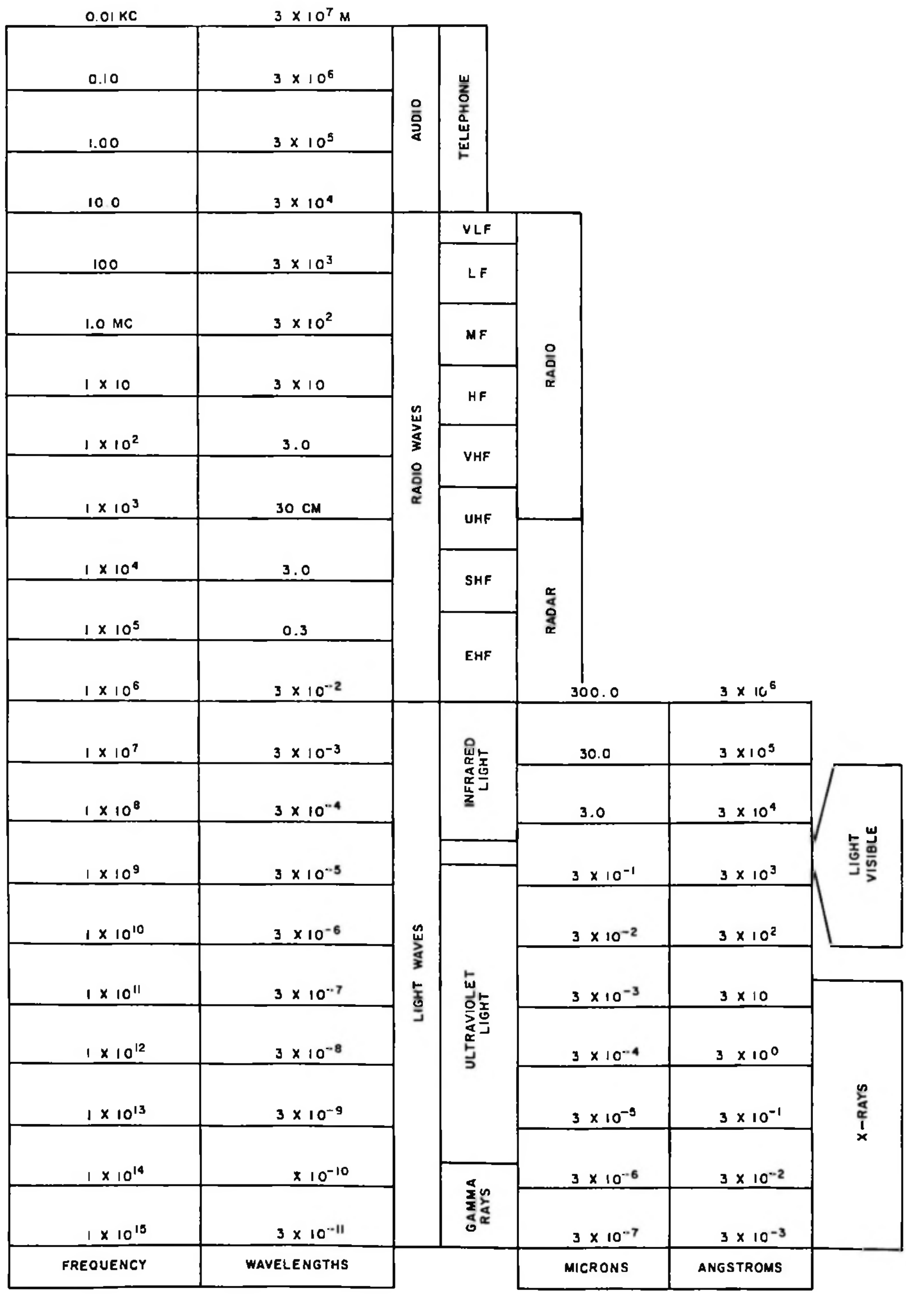


Figure 12-63. Electromagnetic Frequency Spectrum

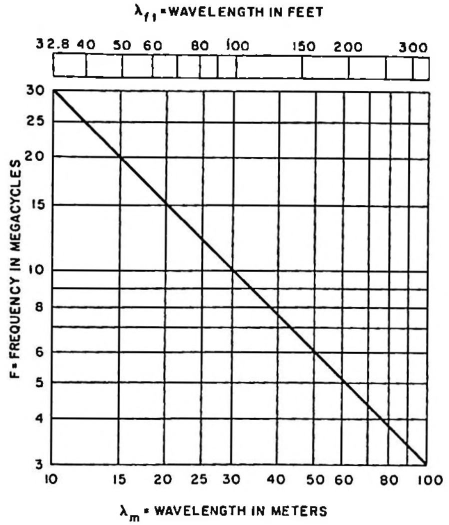


Figure 12-64. Wavelength-Frequency Conversion Chart

and thereby creating a continuous frequency spectrum.

12-128. SPECTRUM SOLUTION.

12-129. Variations to be measured may continuously change in frequency because of a mechanical function caused by varying the tension on a string or because of a capacitance variation in an oscillatory circuit. Methods of determining the instant frequency can be used to find the spectrum distribution when a high-frequency current is frequency-modulated because of a sinusoidal variation in the circuit capacitance, code modulation, direct capacitance modulation, direct frequency variation, or inverse capacitance modulation. This section deals with simple limits such as the requirement that the frequency (f) must be small in com-

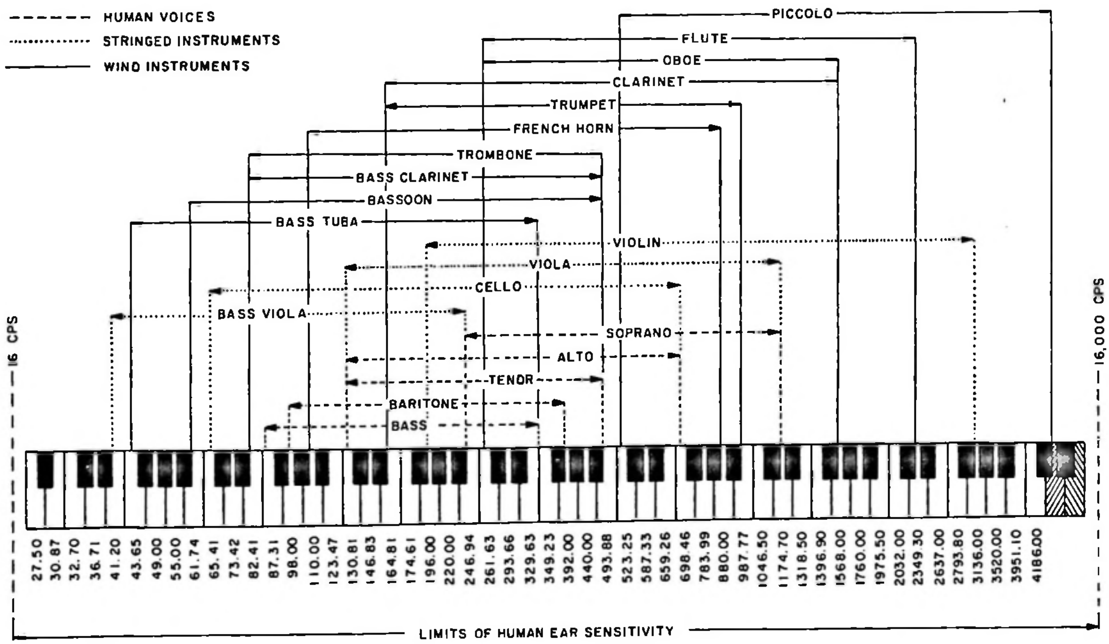


Figure 12-65. Acoustic Spectrum

Table 12-2. Abbreviated Bessel Factor Table

β	$J_0(\beta)$	$J_1(\beta)$	$J_2(\beta)$	$J_3(\beta)$	$J_4(\beta)$	$J_{14}(\beta)$
0.0	1.0000	0.0000	0.0000	0.0000	0.0000	
0.5	0.9385	0.2423	0.0306			
1.0	0.7652	0.4401	0.1449	0.0196	0.0025	
2.0	0.2239	0.5767	0.3528	0.1289	0.0340	
5.0	−0.1776	−0.3276	0.0466	0.3648	0.3912	
10.0	−0.2459	0.0435	0.2546	0.0584	−0.2196	0.0120

parison with the carrier deviation frequency (ΔF), and that (ΔF) must be small in comparison to the mean (center) frequency (F).

12-130. Spectrum analysis and measurements are widely used in microwave engineering when testing or observing the operation of pulsed oscillators or magnetron tubes. Performance measurements of a-m and fm radio-frequency transmitters, noise measurements, and wide-range signal searching techniques employ the spectrum analyzer. In the audio-frequency range, spectrum analysis is used to aid in the design of telephone and telegraph transmission equipment, to study sounds such as voice or music, to locate mechanical noise, etc. Spectrum analyzers are also used in the very-low-frequency range for application in the fields of electro-medicine, geological surveying, and oceanographic measurements.

12-131. SPECTRUM DISTRIBUTION.

12-132. AMPLITUDE MODULATION.

12-133. The modulation energy in an amplitude-modulated (a-m) wave is contained entirely within the sidebands. When the modulation signal is a sinusoidal voltage, the frequency spectrum distribution is contained within one pair of sidebands. The frequencies of the sidebands are (F — f) and (F + f), where F is the carrier frequency and f is the modulation frequency.

12-134. FREQUENCY MODULATION.

12-135. SIDEBANDS. Theoretically, unlike a-m, which can contain only two side currents, fm can contain an infinite number of side currents of frequencies, F ± f, F ± 2f, F ± 3f, etc, in addition to the current of the carrier frequency, F. Therefore, the modulation energy is not entirely contained in the first pair of symmetrical side currents as it is in a-m. The energy contained in fm is spread over several side currents. The amplitude of a particular pair of side currents may be larger than the center-frequency component. This fact also holds true for phase modulation, in that, with the same modulation index (β), you will obtain the same spectrum distribution. The number of important side currents is larger for low frequencies in the signal band than it is for high frequencies.

12-136. AMPLITUDE FACTORS. Table 12-2 provides selected amplitude factors that you must use to multiply the maximum unmodulated carrier current level (I_m) to find the amplitude value of an individual side current pair in the frequency spectrum. As an example of the use of table 12-2, use a maximum center frequency swing (ΔF) = ± 60 kc and a 30-kc signal frequency (f) to

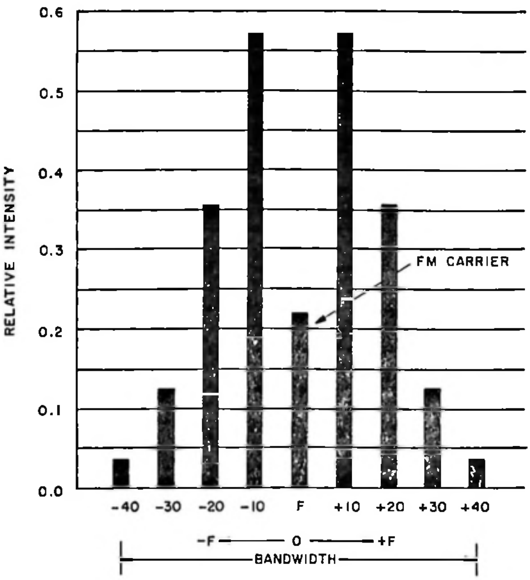


Figure 12-66. Spectrum Distribution for a Modulation Index of $\beta = 2$

find the value of $\beta = \frac{\Delta F}{f} = \frac{60}{30} = 2$. Table 12-2 illustrates that $J_0(\beta) = J_0(2) = 0.2239$, and this is the value multiplied times the maximum unmodulated current (I_m) to obtain the magnitude of the center frequency component (F). The value of $J_1(\beta) = J_1(2) = 0.5767$ multiplied times I_m gives the amplitude of both the first upper side current at the frequency of (F + 30 kc) and the first lower side current at the frequency of (F — 30 kc). The spectrum distribution for a modulation index of $\beta = 2$ is illustrated in figure 12-66. This graph was prepared from data obtained from table 12-2. As all amplitudes are obtained by multiplying the Bessel factors obtained from table 12-2 times I_m, the magnitude of the Bessel factors directly determines the intensity of the side current pairs in the useful frequency spectrum. The side current pairs which are too far from the center frequency (F) are not of sufficient amplitude to be significant. The bandwidth is determined by the number of significant side currents.

12-137. The bandwidth may also be calculated from table 12-2 for a specific frequency swing. For example, let $\beta = 10$ and read the Bessel function values across the chart, from left to right. As you can see, the 14th sideband pair is $0.012\ I_m$, which is not significant. Therefore, the maximum bandwidth for $\beta = 10$ is 2 (both side currents) times 14 sidebands times the signal frequency deviation. For a 30-kc signal deviation, the bandwidth would be 2 x 14(30) = 840 kc, and for a 2-kc note it would be 56 kc. This example shows why a higher-modulation signal requires more frequency space (that is, greater bandwidth) than does a lower-frequency modulation signal. The Bessel factors given in table 12-2 also show that the greater the modulation index (β), the greater the number of significant sidebands. As the modulation energy spreads out from the carrier frequency (F), you can determine at what point you should stop the carrier. For example, in figure 12-67, which was prepared for an index of $\beta = 24$, the largest side currents occurred at the edges of the pass band; thereafter you can continue to read off the Bessel factors until the side current pairs are only 5 percent or less of the unmodulated carrier current level, I_m. Sideband currents beyond this point do not have significant amplitudes for practical consideration.

12-138. Frequency modulation, for modulation index values smaller than 0.2, is similar to amplitude modulation, in that both types of modulation contain only one significant side current pair. Therefore, for a value of $\frac{\Delta F}{f} = \beta = 0.2$ or less, fm behaves exactly like a-m with respect to spectrum distribution. However, unlike a-m, no primary oscillator can be used in the fm transmitter because you must be able to change

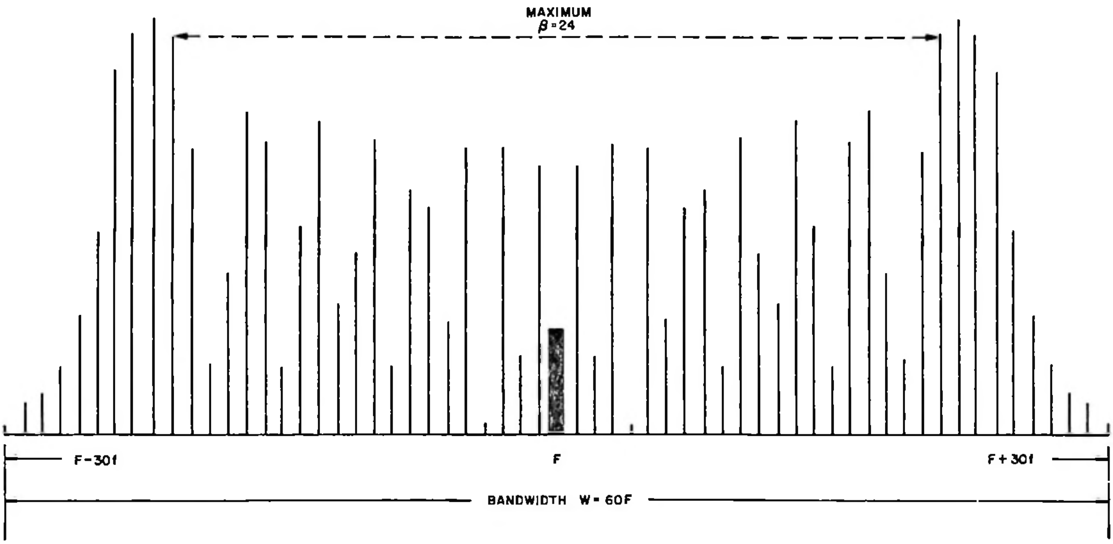


Figure 12-67. Spectrum Distribution for an Index of 24

the carrier frequency during the modulation cycle to produce fm. The desired intelligence in fm or phase modulation (pm) will create more energy distribution, and thus a larger response in a receiver demodulator, than will noise energy. This is the outstanding desirable feature of fm over a-m spectrum distribution.

12-139. PHASE MODULATION.

12-140. In pm, unlike fm or a-m, the carrier current level (I_m), as well as the center carrier frequency (F), remains constant. Only the relative phase (θ) changes. The actual value of θ is not important. The deviation from this value is important, and produces the desired pm. For example, if θ equals 30 degrees and the phase deviation is $\pm$ 40 degrees, this will produce a certain modulation of the carrier. However, if θ had originally been 80 degrees rather than 30 degrees and was subjected to a $\pm$ 40-degree deviation, the same output pm waveform, containing the same intelligence, would have been produced. The original value of θ indicates only the original amplitude at the start of the phase swing. Actually, the effect of a $\pm$ 40-degree deviation would be to give the appearance of wobbling about the carrier frequency (F) as F goes through an angular distance of $\pm$ 360 degrees, regardless of its beginning phase, θ. However, the equivalent instantaneous frequency of the modulated carrier would remain the same. Pm entails the important fact that it is created not only by the maximum phase deviation ($\Delta\theta$), but also by the applied signal frequency (f). Therefore, the frequency shift is greater for higher frequencies than it is for lower frequencies.

12-141. SIDEBANDS. The sidebands foı pm are similar to the sidebands for fm, and the same general formula for the modulation index holds true (that is, $\beta = \frac{\Delta\theta}{f}$).

As with fm, the side currents contain a symmetrical frequency distribution around the carrier. In other words, the first upper side band and the first lower side band have the same numerical value, amplitude, and frequency difference from the carrier; the value of the modulation index (β) is proportioned to the phase deviation ($\Delta\theta$); and for a fixed maximum phase swing it does not matter whether you use a 15-cycle or 15-kc signal to modulate the phase of the carrier. In either case, you will obtain the same number of important side current pairs. However, for the 10th upper and lower sideband in the 15-cycle modulation case, the side current pair is 10 x 15 cycles or 150 cycles above and below the carrier; for the 15-kc case the 10th side current pair is 10 x 15 kc or 150 kc above and below the carrier, and thus requires a much broader bandwidth. Pm and fm, unlike a-m, have a spectrum distribution of the modulation energy which is proportional to the square of the spectrum amplitudes, and does affect the carrier frequency amplitude. However, it does not matter whether you are dealing with a modulation index of 10 for fm or a maximum phase deviation of 10 radians (573 degrees) for pm. This is true for pm as long as the maximum phase swing ($\Delta\theta$), which causes the pm, is fixed. The same spectrum distribution is equally true for fm if the maximum frequency deviation (ΔF), which causes the fm, changes directly with the signal frequency (f). As the bandwidth of pm remains the same regardless of signal frequency changes, the pm bandwidth can be calculated directly from the modulation index (β). For small values of modulation index (0.2 or less), pm will contain only one pair of significant sidebands, the same as for a-m. Actually, this permits you to amplitude-modulate a carrier, suppress the carrier, and shift the modulation product by 90 degrees to provide narrow-band fm or pm. The fm or pm effect can be obtained by applying a signal voltage having a magnitude which is inversely proportional to the signal frequency (f). The phase-shifted product is then combined with the unmodulated carrier.

12-142. Both fm and pm contain a true instantaneous and an equivalent instantaneous carrier frequency. Therefore, the number of cycles per second must remain constant to prevent the center or mean frequency (F) from drifting to some other nearby frequency not originally assigned to the carrier. If this occurred, the entire spectrum centered around the carrier would drift and infringe on other nearby fm channels.

12-143. SPECTRUM ANALYZERS.

12-144. REQUIREMENTS.

12-145. The type of data available will determine the type of spectrum analyzer to be used. If an amplitude-versus-time plot is available, you can use graphical methods of analysis. Some graphical methods involve calculations and tabulations, while others require only mechanical devices to determine the frequency components of a complex wave.

12-146. If you know the voltage representing the complex wave at a particular point in the electrical circuit, you can use the point-to-point or the continuous type of spectrum analyzer to examine the individual frequency components. The point-to-point analyzer is used to measure the amplitude of each frequency component individually, whereas the continuous type of analyzer sweeps the entire frequency range of interest and then presents a simultaneous display of all frequency components in the complex wave.

12-147. PULSED RF WAVES. Microwave signals are normally pulse-modulated. This is especially true in radar facilities where the ratio of reflected power to transmitted power is very small and the transmitter's average power level is made small by the use of a high peak-power pulsed oscillator using a short-duration pulse. The radar receiver normally requires an output pulse of constant amplitude and frequency. This pulse has a duration of between 0.1

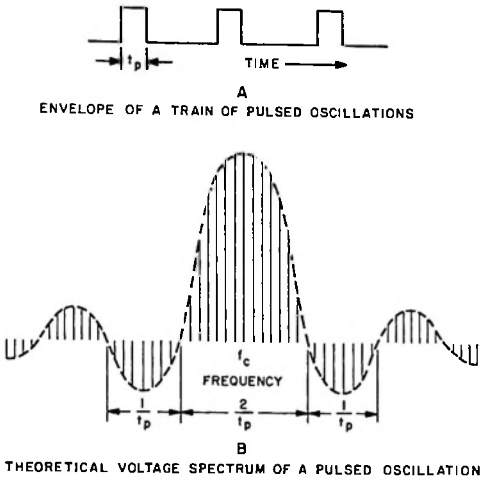


Figure 12-68. RF Pulses

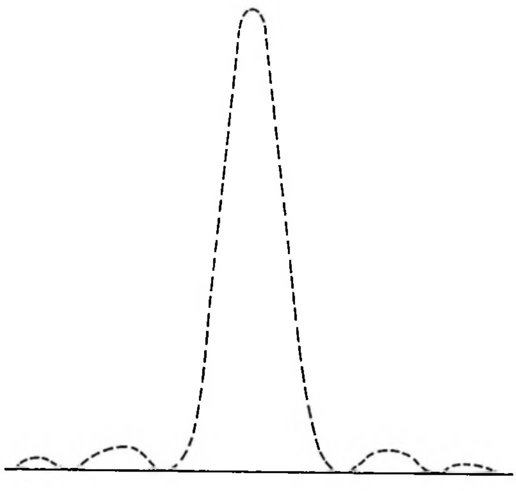


Figure 12-69. Envelope of an Ideal Power Spectrum of a Pulsed Oscillator

and 5 microseconds, with a 500 to 2500-microsecond duration between pulses. Therefore, the leading and trailing edges of the pulses must be extremely steep, with a constant amplitude between them. Incorrect pulse shape will cause frequency spread and pulling, which results in less available energy at the frequency to which the receiver is tuned. Spectrum analysis is required to determine the amount of amplitude and frequency modulation existing in the rf pulse. Figure 12-68 illustrates a train of pulsed oscillations and the voltage spectrum of a pulsed oscillation. However, the voltage spectrum is of secondary importance because the power spectrum obtained by simply squaring the amplitudes of the voltage spectrum, is of primary interest in spectrum measurement. Figure 12-69 illustrates the ideal power spectrum pattern. Any major deviation from this form can be determined with a microwave spectrum analyzer during the radar oscillator performance test.

12-148. Observing part B of figure 12-68, you will notice that the spectrum plot is symmetrical about the center, or mean, frequency (f_c). The spectrum has maximum amplitude at frequencies separated from (f_c) by harmonics of the prf. Zero amplitudes occur at distances determined by the time of pulse duration (T_p).

12-149. HETERODYNE SPECTRUM ANALYZER.

12-150. Figure 12-70 shows the basic block diagram of a narrow-band superheterodyne receiver with a local oscillator operating on a frequency which can be varied as a function of time. The use of the leading edge of a sawtooth waveform causes a linear time variation and synchronizes the linear time variation with the horizontal deflection of a cathode-ray-tube beam. The linear sawtooth variation is at a low frequency; consequently, many cycles of the pulsed microwave signal are sent to the analyzer during each frequency swing of the local oscillator. The incoming pulses are mixed with the local oscillator frequency to produce a difference frequency. This output is injected into a narrow-band intermediate-frequency amplifier. The amplifier output is detected, amplified, and applied to the cathode-ray tube, as shown in figure 12-70.

12-151. RESOLVING POWER. The resolving power of a continuously tuned spectrum analyzer is a measure of the analyzer's ability to distinguish between two adjacent frequencies of equal amplitude. The resolving power is a function of both the sweep rate and the i-f bandwidth. This is true because the amplifier has a fixed bandwidth and center frequency, and its input signal frequency is varying with time. If the circuit has a high Q, the output will oscillate (ring) just past the maximum signal amplitude. The solution to this problem is to select a Q which causes the steady-state response to be equal to the rise and fall time of the transient, as measured at the 3-db (half-power) points. The longest pulse duration being used determines the maximum permissible i-f bandwidth consistent with adequate resolution.

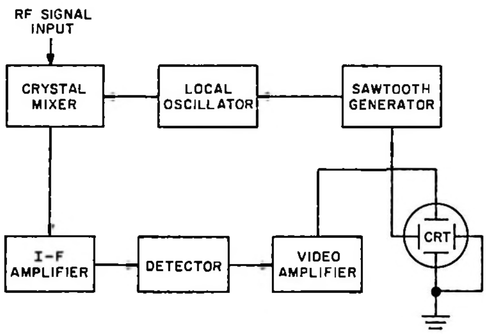


Figure 12-70. Heterodyne Spectrum Analyzer, Block Diagram

12-152. MICROWAVE SPECTRUM ANALYZER EQUIPMENT. Figure 12-71 illustrates a typical rf circuit used in microwave spectrum analyzers. The input rf signal to be analyzed is coupled to a waveguide-type transmission line which is terminated in a crystal mixer. The oscillator power is coupled to the main line through a directional coupler. The mixer output is directed through a coaxial line to the input i-f amplifier of the spectrum analyzer. The directional coupler isolates the local oscillator signal from the rf input, while directing this power to the mixer. The frequency meter is located in a position which permits the measurement of either the rf input sig-

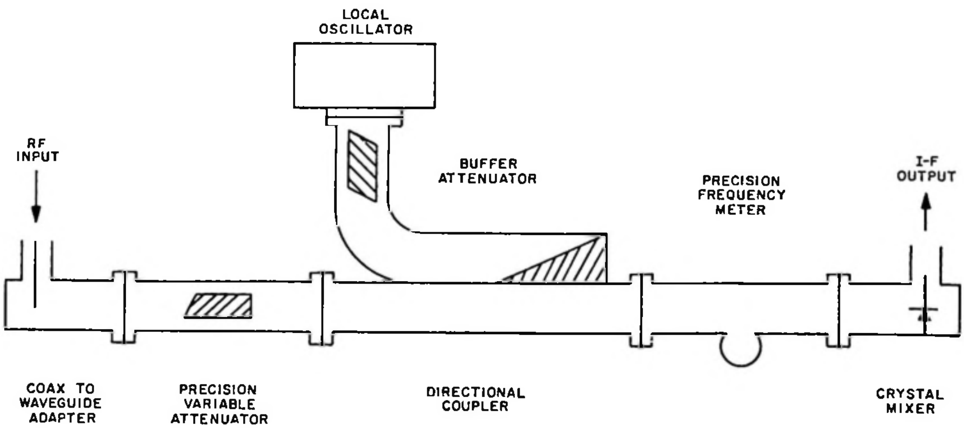


Figure 12-71. Typical RF Circuit Assembly Used in Microwave Spectrum Analyzer

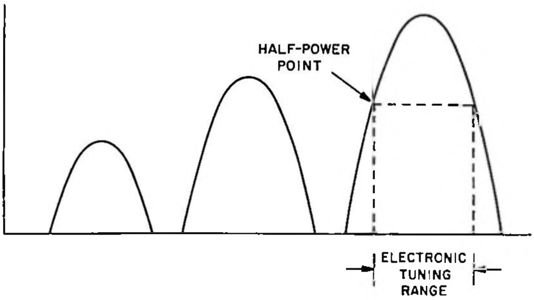


Figure 12-72. Electronic Tuning Range

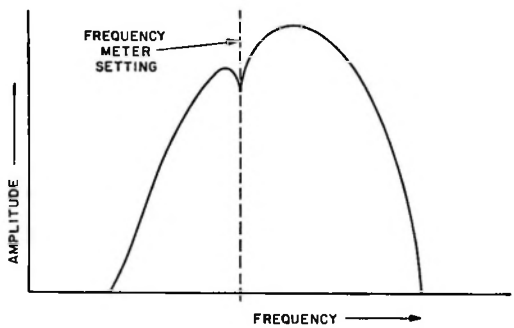


Figure 12-73. Dip in Amplitude of Oscillator Output Due to Frequency Meter Tuning

nal or the local oscillator power. The mixer is positioned where its matching properties can be controlled.

12-153. LOCAL OSCILLATOR. The local oscillator normally used is a reflex klystron. A sawtooth input voltage is applied to the reflector electrodes, to provide a frequency output which increases linearly with time. The dc voltage level of the sawtooth determines the oscillating mode to be used. The reflector voltage level and klystron cavity dimensional variations permit proper mode tracking throughout the klystron's operating frequency range. Setting the cavity dimensions determines the frequency range through which the klystron will sweep when motivated by the sawtooth sweep voltage. The total oscillator frequency sweep is known as the electronic tuning range. This range is primarily determined by the klystron mode width at the half-power points. The electronic tuning range is illustrated in figure 12-72.

12-154. BUFFER RF ATTENUATOR. The rf attenuator decreases the amplitude of the local oscillator pulse being measured, and thus prevents crystal burnout.

12-155. CRYSTAL MIXER. Normally, a silicon diode with a low shunt admittance at the operating frequency is used for the mixer. It is coaxially mounted to prevent leakage and is designed to bypass rf signals and to extract i-f signals. If no rf power is present, this crystal detector provides an input to a video amplifier which applies the mode shape to a cathode-ray tube to provide a visual screen display. This procedure provides a method of properly adjusting the local oscillator before performing the spectrum measurement.

12-156. FREQUENCY METER. The microwave frequency meter employs a coaxial cavity whose length determines the resonant frequency of the cavity. A pickup loop couples the meter to the main waveguide line to permit the extraction of a power sample. If the main waveguide line frequency is equal to the resonant frequency of the cavity, the cavity will absorb power from the main line. This power loss subtracts from the power being delivered to the crystal mixer. If you watch the analyzer screen, you will see a sharp dip in the power spectrum as the oscillator is swept through the cavity resonant-frequency point. This dip in the mode is sharp enough to provide exact frequency measurements, as illustrated in figure 12-73. Figure 12-74 illustrates a method of using the frequency meter to cause a dip in the pattern when the meter is tuned to either the local oscillator frequency or one of the components being analyzed.

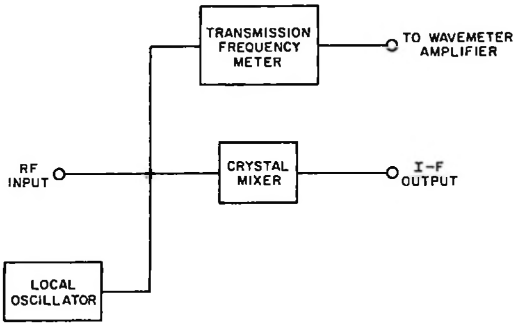


Figure 12-74. RF Circuit Using Transmission Frequency Meter

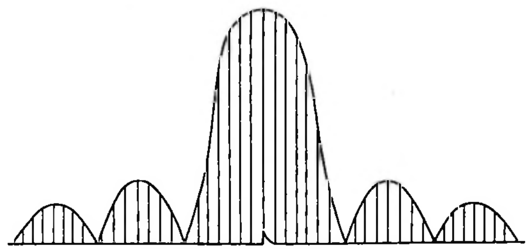


Figure 12-75. Local Oscillator Baseline Dip Caused by Frequency Meter Tuning

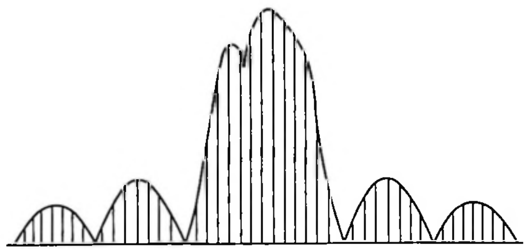


Figure 12-76. Signal Frequency Amplitude Dip Caused by Frequency Meter Tuning

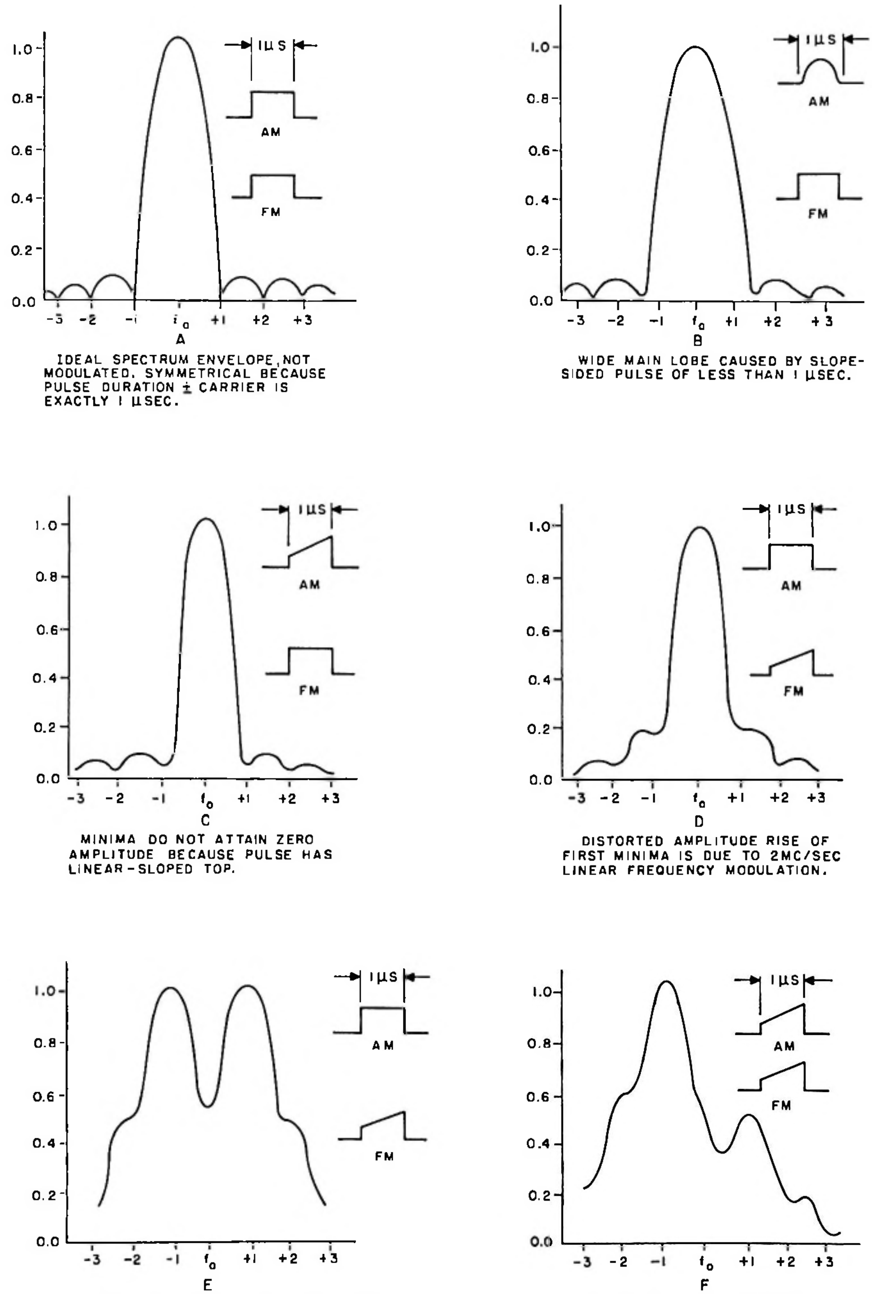
If the oscillator frequency mode is being dipped, the marker will appear on the frequency baseline, as shown in figure 12-75. If an input signal component is being dipped, the marker will appear in the amplitude of that particular spectrum component, as shown in figure 12-76. The frequency difference between any two points within a given spectrum can be measured by tuning the cavity from one point to the other and noting the frequency variation between them. Therefore, mode width and frequency measurements can be accomplished with the aid of an analyzer frequency meter.

12-157. ANALYZING THE SPECTRUM PATTERN. The primary reason for analyzing the spectrum is to determine the exact amount of amplitude and frequency modulation present. The amount of amplitude modulation determines the increase in the number of sidebands within the applied pulse spectrum, whereas an increase in frequency modulation increases the amplitude of the side lobe frequencies. In either case, the energy available to the main spectrum lobe is decreased.

12-158. TYPICAL SPECTRUM PATTERNS. Figure 12-77 contains several illustrations of commonly obtained patterns. These spectra are the result of pulse-modulated waves which are special types of rf carrier amplitude modulation. As you will see, amplitude modulation alone does not seriously affect the frequency spectrum of an rf pulse.

12-159. The type of modulation present can be easily determined because amplitude modulation primarily affects the amplitude of the side lobes and does not affect the shape of the main lobe. Frequency modulation affects the main lobe bandwidth. Spectrum asymmetry, as shown in part F of figure 12-77, occurs only when both amplitude and frequency modulation occur simultaneously.

12-160. The accuracy obtained in spectrum analyzer measurements is dependent on both the amplitude of the vertical pips and the



A — IDEAL SPECTRUM ENVELOPE, NOT MODULATED. SYMMETRICAL BECAUSE PULSE DURATION ± CARRIER IS EXACTLY 1 µSEC.

B — WIDE MAIN LOBE CAUSED BY SLOPE-SIDED PULSE OF LESS THAN 1 µSEC.

C — MINIMA DO NOT ATTAIN ZERO AMPLITUDE BECAUSE PULSE HAS LINEAR-SLOPED TOP.

D — DISTORTED AMPLITUDE RISE OF FIRST MINIMA IS DUE TO 2MC/SEC LINEAR FREQUENCY MODULATION.

E — DISTORTED AMPLITUDE RISE OF FIRST MINIMA IS DUE TO 6MC/SEC LINEAR FREQUENCY MODULATION.

F — DISTORTED TOTAL SPECTRUM IS DUE TO BOTH AMPLITUDE AND FREQUENCY LINEAR PULSE SLOPE. *

* CHANGING THE SLOPE OF ONE TYPE OF MODULATION WITH RESPECT TO THE OTHER WILL PRODUCE A MIRROR IMAGE.

Figure 12-77. Spectrum Patterns

power of the individual frequencies which these pips represent. The frequency meter you select will provide the primary limitation in your frequency measurements. The pattern you obtain is the result of heterodyning the signal with the local oscillator. This should provide linearity between the input signal and the difference-frequency output signal. This linear relationship between the input and output can be maintained by holding the amplitude of the local oscillator at a constant value as it sweeps through its assigned frequency range. The same result can be obtained by holding the local oscillator amplitude at a high value as compared with the signal amplitude at the crystal mixer input, if the mixer is operated in accordance with the square law. In addition to controlling the above functions, you must also have well regulated power supplies feeding the klystron.

12-161. The rf precision variable attenuator, shown in figure 12-71, is used to compare the power levels of different frequencies contained in the same spectrum pattern. It can also be used to determine the relative power levels of two or more incoming continuous-wave signals. To make these measurements, you must increase the attenuation until the signals containing the larger amplitudes are decreased to the original amplitudes of the smaller signals and note the exact attenuation required to perform this decrease. Since the normal attenuator is calibrated in db at the exact center of the analyzer frequency range of operation, its frequency sensitivity will produce some error. For example, the typical waveguide attenuator used in the 5000 to 6000-megacycle range will have a frequency sensitivity of approximately 3 db with an insertion loss of 50 db over its entire range.

12-162. The primary limitation imposed on the frequency range of a microwave spectrum analyzer is the local oscillator klystron. The limitation can be overcome by using two klystrons, one swept and one of fixed frequency, heterodyned in the crystal mixer. The difference frequency output from the crystal mixer represents a local oscillator operating at a lower frequency than either individual klystron. There is a sensitivity loss due to the double mixing action required, but wider-band analyzers are achieved in this manner because the percentage change in frequency is relatively large at the difference-frequency levels. In addition, stability requirements become greater because a small percentage shift at either klystron frequency represents a large percentage shift at the lower difference frequency.

12-163. The limitations on resolving power preclude the use of a conventional-type analyzer if you are attempting the analysis of a narrow-band microwave spectrum. However, even this limitation can be overcome if you mix the narrow-range signal with the output of a stable microwave oscillator operating at approximately the same frequency. These two very close frequencies will produce a beat in the audio frequency range. This audio beat can then be applied to an audio frequency analyzer, which can measure the relative power of each frequency. Figure 12-78 illustrates this method of using an audio analyzer. If you used an audio

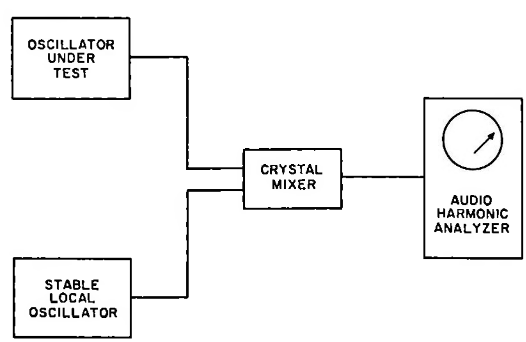


Figure 12-78. Typical Audio Harmonic Spectrum Analyzer Hookup

analyzer with a frequency range between 50 cycles and 16 kilocycles, the analyzer would achieve up to 40 db attenuation at an approximate bandwidth of $\pm$ 30 cycles about the center frequency. The resolving power would be well under 60 cycles, so that a 10-kilocycle signal could be resolved. However, the local oscillator must have negligible drift in comparison with the narrow frequency range being analyzed. This method is very accurate, but it has the disadvantage of being a point-to-point technique because you must manually tune the audio analyzer to obtain relative power readings at discrete frequencies.

12-164. TUNED-CIRCUIT SPECTRUM ANALYZERS.

12-165. RESONANT CAVITY. The resonant cavity is the most common tuned circuit in a spectrum analyzer for use in the microwave frequency range. A type of resonant cavity analyzer is illustrated in figure 12-79. The signal to be analyzed is coupled to the calibrated resonant cavity, and, when the cavity frequency is equal to the signal frequency, a reading is obtained on the meter. Point-by-point spectrum analysis is obtained as you record the meter readings for each different setting of the tunable cavity. As no sweep is involved, the resolving power is proportional to the Q of the cavity.

12-166. If you wish, you can make the tuned circuit spectrum analyzer automatic by using a motor to sweep the resonant cavity through its frequency range in synchronism with the horizontal sweep on an oscilloscope. The output of the detector is coupled to the vertical plates of the oscilloscope rather than to the frequency meter. You can adjust the phasing between the frequency setting of the cavity and the cathode-ray-tube scan voltage by indirectly adjusting the position of the electrostatic or electromagnetic field.

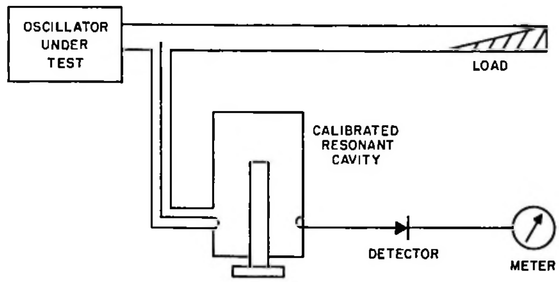


Figure 12-79. Typical Resonant Cavity Analyzer

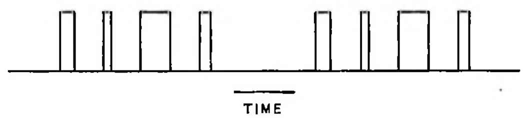


Figure 12-80. Random Width Constant Amplitude Coded Pulse Groups Used in Spectrum Analysis

12-167. MULTIPULSE ANALYSIS. Coded pulse groups, rather than a single repeated pulse, are coming into more common use in microwave transmission equipments. Figure 12-80 illustrates this type of random-width constant-amplitude signal. The spectrum analyzer used must be keyed to analyze the spectrum of a single pulse within this group. The typical circuit used as the keyer is shown in figure 12-81. The multipulse signal shown in figure 12-80 is heterodyned down to the first i-f frequency and applied to a crystal modulator and a video amplifier. The video signal is then applied to an oscilloscope for visual presentation of the entire group of random-width pulses. A particular pulse of interest is intensified by additional circuits, and this pulse is then used for analysis.

12-168. MATHEMATICAL AND GRAPHICAL SPECTRUM ANALYSIS. When a periodic waveform has an amplitude resulting from a function of time, its frequency spectrum

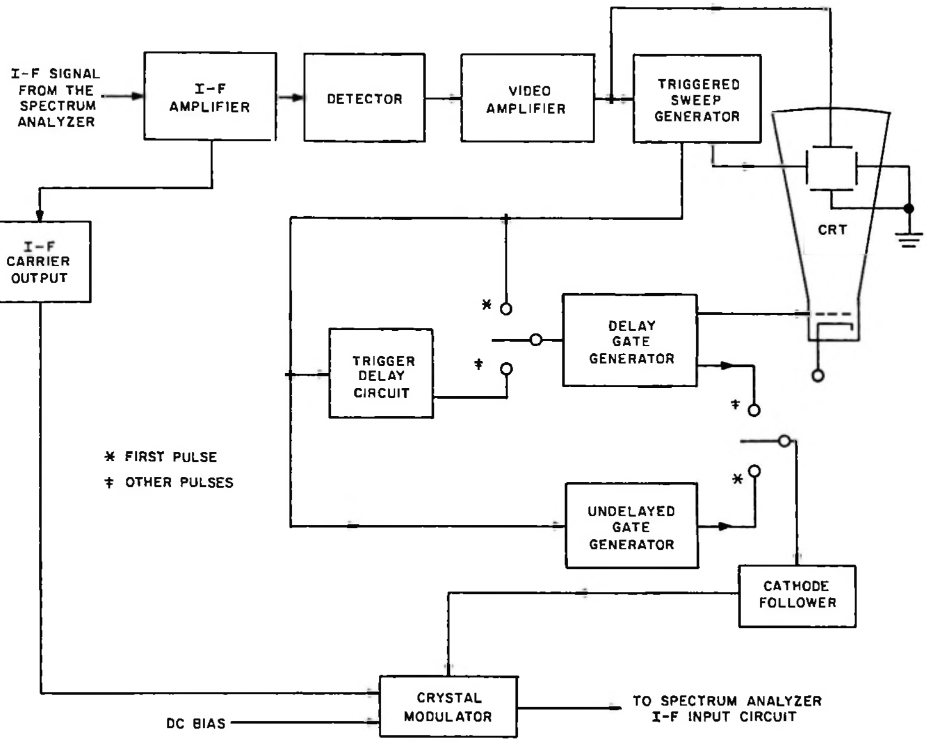


Figure 12-81. Multi-pulse Selector Circuit Used in Spectrum Analysis

consists of the fundamental frequency and the harmonics of that fundamental frequency. The analysis can be performed and graphed with the aid of advanced mathematics or by the use of a mechanical slide rule designed for this application. However, the problem of performing an analysis on a complex waveform may be simplified by using an electromechanical computer. These instruments automatically determine the amplitude of any individual frequency or combination of frequencies.

12-169. WIDE-BAND SPECTRUM ANALYZERS.

12-170. COMMUNICATION RECEIVER ANALYZER TECHNIQUE. A simple technique used to perform wide-band spectrum analysis at frequencies below 1000 megacycles is illustrated in figure 12-82. This technique employs a communications receiver to compare the unknown signal with a signal of known frequency and amplitude. While the unknown signal is applied, the receiver is tuned until a meter reading is obtained. The receiver input is then switched from the source of the unknown signal to the source of the known signal. The signal generator providing the known output is adjusted until the meter shows an output reading. You can read the generator's calibrated dial at this point to obtain the frequency of the unknown signal. Next, by adjusting the calibrated attenuator until the meter reading is identical to the reading obtained with the unknown signal applied, you will know the exact amplitude of the unknown signal. The foregoing will provide you with a point-to-

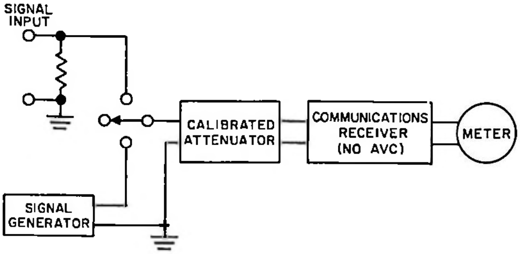


Figure 12-82. Single-Receiver Circuit Hookup for Wide-Range RF Spectrum Analysis

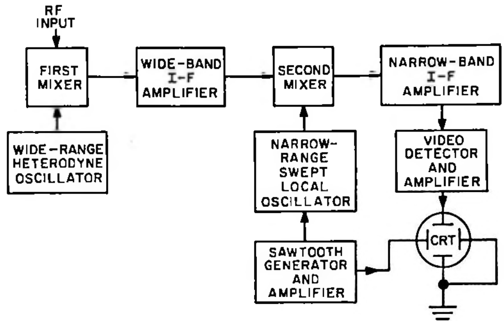


Figure 12-83. Double Heterodyne Circuit Used for Wide-Band Analysis

point spectrum analysis of any unknown signal.

12-171. DOUBLE-HETERODYNE CIRCUIT TECHNIQUE. A second technique involving a double-superheterodyne wide-range analyzer, for frequencies below 1000 megacycles, is illustrated in figure 12-83. In this technique a local manually tuned oscillator, providing the wide frequency range, is mixed with the unknown signal being analyzed. The two mixed signals provide difference frequencies, which are sent through an i-f amplifier and mixed with a swept local oscillator signal, which determines the spectrum bandwidth of the analyzer. The difference frequencies resulting from this second mixing process are sent through a narrow-band amplifier, after which they are detected, amplified, and applied to a cathode-ray tube for visual presentation. The output of the first i-f amplifier can also be electrically connected through an audio detector; earphones can be used to locate cw signals.

12-172. RESOLUTION AND ACCURACY. When a spectrum analysis depends on the sweeping of a local oscillator over a wide frequency range, the optimum resolution depends on the rate of frequency change within the local oscillator. If a frequency range on the order of 150 to 1 were to be covered in a single sweep, the sweep time would have to be exceptionally long or the resulution would suffer deterioration at the low end of the band. For example, if an audio-frequency analyzer were used to sweep between 100 and 15,000 cycles, the accuracy of frequency discrimination between 100 and 200 cycles is different from the accuracy between 14,000 and 15,000 cycles. This disadvantage could be eliminated by replacing the linear sweep with a logarithmic sweep rate. Thereafter, the physical separation between 100 and 500 cycles would be approximately the same as the spacing between 500 and 15,000 cycles. This would cause the frequency discriminator to remain constant over the entire sweep range.

12-173. Analyzer resolution is a function of both the sweep rate and the i-f bandwidth. The sweep rate must be a logarithmic function of time to maintain optimum resolution. These two facts make necessary the simultaneous variation of the i-f bandwidth as a function of the square root of the sweep rate. This can be accomplished by using a crystal filter as the input of the i-f amplifier's tuned circuit. Now, since the bandwidth is a function of the Q of the crystal, the internal resistance of a triode is placed in shunt with the crystal, and the triode grid

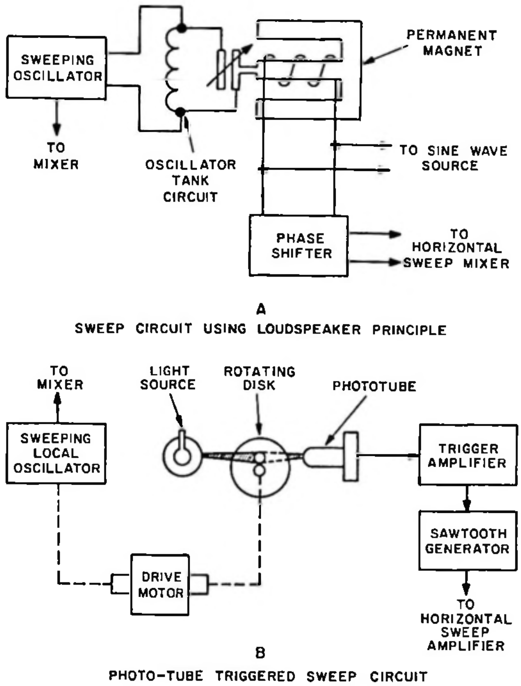


Figure 12-84. Sawtooth Sweep Circuits Used for Low-Frequency Spectrum Analyzers

voltage will vary the triode resistance with respect to time. This will, in turn, vary the selectivity of the crystal tuned circuit. The grid voltage performing this function is obtained from the same source which sweeps the local oscillator. Therefore, the bandwidth is varied in synchronization with the sweep rate of the analyzer. When the bandwidth is varied, the i-f amplifier gain varies. This may be compensated by adding another stage, with its gain swept in such a way as to create uniform gain over the entire frequency range of the analyzer.

12-174. SWEEP TECHNIQUES. When the input voltage of a tuned circuit is swept past the resonant frequency point, the frequency response of the tuned circuit will be distorted. This is true because the peak response of the tuned circuit has a tendency to sweep in the direction of the frequency sweep. The maximum sweep rate can be determined by using a triangular sweep voltage, which sweeps from the low through the high frequencies and then from the high through the low frequencies. If the pattern on the oscilloscope is displaced first to the right and then to the left in a double pattern, you must reduce the triangular sweep voltage until the patterns blend into one shape. This is the maximum permissible sweep rate.

12-175. In those spectrum analyzers which do not use the point-to-point technique, you must still provide synchronous sweeping of the local oscillator frequency and horizontal positioning of the crt beam. You may use the same sweep voltage for both functions when applied to a klystron oscillator in a microwave analyzer. However, below 1000 megacycles, where low sweep rates and wide frequency changes prevail, other methods of sweep must be employed. One such method uses a sawtooth generator to provide the crt horizontal drive and also the drive for a reactance tube circuit. The reactance tube circuit, in turn, sweeps the local oscillator frequency. Figures 12-84 and 12-85 illustrate other methods of obtaining a sweep voltage for application to the horizontal amplifier for the crt beam sweep.

12-176. Part A of figure 12-84 illustrates a sweep method which employs the electrodynamic loudspeaker principle. A voice coil equivalent varies the tuning capacitor spacing to vary the resonant frequency of the tank circuit and thus to sweep the local oscillator.

12-177. Part B of figure 12-84 illustrates a local oscillator motor which triggers the horizontal sweep voltage. However, the light source through the slotted aperture triggers only once each sweep; it does not maintain control after the sweep begins. Therefore, frequency accuracy

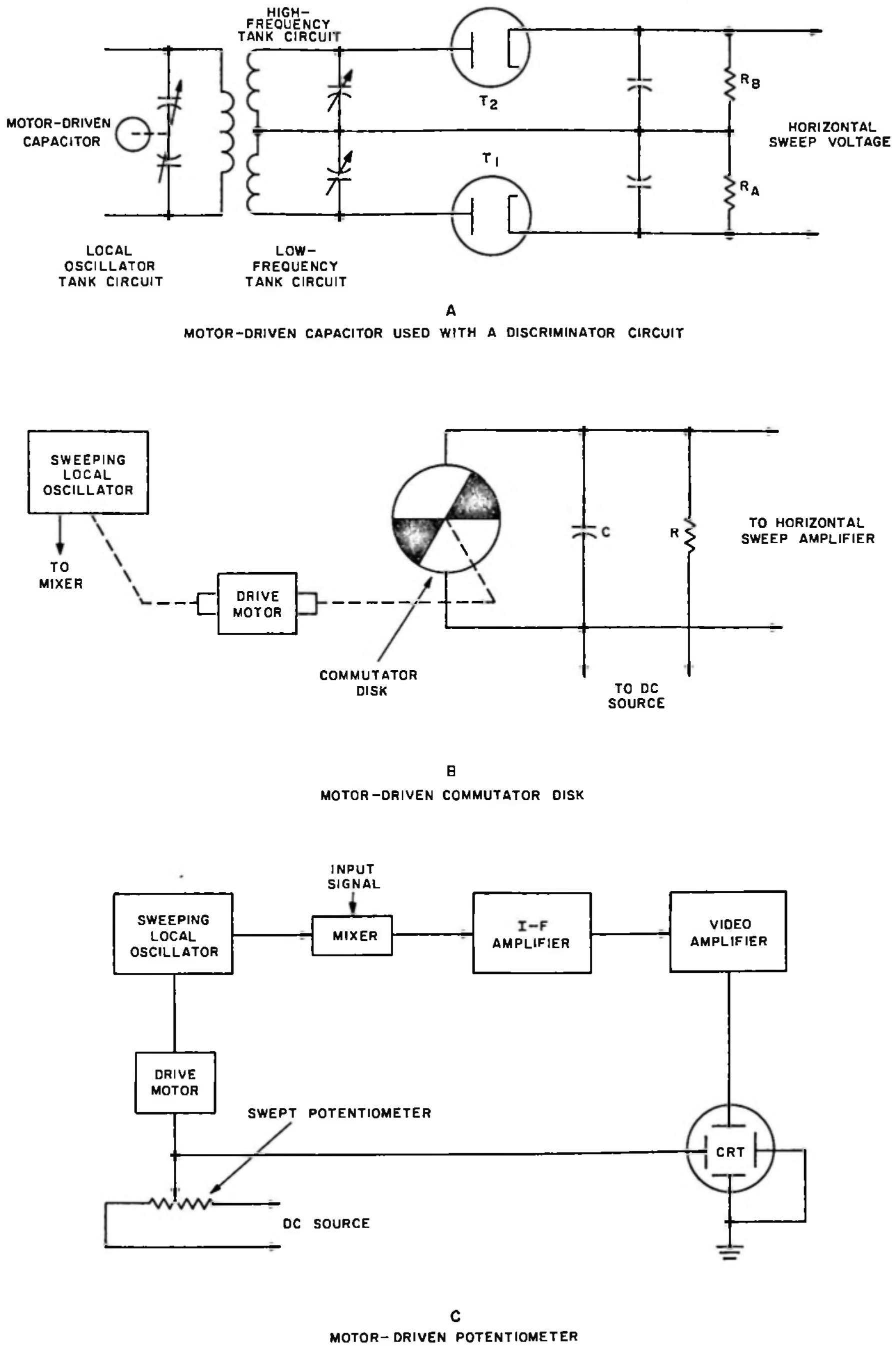


Figure 12-85. Motor-Driven Sawtooth Sweep Circuits Used for Low-Frequency Spectrum Analyzers

depends on the linearity of the oscillator sweep.

12-178. Part A of figure 12-85 uses motor-driven capacitors in the local oscillator tank circuit. A discriminator with two resonant circuits is connected to the tank output. The low resonant circuit is connected to the tank output. The low resonant circuit tube (T_1) conducts through R_A, and the high resonant circuit tube (T_2) conducts through R_B. In the center of the sweep range, the current through R_A and R_B cancels, and there is no output signal. Therefore, the output sweep voltage from the discriminator varies as a linear function of the local oscillator.

12-179. Part B of figure 12-85 illustrates the motor-driven oscillator discussed in paragraph 12-177, and illustrated in part B of figure 12-84, which is coupled to a potentiometer. When the potentiometer is connected across a dc source, a linearly varying output sweep voltage is obtained. This method is employed when high resolution and wide bandwidth are required. High resolution and wide bandwidth requirements are usually associated with low sweep frequencies.

12-180. Part C of figure 12-85 illustrates another disk technique which uses the charge and discharge of an RC network to obtain the sawtooth waveform for each disk revolution. The output sawtooth is applied to the crt deflecting plates directly if no amplification is required.

12-181. AUDIO-FREQUENCY SPECTRUM ANALYSIS. Since the bandwidth ratio in the audio-frequency range is 500 to 1, you are still primarily considering wide-band analyzer techniques. Audio-frequency spectrum analyzers may be generally grouped as follows:

a. Graphic
b. Resonant-circuit
c. Heterodyne
d. Stroboscopic

12-182. The previously discussed graphic analyzers, in common with all analyzers of this type, completely specify the frequency, amplitude, and phase of each spectral component contained in each waveform. Graphic analyzers are the only instruments which provide all three items in an analysis of the spectral components of a wave.

12-183. Resonant-circuit analyzers have the disadvantage of containing large, continuously variable coils and capacitors. However, these analyzers were used because heterodyne analyzers contained a fixed bandwidth and could be used on only one end of the swept frequency range. The resonant circuit analyzer uses the technique of passing the complex wave through a tuned circuit and obtaining the relative amplitude of each passed resonant frequency. The tuned circuit bandwidth depends on its Q; therefore, automatic variation of the selectivity takes place from one end of the tuning range to the other.

12-184. The disadvantage of large tuning components in resonant-circuit analyzers can be removed by using negative feedback for all frequencies except the one selected. This circuit has good stability, which is obtained from the effect of lumped constants.

12-185. HETERODYNE AUDIO ANALYZER. The heterodyne analyzer principle has been previously discussed in connection with microwave analyzers. The input wave is heterodyned with a high-frequency carrier oscillator which is tunable from 11 to 16 kilocycles, as illustrated in figure 12-86. The balanced modulator shown removes or reduces the carrier oscillator frequencies and any higher frequencies. The mechanical resonant filter uses a steel rod clamped at its center, to assure sufficient circuit Q to discriminate between the lower fre-

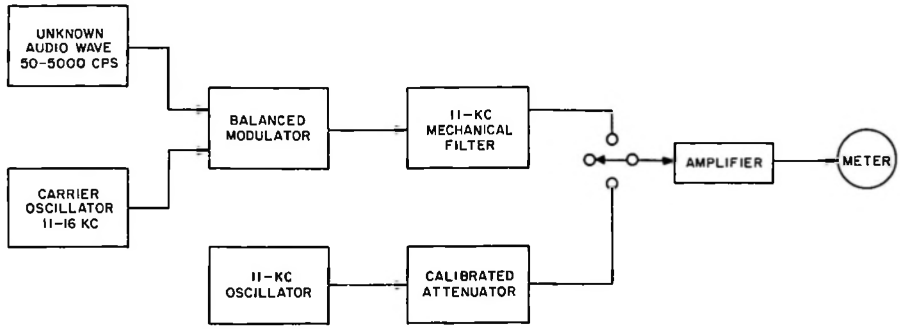


Figure 12-86. Voice Transmission Spectrum Analyzer

quencies. The steel rod provides coupling between the modulator on one side to the amplifier on the other. A Q of over 15,000 may be obtained in this manner, as compared with a Q of 200 obtained with lumped constants.

12-186. STROBOSCOPIC ANALYSIS. For low-frequency ranges, the stroboscopic analysis technique may be employed. The stroboscope frequency meter permits the measurement of frequencies in the range of 30 to 4000 cycles with an accuracy of 0.05 percent. The unknown frequency triggers a neon lamp that illuminates the stroboscopic disks from behind. Each disk rotates at different speeds, corresponding to primary frequencies within the range of the instrument. The disk speed is governed by a tuning-fork-controlled synchronized motor. At some constant speed of rotation, each disk represents a different frequency. When an individual frequency coincides with one of the circulating disks, that disk appears to stand still. Relative amplitudes are difficult to measure by this method. The analyzer frequency range can be shifted by changing the rotational speed of the disk.

12-187. The harmonic analysis of complex audio-frequency waves can be performed by any of the methods discussed in the past few paragraphs. Graphic and graphic-mechanical techniques provide an amplitude, frequency, and phase analysis of the individual frequency components of a wave. A heterodyne analyzer, using a calibrated local oscillator and balanced modulator, is normally used for sound and vibration analysis, provided that the phase relationships are not important.

12-188. One simple method of obtaining an analysis of a low-frequency signal is to gate an amplifier on and off with a square-wave input. If a sine wave of the same frequency as the gate signal is applied, output should consist of exactly one half of the input sine wave. However, which part of the input is amplified will depend on the phasing between the square wave and the sine wave. A dc meter, placed in the output of the amplifier, will read the average value for that portion of the sine wave that is gated through the amplifier. Therefore, the meter indication can vary between a maximum positive and a maximum negative value as the phase between the sine wave and the gating signal is changed. When this method is used, an unknown signal is applied to the amplifier and a known square-wave signal is applied to the amplifier gating circuit; the frequency of the square wave is then varied through the low-frequency spectrum. Each time the square wave frequency matches a basic component of the unknown signal,

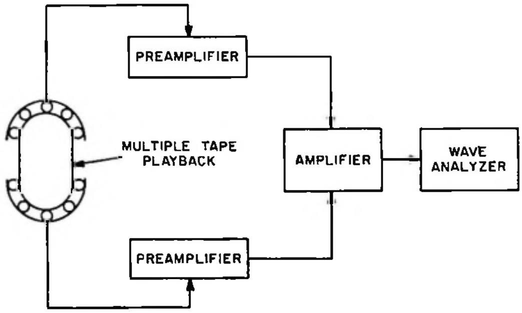


Figure 12-87. Audio Spectrum Analyzer

maximum amplifier output will be shown on the meter.

12-189. VERY-LOW-FREQUENCY AND TRANSIENT ANALYZERS.

12-190. TRANSIENT SIGNALS. Normally, a steady pattern is obtained on the screen from a conventional heterodyne analyzer which is used to measure the spectrum of a steady-state, non-changing input signal. However, if the signal changes during analysis, the crt screen pattern will obviously change also.

12-191. If the input signal changes are slow in comparison with the local oscillator sweep signal, the screen will reflect these changes in the pattern. The screen persistence characteristic will determine both the clearness and the distinctness of these signals. When the change becomes more rapid, and increases beyond some maximum frequency point, the pattern becomes too distorted and blurred for analysis of the wave. Considering that the analyzer sweep rate cannot be increased without sacrificing the resolution, the analysis of rapid transient signals is difficult.

12-192. An analysis of a transient signal in the audio range can be accomplished by recording the rapid-change signal on tape and splicing the tape end-to-end as a continuous repeating circle. This circular looped tape is then placed on a playback machine, which repeats the signal over and over. The number of scans required to obtain an adequate spectrum analysis depends on both the transient spectrum width and the sweep range of the analyzer. However, if you make a long-time photograph of the crt screen pattern, you can obtain an average spectrum plot of the transient response.

12-193. COMPOSITE SIGNAL ANALYZER. The composite signal audio analyzer uses the circular looped tape technique. Adequate speech spectra can be obtained by using this method. Speech, as such, does not lend itself to spectrum analysis. However, the speech pattern of an individual can be analyzed from a 2-minute random reading sample. Figure 12-87 illustrates a method of analyzing a speech signal.

12-194. The basic heterodyne analyzer has a limited sweep rate for a given resolution because of the delay created by the i-f filter buildup time. However, by using a high-speed resolution analyzer you can circumvent this sweep rate limitation. In this method, the subject frequency range is applied simultaneously to a group of filters, and each filter responds to a different basic frequency of the input spectrum. The filter outputs are detected in series, amplified, and applied to the vertical deflection plates of the oscilloscope.

12-195. SUB-SONIC SPECTRUM ANALYZERS. The average audio analyzer is designed to operate effectively down to 50 cycles per second. However, there are many instances where lower frequencies must be analyzed. For example, in the study of structural vibration, geophysical investigation, and medical electronics, you may be required to spectrum-analyze frequencies down to one cycle per second. Vibration analysis of a steel bar disclosed a

125-cycle resonance condition and a total bandwidth of only one-half cycle. The subsonic analyzer requires a filter output to operate a recording pen. A motor is used to synchronize the local oscillator with the feeding of the paper past the recording pen.

SECTION VI

NOISE FIGURE MEASUREMENTS

12-196. GENERAL.

12-197. NOISE SOURCES.

12-198. The useful operating range of all radio and much of the higher frequency equipment is limited by noise interference. Atmospheric, man-made, and receiver, and antenna noise are the three primary sources of limiting noise interference. These types of noise are shown in graphical form in figure 12-88 for your convenience.

12-199. ATMOSPHERIC NOISE. This type of noise is dependent on frequency, time, weather, season, and location. This is true because most atmospheric noises are created by lightning discharges occurring in local thunderstorms. Noise usually decreases with increasing latitude on the surface of the globe. Therefore, noise will be severe during the rainy seasons in the Caribbean, East Indies, Equatorial Africa, Northern India, etc. The median values of atmospheric noise for our nation or other regions lying in the same median (30 to 50 degrees north or south latitude) are illustrated in figure 12-88. The curves of figure 12-88 assume a bandwidth of 10 kilocycles. The curves also assume the use of a half-wave dipole antenna and no transmission line losses. Atmospheric noise is the primary limitation on the transmission of intelligence at frequencies below 30 megacycles. Above 30 megacycles, atmospheric noises are minor compared to receiver noises. The peak value of interfering at-

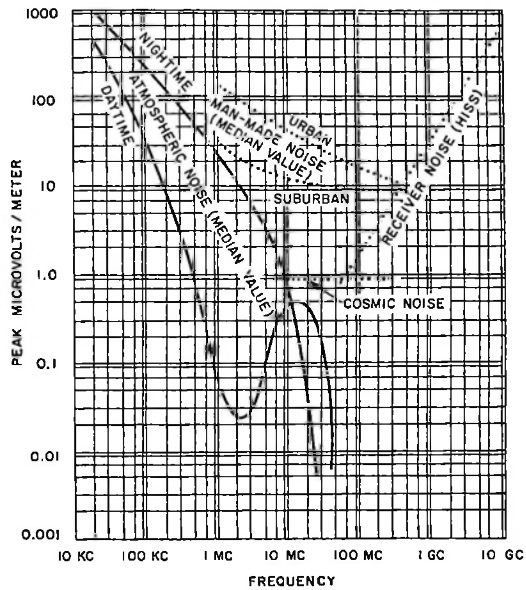


Figure 12-88. Radio-Frequency Noises Applicable to Regions Between 30 and 50 Degrees Latitude North or South

mospheric noise is normally directly proportional to the square root of the receiver bandwidth. The values provided in figure 12-88 may be multiplied by the factors in table 12-3 to obtain the atmospheric noise factors in regions other than those provided for the United States.

12-200. COSMIC NOISE. Cosmic noise interference normally has a lower peak volt-

Table 12-3. Atmospheric Noise Multiplication Factors for Regions Other Than the United States

DEGREES OF LATITUDE	NIGHTTIME		DAYTIME	
	100 KC	10 MC	100 KC	10 MC
90-50	0.1	0.3	0.05	0.1
50-30	1	1	1	1
30-10	2	2	3	2
10- 0	5	4	6	3

age value than any other type of radio interference. However, if atmospheric or man-made noises are not present or are extremely low, cosmic noise will become the primary limiting factor within the frequency range of 10 through 300 megacycles. Cosmic noise is received from three distinct sources: (a) galaxy noise, which has the same characteristics as thermal noise in the primary region between 150 and 200 megacycles (this interference apparently comes from the general direction of the Milky Way and has a spatial distribution); (b) thermal noise due to celestial bodies appears between 3 and 30 gigacycles; and (c) anomalous solar radiation, which appears around the frequency of 30 megacycles and is dependent on sunspot type disturbances.

12-201. Atmospheric noises are sometimes called "static" because of their peculiar sound. As has been previously stated, static noise exerts the most disturbance during thunderstorms, which occur more during the six warmer months of the year than they do during the six colder months.

12-202. The only static limiting methods available at this writing are as follows:

a. Insert limiter stages within the audio section of the receiver to limit the amplitude of all static signals above a set amplitude level and thus limit the volume crashes resulting from lightning bursts to approximately the same volume as the desired signal.

b. Limit the reception bandwidth. Since most atmospheric noises are spread over a broad frequency range, a narrow reception bandwidth will eliminate the noise distrubuted outside this bandwidth.

12-203. MAN-MADE NOISE. This type of noise is most prevalent around densely populated or industrial areas. It comes from such interference sources as automobile ignitions, electric motors, electric switching devices, high-tension line leakage, diathermy machines, heating generators, and a host of other man-made or man-permitted electrical disturbances. Figure 12-89 is a bandwidth factor chart which provides a multiplication factor for man-made interference in receiver bandwidths greater than 10 kilocycles. This factor should be multiplied by the value obtained in figure 12-88 for data beyond the scope of the bandwidth represented in this figure.

12-204. Man-made noises are so variable in both source and bandwidth that a simple rule to convert 10-kilocycle bandwidth noise measurements to other bandwidth consid-

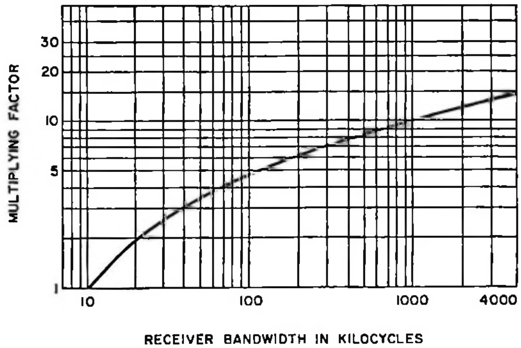


Figure 12-89. Noise Multiplication Factors for Man-Made Noise in Bandwidths Greater than 10 Kilocycles

erations is not available. The typical median multiplication factors provided in figure 12-89 simply mean that approximately 50 percent of all receiver sites must have lower noise values than those represented by these curves.

12-205. THERMAL NOISE AND TUBE HISS. Even in the complete absence of atmospheric and man-made noises, there still remains a practical limit to the maximum value of the weakest signal that can be used as intelligence. The primary limitation is due to either thermal noise or tube hiss.

12-206. Thermal noise is the result of the friction heat created by electrons within a material. Generally speaking, the greater the degree of applied heat, the greater number of affected electrons and, consequently, the more friction heat created. Thermal noise has a uniform power distribution throughout the radio-frequency spectrum. This distribution factor distinguishes true thermal noise from the galaxy noise explained in paragraph 12-200.

12-207. Noises produced within a tube are usually due to the fact that electrons emitted from the cathode hit other electrons or tube elements in the electron's normal cathode-to-plate transit path. Naturally, any hitting, bumping, or other process of slowing down of electron speed will create plate current fluctuations which will be reflected in the circuit output as undesirable noise. The maximum value of this tube noise limits the minimum value of voltage, representing applied intelligence, that can be amplified to provide an output discernible as the original applied intelligence. Therefore, the majority of tube noises are created by shot effect (electrons colliding with each other or with tube elements). In multielement tubes (those containing more than a plate and cathode), additional tube noise will occur because of fluctuations created by the division of current between the different electrodes. For example, low-noise triode amplifier tubes have a noise resistance on the order of 200 ohms; low-noise pentode amplifier tubes have a noise resistance on the order of 700 ohms (more tube elements); pentode mixer tubes have a noise resistance on the order of 3000 ohms; and frequency-converter tubes have a noise resistance on the order of 200, 000 ohms. These noise resistances are considered as additional resistances appearing in series with the internal plate resistance of the tube.

12-208. RANDOM NOISE. Random noise is created as the result of many frequencies (unrelated in phase) beating with the carrier or with each other, to provide difference and sum frequencies not contained in the original intelligence. These newly created frequencies produce audible noise as a result of being demodulated at the detector or discriminator. Figure 12-90 illustrates the comparison between a-m and fm audible noise.

12-209. NOISE AND THE DEVIATION RATIO. The deviation ratio is a term representing the ratio of the maximum carrier swing to the highest audio-frequency value.

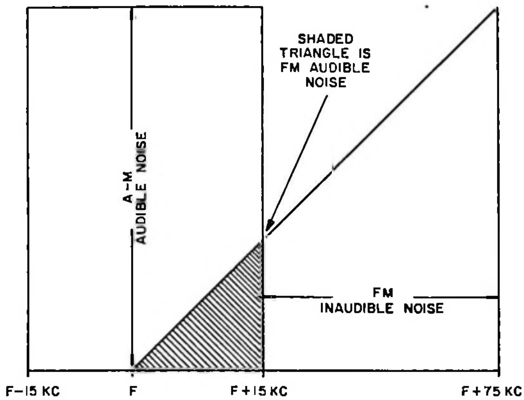


Figure 12-90. A-M Versus FM Audible Noise Comparison

For example, for a carrier shift of 45 kilocycles and an audio frequency of 15 kilocycles, the deviation ratio would be 3. However, the signal-to-noise ratio varies in direct proportion to the deviation ratio. Therefore, a small deviation ratio will result in a small signal-to-noise ratio; consequently, more noise will be present. From the foregoing discussion, it is obvious that the deviation ratio should be as high as possible to produce the minimum noise level in the output signal.

12-210. PRE-EMPHASIS AND DE-EMPHASIS. The majority of noise energy is contained in the lower frequencies, and the greatest amount of irritating noise is generated in the audio range above 5000 cycles. A pre-emphasis network may be inserted within the audio section of the transmitter to reduce the unwanted noise voltages. The pre-emphasis network attenuates the lower frequencies, where most of the noise is contained, more than it attenuates the higher frequencies. When the proportionately attenuated signal reaches the receiver, a de-emphasis network inserted within the audio section of the receiver will reverse the effects of the pre-emphasis network contained in the transmitter. Therefore, the transmitted signals are returned to their proportionate original values. However, as the noise was lost through pre-emphasis, it will not be regenerated and injected into the de-emphasized signal.

12-211. IMPULSE NOISE. As the name implies, impulse noise consists of sharp, short bursts of noise energy created by either internal or external agencies, and may be the result of either man-made or natural causes. Typical examples of this type of noise are the familiar sparking generated by electrical machines or lightning bursts occurring during a local thunderstorm. The primary method of removing this type of noise is to employ a limiter stage within the receiver to limit or cut off the noise peaks that extend above the peaks of the signal containing intelligence.

12-212. HUM. The type of noise called hum is normally the result of insufficient filtering of the 60- or 120-cycle line frequency used to produce the pure direct current required as the equipment operating power. Hum may also result from insufficient choke or transformer shielding. The normal method of decreasing this type of noise is to use a capacitor to bypass all dc connecting points and to provide adequate shielding for the chokes and transformers contained within the equipment power supply. Wires and lines containing low frequencies should be positioned as far apart as possible and laid at right angles to each other, to prevent feed-over from one wire to another and consequent modulation of the signal in an adjacent wire or line. The total amplitude of this undesirable modulation would be converted into hum noise.

12-213. MEASUREMENT OF NOISE.

12-214. INTERNAL NOISE. Static and noise voltages are difficult to measure because of their varied characteristics. A continuous noise, such as a hiss or hum, can be measured in terms of the field strength of a just audible tone of known amplitude and frequency. However, the measurement of all types of noise is based on the comparison of the noise signal with a signal of known characteristics, and this method generally results in a close approximation rather than an exact result.

12-215. EXTERNAL NOISE. Atmospheric, cosmic, and man-made noises can be measured by the comparison method discussed in paragraph 12-214. However, since internal noises consist primarily of steady voltages, and external noises consist primarily of sharp, short bursts of energy, it is necessary to measure the average internal noise and the peak external noise pulses. Average values can be measured by an audible method, but pulse noise should be measured visually with the aid of an oscilloscope. Automatic recorders should be employed to aid the measurement of noise signals occurring over a wide time limit.

12-216. NOISE FIGURE.

12-217. GENERAL.

12-218. A more precise evaluation than audible or visual comparisons of the noise emanating from an equipment can be obtained by means of the noise figure. The term noise figure represents the results of a measurement of the noisiness of an equipment. The noise figure is the ratio of the noise power output from an equipment to a noiseless output from the same equipment (not counting the noise originating from the source generator applied to both the noisy and noiseless equipment). The noise figure principle can be applied to any type of electrical noise device. However, this section deals with the representative area of its widest use--radio transmission and reception. The noise figure is actually a useful, but restricted, figure of merit. All other factors being equal, the quality of an equipment can be determined. However, an equipment required for broadband applications may have a greater noise figure than a narrow band equipment, and yet be better qualified for its application than any other equipment.

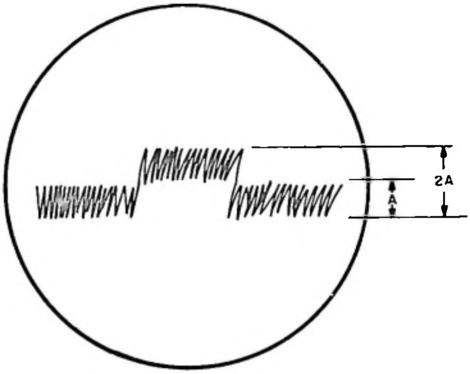


Figure 12-91. Tangential Signal

12-219. The noise figure measurement method can be used to determine the quality of an equipment between its input terminals and some other later stage of the equipment. However, other sensitivity measurements have been, and are still being, used to determine the quality of an equipment with respect to its undesirable noise content.

12-220. TANGENTIAL SIGNAL. The term tangential signal is sometimes used to define the quality of an equipment with respect to noise. As shown in figure 12-91, the tangential signal is defined as the result of raising the original noise level to twice its former amplitude. The tangential signal required to double the amplitude of the original noise voltage has been determined, by experimental measurements, to be approximately 8 db.

12-221. MINIMUM DISCERNIBLE SIGNAL. A pulsed signal which is just barely discernible from the surrounding noise signals

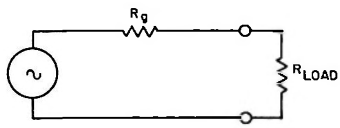


Figure 12-92. Power Network Match or Mismatch

on an "A" type oscilloscope is termed the minimum discernible signal (mds).

12-222. When using the sensitivity terms defined in the two paragraphs above, you must realize that both the tangential signal and the mds signal will provide only a close approximation. This is true because there is no well defined line that you can consider to be the prevailing noise level; thus your measurement results may vary because of difficulty in distinguishing the reduced mds signal amplitude from the prevailing noise level. Also, depending on the method and type of presentation, signal pulse width, bandwidth, sweep speed used during measurement, etc, the measurement result will vary.

12-223. For higher frequencies (above the 10 megacycle point), the received noise signal is added to the noise normally generated within the first few stages of the equipment and this result may be compared with the received signal. However, following amplification by the beginning stages of an equipment, you can compare the original input signal with the amplified signal and determine the approximate amount of noise generated within the equipment being used. You must realize that the input noise will be amplified along with the normal signal it accompanies, and therefore, noise generated in later circuits cannot compete with the large volume of amplified signal noise. If you can decrease the input signal-to-noise level and thereby increase the equipment reception qualities, the equipment can make better use of the information received at its input terminals. The noise figure is an objective measure of receiver sensitivity, and, since it eliminates human error (and other variance factors), it has become the primary method of determining equipment quality with respect to noise.

12-224. THEORY.

12-225. POWER. Consider the circuit of figure 12-92; when the value of R equals the value of R_g, maximum power will be delivered to the load. In any case, the power can be calculated by the application of Ohm's law formulas. The power delivered to the load will always be less than the power available at the input. If R and R_g are not equal, the mismatch loss will be even greater, and less of the input power will be available at the load. The available power at the load is independent of the actual impedance value of the load. Only the fact of match or mismatch between impedances (R and R_g) is important. A circuit naturally contains less noise under the conditions of mismatch because some of the noise, along with the signal, is lost.

12-226. GAIN. The gain of a stage is simply the ratio of power output to power input. If, as shown in figure 12-92, P_1 is the power input and P_2 is the power output, then the gain (A) equals P_2/P_1. This simple fact holds true for cascaded stages as well. In fact, you can break the cascaded circuits at some point and use the same simple calculations to determine the gain between the input and point of breakage. As was stated previously, the gain of a stage or cascaded stages is independent of the load impedance. However, the gain is dependent on the stage input impedance because the signal is developed across these terminals. The gain calculated or measured at the output of the first stage can be multiplied by the gain of the second, third, fourth, etc, to obtain the total gain of the equipment. Therefore, the total gain of an equipment is the product of the gains of all stages.

12-227. Knowledge of both the gain and power available within an equipment will permit you to formulate the noise figure into definite measurable terms. The following formula can be used to calculate the noise figure:

$$NF = \frac{P_{s1}P_{n1}}{P_{s2}P_{n2}}$$

where:

NF = Noise figure (when the source impedance temperature is 290 degrees K (17 degrees C or 62.6 degrees F)

P_{s1} = Signal power input

P_{s2} = Signal power output

P_{n1} = Noise power input

P_{n2} = Noise power output

Other pertinent facts of noise power should be realized. In the case of a metallic load resistor, the thermal noise output (P_n) is equal to KTB, where K is Boltzmann's constant (1.374 x 10^{-23} joule per ^{o}K), T is the absolute temperature of the resistor in degrees Kelvin, and B is the bandwidth over which the noise is observed. A carbon resistor generates an additional noise output when direct current flows through it. This noise output is proportional to the square of the direct current through the resistor, and may be far in excess of the thermal noise. The total available output noise power (P_{n2}) obtained when the noise figure is derived from the output of a stage being fed from a resistive source is $NFAKT_oB$, where each letter symbol except T_o represents the same terms previously discussed; T_o is an IRE standard and equals 290 degrees Kelvin (17 degrees centigrade or 62.6 degrees Fahrenheit). In fact, the noise power output (P_{n2}) is approximately 4 times 10^{-15} for this exact temperature, and, unless an exact noise figure is required, no calculations are necessary. However, you must remember that the noise figure is dependent on an exact temperature and these conditions are rarely obtainable. The true operating noise figure (NF_o) depends on the calculation of source temperature, etc, and is rarely usable in the field. Therefore, NF is normally used when comparing the noise figures of two separate equipments. In addition, the foregoing formulas assume equal bandwidths in all stages of the equipment, and, as you know, this is not necessarily true. In the event that all bandwidths of all stages are not equal, you must make calculations based on the available criteria contained within each separate stage. Remember, all stages within the equipment must be linear or the foregoing formulas, approximations, etc, are not applicable.

12-228. BANDWIDTH. The previous discussion assumed that the over-all bandwidth (B) of the equipment was the same as the usable bandwidth (B_1), and this is not always true. When B is larger than B_1, this "extra bandwidth" will permit an increase in noise power, but the original signal (without noise voltages) will remain the same. In some cases, the useful bandwidth has been restricted deliberately so as to permit the amplification of only part of the available signal. In this case, an increase in bandwidth will permit amplification of more of the original signal. But, an increase in bandwidth will permit a greater increase in the noise power signal than in the desired signal. Spurious response within an equipment represents an increase in bandwidth. The elimination of spurious response will improve the noise figure, as shown by figure 12-93. NF, shown in the figure, is the noise figure including spurious response, and NF_1 is the noise figure without the

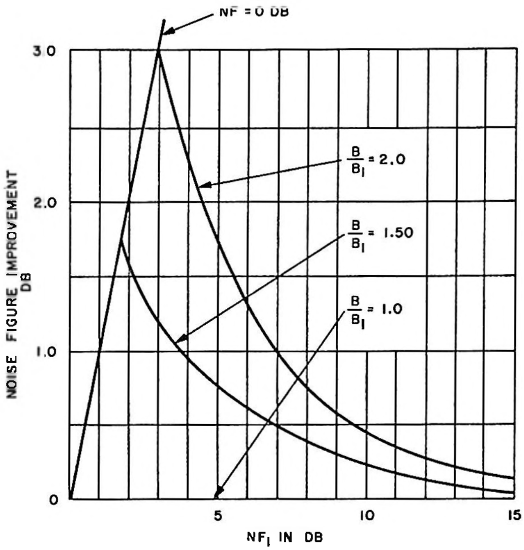


Figure 12-93. Noise Figure Improvement Chart

spurious response present. As illustrated in the figure, as the ratio B/B_1 increases, the noise figure improves and less noise is apparent within the equipment.

12-229. MEASUREMENT REQUIREMENTS.

12-230. You must apply a power signal of known characteristics to the input of the equipment to be measured for its quality of signal reproduction versus noise. You must next compare the equipment signal output, obtained as a result of the normal input, with the signal output obtained when using a signal source of known characteristics. You are actually obtaining a resultant from two separate measurements--the output noise factor with normal signal (P_2) and the output noise signal with an artificial input signal (P_3). Therefore, $\frac{P_3}{P_2}$ represents P, which is sometimes termed the Y factor. This P or Y factor represents a single calculation figure which can be used

Figure 12-94. Noise Figure Measurements, Basic Equipment Arrangement

to determine the noise figure. All of the different measurement techniques are merely variations of this method. Figure 12-94 illustrates the basic arrangement of equipment used to measure noise. The input noise source shown in figure 12-94 can be a signal or noise generator with either a constant or variable output. The output noise power measurement can be accomplished with a combination linear amplifier and quadratic detector. A linear amplifier is required because a diode operating in the square-law region of its characteristic curve can be used for input signals above a relatively low level to measure output noise power, but for low-level signals, noise generated within the diode will cause errors. A crystal detector operating in the square-law region can be used to measure noise output, but its dynamic range is limited. Bolometers, thermistor, or thermocouple meters are usually the best type of meters to use for the measurement of noise. However, thermal meters have a tendency to drift with operational time; therefore, continual zero-setting and recalibration is necessary. You can use the thermionic or crystal detector in a non-square-law region and calibrate the detector from a standard signal source before each use. Non-thermal detectors operating outside of the square law region will react differently when a noise signal is added to a noise signal than when a noise signal is added to a sine wave signal. The rectified detector current will change from a factor of 1.414 for noise plus noise to a factor of 1.45 for noise plus signal.

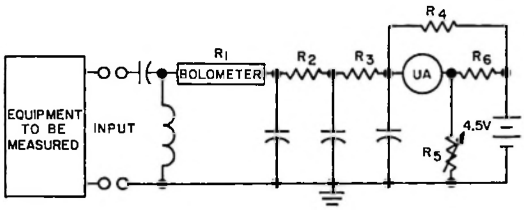


Figure 12-95. Typical Bolometer Bridge Detector Arrangement for Measurement of Noise Power

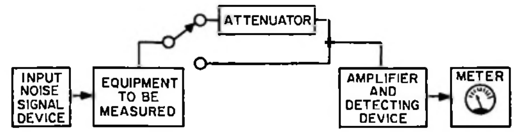


Figure 12-96. Attenuator Method of Noise Figure Measurement

12-231. The output indication can be the result of directly measuring the power level at different points within the measured equipment. The noise signal can also be measured by using a power level monitor in conjunction with an attenuator to establish equal power levels at the monitor equipment for both the initial and final measurements.

12-232. TYPICAL EQUIPMENT ARRANGEMENT FOR NOISE FIGURE MEASUREMENT.

12-233. TYPICAL BOLOMETER BRIDGE DETECTOR. Figure 12-95 illustrates a typical bolometer bridge method of measuring the noise power output of an equipment. The bolometer (R_1), together with R_2 and R_3, forms one arm of a Wheatstone bridge. The other arms are formed by R_4, R_5, and R_6. The bridge is balanced by adjusting R_5 while the input of the detector is disconnected. When the detector is connected to the equipment to be tested, the input noise power causes the bolometer resistance (R_1) to change value. The changing value of R_1 will upset the bridge balance, and, if the equipment being measured is linear, the output test meter reading will be proportional to the noise power output.

12-234. OUTPUT INDICATOR DEVICE. As shown in figure 12-96, if an output device is used as a detector only, an attenuator must be placed at a point following the input-noise-contributing elements and ahead of the point where equipment linearity becomes a factor. Connect the equipment as shown in figure 12-96, and set the indicating device following the detector to a value representing the quiescent condition of the equipment to be measured. Next, apply the signal to the equipment under measurement. The noise power will be that amount of attenuation required to reduce the indication device back to the value set for the quiescent noise. If the device reacts the same to the result of noise signal plus noise signal as it does to noise signal plus sine wave signal, you can make both signal generator and noise generator type measurements with the same set-up.

12-235. The attenuator method of noise measurement illustrated in figure 12-96 uses a low-impedance coaxial cable connection between the attenuator-amplifier and the detecting device. Normally a 75-ohm load is used to help create a broad bandwidth within the measuring device; unfortunately, a loss in gain will be experienced. The amplifier just before the detecting device compensates for this gain loss. The detecting device should be well shielded and contain its own power supply, to prevent feedback to the equipment under test.

12-236. EQUIPMENT VERSUS BANDWIDTH. The actual measuring equipment should not affect the equipment being tested. This means that the signal source must be properly terminated not only for the desired signal channel but also for the spurious channels which provide image and harmonic response. For example, if the harmonic response channels are not properly terminated, conversion losses may result.

12-237. ERRORS. The measurement of noise figure requires that the error be minimized. Therefore, you must take your measurements at a point within the equipment where the signal level has been established and beyond which no significant noise contribution normally occurs. No nonlinearities from preceding stages within the test equipment can be permitted. Considering that noise signals are not linear and that noise peaks may exceed the rms level, the linearity of the equipment being measured must be maintained up to an amplitude of at least four times the rms value of the noise power. Therefore, direct noise power output measurements have limited application. The signal source used in noise figure measurements should have a fixed value of output power with a ratio (P), as discussed in paragraph 12-230, of at least 2 and preferably 10. In addition, if the input signal is provided by a "white noise generator," the ratio represented by **P** should be at least 40. If the ratio is as high as 40 and a diode or crystal detector is used to measure the power level, then the detector must follow the square law or linear law over a 16-db range, or have a known relationship between the input and output levels for long periods of time.

12-238. In accordance with the law of output measuring devices, you must know the detailed output characteristics of your output indicator. For example, you must be able to obtain a response as a result of detecting a noise-plus-noise-signal mixture or a noise-plus-sine-wave-signal mixture. If you do not use the constant output measurement method, you must know the output indicator characteristics, such as stability with respect to time and calibration accuracy.

12-239. Other factors which create measurement errors include useful channel bandwidth, assumed value of generator output, and source impedance. An error would result if you used an incorrect value of useful channel bandwidth B_1 of the equipment being tested or an improperly corrected output power from the signal source in cases where B/B_1 is not equal to 1. The output from the source signal generator may not be the value stamped on the equipment. This change in output could have been caused by gassy tubes, worn out components, attenuation not calibrated recently, or many other factors. Therefore, never assume that the stated output is correct. Always check the source generator output to assure that it is correct and that the output meets your requirements. The source generator must also have the proper impedance in both its on and off conditions. The self-evident fact that an equipment may be specifically designed to operate from a 50-ohm-impedance source requires that you provide a source having this exact impedance. Unfortunately, failure to obtain exact impedance matching is the most common cause of measurement error. Even though the impedances concerned are resistive, you cannot correct the mismatch with a simple computation because the noise figure may be critically sensitive to source impedance values. The corrective procedures become even more complex if the impedances concerned are reactive rather than resistive.

12-240. In general, the most precise measurement technique with a minimum of error requires the use of a precision variable i-f attenuator in conjunction with a constant-output type indicator. The accuracy of measurement should be as precise as the attenuator calibration. Considering that secondary standard variable attenuators are readily available via commercial sources, the **P** (as discussed in paragraph 12-230) can be measured with either variable or fixed-signal sources. At microwave frequencies, the signal source can be a fixed-output gas discharge source. However, as the ratio of the P factor increases, for a fixed input, the direct output power measurement becomes unusable because of the increasingly large dynamic ranges required.

12-241. MEASUREMENTS.

12-242. GENERAL. In the event that the ratio represented by P (as discussed in paragraph 12-230) equals 2, then the input power can be varied and used in conjunction with an i-f attenuator. However, if the input is not variable, and P is selected as 2, then a fixed 3-db pad can be switched in or out of the circuit to establish the measurement conditions. This method, using the 3-db pad, is accomplished by setting the gain of the equipment being measured to a convenient reference level, as seen on the indicator, while the equipment being measured contains only normal internal noise levels. Next, the 3-db pad is inserted at a point within the equipment following the primary input noise contributing elements. The insertion of this pad will lower the indicator reading. Now increase the input signal until the indicator increases to the value set before the 3-db pad insertion. The increase of input signal required to reposition the indicator to its original point corresponds to P equals 2. Using these known factors, you can compute the noise figure.

12-243. When the applied input power source cannot be varied, you can still use the ratio P equals 2 by adding a calibrated attenuator between the signal source and the equipment to be measured. The attenuator acts as a control over the source generator output, and this will permit you to compute the signal available at the equipment input terminals. The method of computation is as follows: Set both the 3-db pad and the rf attenuator for zero attenuation, and turn off the source power. A signal reference level can then be established with the equipment gain control of the output indicator. Now, turn the power back on, insert the 3-db pad and the rf pad to bring the signal level back to its original point, where the input signal power equals the noise power of the equipment with reference to the equipment terminals. Using these known factors, you can now compute the noise figure.

12-244. In the event that you cannot establish the condition of P equals 2 because the noise figure being measured is high and the input signal power is limited so that P is less than 2, or you use a gas-discharge generator whose output cannot be varied, you can use a precision variable i-f attenuator to determine the exact P ratio. The method of P measurement consists of turning off the applied input power and setting the indicator to the desired reference level. Next, turn on the source input power and insert enough attenuation to return the indicator level to its former reading. The value of P is provided directly by the amount of attenuation inserted.

12-245. AUTOMATIC MEASUREMENT. The measurement of noise figure consists of measuring the ratio of the outputs from an equipment under two input conditions. Therefore, if you set up a switching arrangement to switch rapidly back and forth between the two input conditions and measure the amplitude of the output change at the switch frequency, you will obtain a direct measure of noise figure. Many direct-reading noise figure indicators are based on this measurement technique.

12-246. The automatic indicator uses a modulated noise generator as an input signal source. The output of the equipment being tested will contain two noise levels, one representing equipment noise and the other representing the combination of equipment noise plus the input generator noise. The ratio of these two noise levels is the P-factor. Both noise levels will have some average level. However, short peak fluctuations within the average is normal. The ability of an indicator to indicate small differences between the two noise power levels (possibly corresponding to large noise figures) is directly proportional to its ability to reduce these fluctuation peaks to less than the difference between the two signals. The indicator's ability to reduce these short term fluctuations corresponds to its ability to select and measure the magnitude of the switch-frequency component and its harmonics from the noise background. For noise figures up to 20 db, measurements taken through the noise background can be obtained by using narrow-band filters, tuned to the switch frequency, to remove the background noise prior to measurement. To measure higher noise figures, a cross-correlator can be used. In the cross-correlator, the signal from the equipment is switched from one integrator to another when the noise generator is switched from on to off. Therefore, the output can be averaged in each position over any desired period of time. This type of indicator (using the cross-correlator) has been built to read noise figures up to 40 db.

12-247. To obtain a single representative noise figure from the noise signal power being measured, it is necessary to read the ratio of the combination of the two output signal powers (P). You could use a ratio meter to accomplish this result. Another method of measuring this single resultant is to apply an automatic gain control function to the equipment being measured. The gain control function is obtained as a result of either one of the two power levels being measured or as a result of the average power level of the ratio of the two noise levels being measured. When the gain function is obtained as a result of either noise level, it is used in the synchronous integrator type of indicator combination. However, if it is the result of both signal noise ratios, it is used in the simple filter integration type indicator combination. The source of the agc noise voltage determines what laws are being followed by the output indicator. The applications of this type of measuring device are many; direct production line equipment testing permits the determination of the combined input operating components with respect to the output noise figure, and permits pass band adjustment with respect to the output noise figure. This latter application is useful during the testing of different input voltages to a traveling-wave tube where the input voltage affects the output noise figure.

12-248. The automatic method of measurement reduces the effect of any equipment signal gain drift during the measurement process. By using the switching (on-off) automatic arrangement, sampling the noise ratio many times per second, and using agc long-term drift correction, the effect of signal gain drift is reduced. For the automatic measurement method, where the on-impedance must equal the off-impedance of the source generator, the noise source generator must be of the transmission type. Gas-discharge noise generators in the E-plane and the helical type are excellent for use in the vhf, uhf, and shf noise regions. However, when noise diodes are used under the same conditions, the diode current must be completely shut off during the "off" condition, and the switching waveform must not produce an output at the radio frequency of the equipment, especially in the frequency region used by the noise diodes.

12-249. POWER SOURCES. Noise figure measuring techniques depend on a known amount of noise signal power, developed across a known source impedance, being applied to the input terminals of the equipment under test. Source noise signal power calibration error or incorrect assumptions concerning the actual signal value will directly affect the accuracy of measurement.

Therefore, you should make a careful selection of the proper signal source for your particular application. The most commonly used sources are general purpose signal generators, temperature limited diodes, gas-discharge-tube generators, and, recently, impulse noise and mercury droplet generators.

12-250. The general purpose signal generator is available in almost all frequency ranges of interest. However, its use is restricted because it has poor measurement accuracy. Therefore, this type of power source is normally used to obtain only rough approximations.

12-251. The ratio of the output noise power (P) from an equipment when that equipment was terminated in a source impedance having a particular temperature (290 degrees Kelvin) is discussed in paragraph 12-230. However, you probably know that the operating temperature may not always be exactly 290 degrees Kelvin. Therefore, the ratio of the output noise power (P) is calculated as discussed in paragraph 12-230 with the constant temperature assumption of 290 degrees Kelvin, and then P can be recalculated when additional noise signal power is introduced to the input of the equipment under test. The output power at the constant temperature is termed P_2, and the changed temperature power becomes P_3; therefore, P equals $\frac{P_3}{P_2}$, from which the noise figure can be determined. The following formulas provide the criteria for measurement:

$$P_2 = AKT_1B_1 + (NF_1 - 1)AKT_oB_1$$

where:

A = Gain

T_1 = Temperature of equipment

T_o = 290 degrees Kelvin

B_1 = Useful bandwidth

NF_1 = Measured noise figure

K = Boltzmann's constant (1.38×10^{-23} joules per degree Kelvin)

$P_3 = \Delta P + P_2$ (ΔP = additional signal noise power at input)

Therefore, since $P = P_3/P_2$, the measured noise figure (NF_1) equals

$$\frac{\Delta P}{(AKT_oB_1)(P-1)} - \frac{(T_1 - T_o)}{T_o} + \frac{T_1}{T_o(P-1)},$$

or, if the ratio of P equals 2, and $\frac{T_1}{T_o}$ is approximately 1, the noise figure (NF_1) equals $\frac{\Delta P}{AKT_oB_1} + 1$. Figure 12-97 represents a graphic determination of the noise figure for various useful bandwidths (B_1). You can add or subtract 10 dbm algebraical-

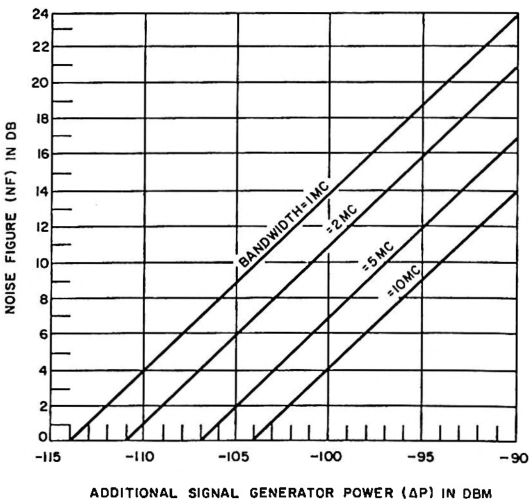


Figure 12-97. Noise Figure as a Function of Signal Generator Power for P = 2

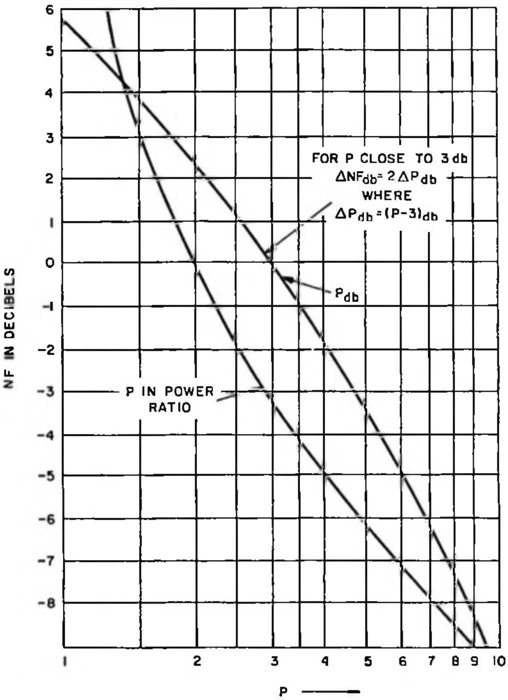


Figure 12-98. Correction Factor (ΔNF) To Be Added to Noise Figure (NF) When P Is Not Equal to 2

ly from the ordinate for each factor of 10 in the bandwidth to extrapolate these sets of bandwidth curves to fractional values. For example, to divide the bandwidth by 10, add —10 dbm to the ΔP scale, and, to multiply the bandwidth by 10, add +10 dbm to the ΔP scale. Figure 12-97 assumes that the ratio P equals 2. If P does not equal 2, a correction equal to $1/(P - 1)$ must be made to the computed noise figure, and, if T_1/T_o does not equal 1, you must subtract

$$\left(\frac{T_1 - T_o}{T_o}\right)$$

from the calculated noise figure. The graph shown in figure 12-98 illustrates a plot of the conversion factor $1/(P - 1)$ required when P is not equal to 2. In addition, when T_1/T_o is not equal to 1, the factor $(T_1/T_o - 1)$ must be algebraically added to the measured noise figure. This factor is added as a small change (ΔNF) in noise value. Figure 12-98 shows this change on the vertical axis of the graph. The horizontal axis denotes the value of P. However, the graph provides only the noise figure correction factor when P is not equal to 2. It does not provide a noise figure sum or total noise figure. When P is not equal to 2 and T_1/T_o is not equal to 1, the noise figure is calculated as follows:

$$NF_1 = \frac{\Delta P}{AKT_oB_1} + 1$$

$$\Delta NFdb = -(T_1/T_o - 1)$$

$$NF = NF_1 + \Delta NF$$

where:

NF_1 = Measured noise

ΔNFdb = Small change in noise representing a correction factor

NF = Actual noise figure total

The other symbols used above have been previously explained.

12-252. In addition to the correction factor (ΔNF) added to the measured noise figure (NF_1) when P is not equal to 2 and ΔNF is not equal to 1, you must also add a correction factor to the measured noise (NF_1) to compensate for the temperature of the source impedance. The amount of compensation to be added to the measured noise figure because of source temperature is shown in figure 12-99. As shown in the graph, as the noise figure increases, the correction factor approaches zero.

12-253. NOISE MEASUREMENT METHOD, USING SIGNAL GENERATOR. As has been discussed in paragraphs 12-242 through 12-252, a signal generator is connected to the

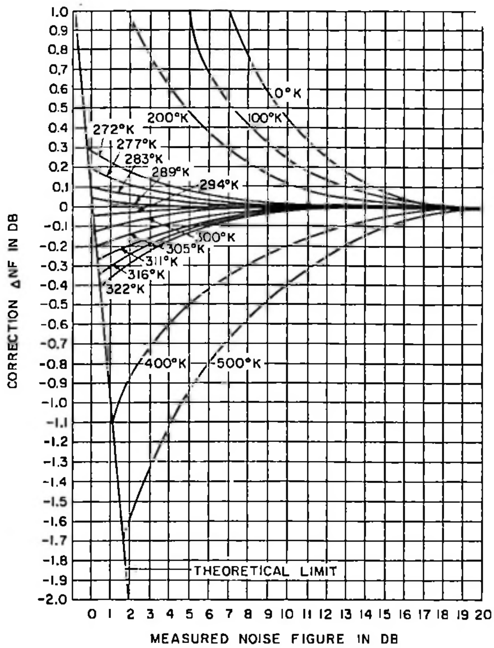


Figure 12-99. Correction Factor To Be Added to Measured Noise Figure, Required as a Result of Source Temperature

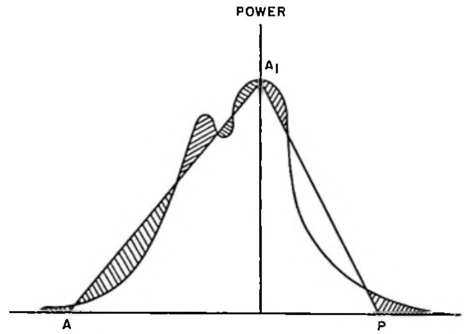


Figure 12-100. A Method of Determining Useful Channel Noise Bandwidth

input terminals of the equipment to be tested. The signal generator must not affect the response of the equipment to be tested, and it must have known temperature characteristics. Now, with a proper output indicator, the signal generator output power required to double the tested equipment output power is determined. The measured noise figure (NF_1) can be obtained from the graph of figure 12-97, or it can be computed from the equations contained in paragraph 12-251. If T_1 does not equal T_0, or P does not equal 2, you must apply the correction factors, contained in both figures 12-98 and 12-99, to obtain the actual noise figure (NF).

12-254. The noise bandwidth (B_1) can be determined by plotting the response curve for the power contained in the useful channel. However, you must know the exact frequency (f) at which ΔP, the additional signal noise power at the input, is measured. This exact frequency (f) occurs at the point where maximum gain is obtained. Unfortunately, when dealing with multiple response curves, the maximum gain point may be difficult to determine. One possible method of obtaining B_1 is to measure the physical area contained under the response curve and to divide this figure by the power gain obtained at frequency f. Figure 12-100 illustrates a simple method used to determine the bandwidth. Two lines are laid out, one from A to A_1 and one from P to A_1. Each line is placed in such a fashion that the cross-hatched area above the line equals the cross-hatched area below the line. The bandwidth (B_1) is equal to the distance from A to P divided by 2.

12-255. A random noise generator (temperature-limited diode) will provide a signal source which generates a known amount of power at the required level for noise figure measurements. This type of signal source does not require large precise values of attenuation, and the possibility of error caused by stray radiation from a high-power source is reduced. All of the electrons emitted from the cathode reach the plate and cause anode current in the temperature-limited diode noise generator. Normally, the plate voltage is increased until a point is reached where an increase in plate voltage does not increase plate current. Above this point is the so-called temperature-limited region, where the anode current is limited only by the number of electrons emitted by the cathode. Actual circuit operation is explained under paragraph heading 8-827 of Volume VI.

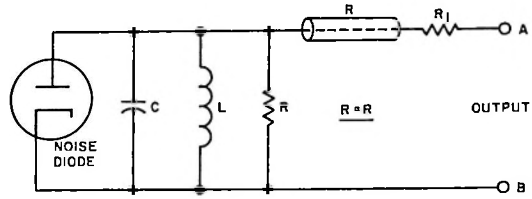


Figure 12-101. Matched-Line Diode Noise Generator

12-256. The shunting effect of stray and wiring reactances across the output of the source signal generator at higher frequencies will distort the input to the equipment under test unless compensation is employed. You can place an inductance across this resistive output and cancel the capacitive effects at a particular frequency. However, this type of circuit arrangement would be resonant at only one frequency. In other words, the source input impedance would be frequency-sensitive and would require readjustment for each new frequency at which measurements are to be made. In addition, if the source impedance is resonated to a particular frequency, and this is the frequency of measurement, this resonant frequency may not permit the proper impedance to be offered to the sideband responses of the tested equipment.

12-257. MATCHED-LINE NOISE SIGNAL GENERATOR. The matched-line noise signal generator circuit is illustrated in figure 12-101. The diode noise generator is terminated in R, L, and C, which are all adjusted to be resonant at the frequency of interest. A cable, with the same value of resistance as the R value connected across the diode, is connected in series from the diode to the point of measurement. A series limiting resistor (R_1) is connected in series at the point of measurement. Now any load can be placed across the output terminals (A to B) to make the source impedance conform to any value desired. Since the circuit Q at the generator is very low, the generator reactance is also very low and may be neglected in the computation to determine the total source impedance. The matched line noise generator requires a large current to measure a given noise figure. For the high-impedance version of the noise

generator $I = \frac{F}{20R} = 0.2$ ma, and for the

matched-line generator $I = \frac{(R + R_1)F}{20R^2} = 20$

ma, or 100 times larger current.

12-258. COAXIAL-LINE NOISE SIGNAL GENERATOR. The coaxial line used with the noise diode will extend the upper frequency limit of measurements. The plate forms the outer conductor, and the filament forms a loop at right angles around the center conductor. This arrangement eliminates the lead inductance. The limitation on the highest frequency at which the tube can function properly is the transit angle. Normally, 3000 megacycles is set as this limit. The output of the noise signal generator is generally connected through a multiposition switching arrangement, to provide flexibility in resonant frequencies by switching to alternate output impedances. Unfortunately, at higher frequencies, the switch introduces additional undesirable lead reactances.

12-259. For the measurement of high noise figures, the noise diode is one of the best signal sources at frequencies where high-frequency errors are not a primary factor.

A temperature-limited diode generally has a known and controllable output noise power. Unfortunately, the output is limited by heat dissipation. However, heat dissipation is not a factor unless you are measuring the high output required for poor noise figures. If great accuracy is not a requirement, P can be made smaller than 2. If P is greater than 2, gain stability is not as important as when P approaches 1.

12-260. GAS DISCHARGE TUBE SIGNAL GENERATORS. The apparent noise output from gaseous discharges is large enough to be detected with sensitive equipment at microwave frequencies. The source of these noise-producing gas discharges may be a tr tube, neon bulb, fluorescent lamp, etc. In particular, the fluorescent lamp can be impedance-matched over a wide band of frequencies to a waveguide type circuit, and the fluorescent noise output will be comparatively constant from one lamp to another. Actually, the noise power is a result of the random velocities of the electrons contained in the discharge, between the times of collisions with the positive ions contained in the beam. The electron temperature of the gas discharge is a function of the type of gas employed, the diameter of the full gas discharge, the gas pressure, etc. Neon and argon are the principal gases used. The microwave noise temperature is equal to the electron temperature. Therefore, the output of the generator can be described by treating the output as a physical body with a temperature equal to the electron temperature. The ratio factor (P) must be again calculated for this type of generator, as it has been for diode noise generators. The equations are essentially the same as those previously used. In fact, P_2 is exactly the same, and P_3 is changed only by the additional electron temperature factor (T_2), as shown below.

$$P_2 = AKT_1B_1 + (NF - 1)AKT_oB_1$$

(Where all symbols are defined in paragraphs 12-230 and 12-251)

$$P_3 = AK(T_2 - T_1)B + AKT_1B_1 + (NF - 1)AKT_oB_1$$

(Where all symbols are defined in paragraphs 12-230 and 12-251, except that T_2 = generator output electron temperature)

$$P = \frac{P_3}{P_2}$$, therefore, if $T_1/T_o = 1$, then

$$NF = \frac{(T_2/T_1 - 1)\,\frac{B}{B_1}}{P - 1}$$, and if $B/B_1 = 1$,

$$\text{then NF} = \frac{\left(\frac{T_2}{T_o} - 1\right) - P\left(\frac{T_1}{T_o} - 1\right)}{P - 1}.$$

If both conditions exist, that is, if

$$\frac{T_1}{T_o} = 1 \text{ and } \frac{B}{B_1} = 1, \text{ then NF} = \frac{\frac{T_2}{T_o} - 1}{P - 1}.$$

Charts and graphs are available from many sources to provide the same correction factors which were discussed for the diode noise generator.

12-261. TRANSMISSION LOSS. The two types of discharge generators used in conjunction with a waveguide have individual characteristics. In one type, the gas-discharge tube is mounted perpendicular to the axis of the guide in the H-plane (parallel to the broad side of the guide); in the other, it is placed parallel to the narrow side of the guide. The gas-dischrage tube type which is placed parallel to the narrow side of the guide provides a broadband match to the guide when fired, and permits the use of a single generator over the entire rec-

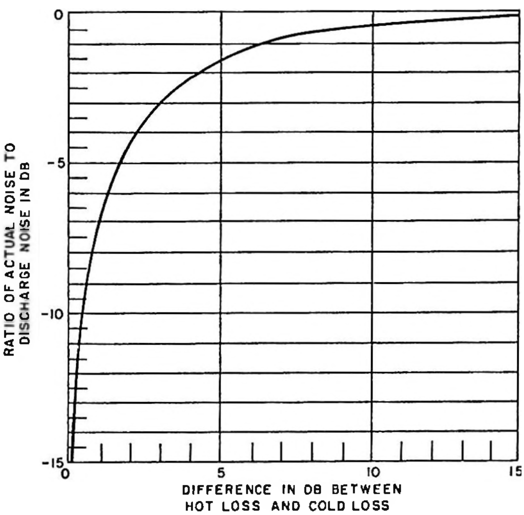


Figure 12-102. Effect of Transmission Losses of a Discharge Generator

ommended frequency range of a given waveguide. When the generator is in the "off" condition, that is, when it is not being fired, it acts as a simple section of transmission line because of its negligible mis-match characteristic. Therefore, a matched termination placed at the end of the generator (away from the equipment under test) may be used for the T_1 termination. Considering that the effective output temperature is less than the temperature of the gas discharge, a resultant correction must be made. The noise temperature ratio of an impedance, as was previously discussed, is the ratio of its apparent temperature to 290 degrees Kelvin. If the apparent temperature of the generator (T_g) minus 1 is divided by the temperature ratio of the gas discharge (T_d) minus 1, it will equal 1 minus the cold transmission loss (L_c) in the off condition divided by the hot transmission loss (L_h) in the on condition, as shown by the relationship:

$$\frac{T_g - 1}{T_d - 1} = 1 - \frac{L_c}{L_h}$$

The reduction of noise in db between the hot fired (on) condition (L_h) and the cold (off) condition (L_c) is illustrated in figure 12-102. As shown in figure 12-102, any circuit which provides a 12-db transmission loss when hot more than it provides when cold, can be used as a noise generator. Coaxial noise generators using gas-discharge tubes have been developed, to cover the range from 200 to 2300 megacycles, in a single unit. The criteria of a good generator is that the transmission loss difference between hot and cold (on - off) must be greater than 12 db, and the standing wave ratio of the line must be small as seen from either end of the line.

12-262. SPOT FREQUENCY VERSUS AVERAGE NOISE FIGURE. The spot frequency noise is that noise obtained over an extremely small bandwidth. The average noise figure is obtained by averaging the spot noise figures over the useful noise bandwidth. The spot noise figure becomes important only in the case of cascaded networks.

12-263. OPERATING NOISE FIGURE. Previous calculations defined the noise figure in terms of a standard temperature source (290 degrees Kelvin). The operating noise figure is defined in terms of a source impedance possessing a temperature other than the standard (290 degrees Kelvin). As has been previously stated, the noise figure is a factor used to define the over-all sensitivity of an equipment. As the sensitivity of an equipment is improved, the noise figure should decrease.

12-264. The effective noise figure is an attempt to reflect the effect of a nonstandard antenna temperature when the antenna is attached to the equipment under test. This method compares the equipment under test with a similar perfect equipment terminated in a nonstandard antenna. Unfortunately, the noise figure increases with a decrease in antenna temperature, whereas for a fixed source noise signal input, the signal-to-noise ratio is improved for the same antenna temperature change. Therefore, the modified operating noise figure takes into account the case of a receiver whose useful channel bandwidth is not equal to the over-all noise bandwidth, as shown by the following equation:

$$NF_o = NF + (T_a - 1)\left(\frac{B}{B_1}\right)$$

where NF_o is the operating noise figure, NF is the noise figure at 290°K, T_a is the antenna temperature ratio, B is the over-all bandwidth, and B_1 is the useful bandwidth. This formula provides a noise figure which represents the ability of the equipment to detect a fixed amplitude of signal power as available from the antennas. Considering that most noise figures are obtained when the source impedance is other than the standard temperature (290 degrees Kelvin), the measurement of noise figure will normally be the operating noise figure at a nonstandard temperature. In this case, you must apply the correction factors provided in the graph of figure 12-102 to obtain the true noise figure.

SECTION VII

THE DETERMINATION OF IMPEDANCE AND DISTRIBUTED CONSTANTS AT MICROWAVE FREQUENCIES

12-265. GENERAL.

12-266. At the lower frequencies the electrical wavelength may be several miles long. Consequently, to simulate the electrical properties of a capacitor or inductor, a wire, pipe, waveguide, or other physical conducting media must possess at least 1/8 the physical length of the electrical wavelength, and even this 1/8 value may be several miles long. When frequencies are increased, the wavelength of each cycle becomes shorter to permit more cycles to occur during a given time interval. This is illustrated by the following basic relationship:

$$\lambda = \frac{982.08 \times 10^6 \text{ ft}}{F}$$

where: λ = Wavelength in feet

982.08×10^6 = Constant for measurement in feet

F = Frequency of measurement in cycles per seconds

For example, to find the wavelength for frequencies of 0.98208 megacycle, 9.8208 megacycles, and 982.08 megacycles, solve the above equation for each frequency as follows:

$$\lambda = \frac{982.08 \times 10^6 \text{ ft}}{0.98208 \times 10^6 \text{ cycles}} = 1000 \text{ ft}$$

$$\lambda = \frac{982.08 \times 10^6 \text{ ft}}{9.8208 \times 10^6 \text{ cycles}} = 100 \text{ ft}$$

$$\lambda = \frac{982.08 \times 10^6 \text{ ft}}{982.08 \times 10^6 \text{ cycles}} = 1 \text{ ft}$$

As you can see from these simple calculations, as the frequency increases the wavelength decreases in direct proportion. This fact makes it practical to use a waveguide, pipe, or other physical conducting media to replace physical components at the higher frequencies but not at the lower frequencies. One mile of waveguide would be required to represent a complete wavelength at a frequency of 186, 000 cycles per second, while only one foot of waveguide would be required to represent a complete wavelength at a frequency of 982.08 megacycles.

12-267. IMPEDANCE.

12-268. In electrical circuits, impedance is the quality which, at any frequency, opposes, retards, and limits the current flow, or free movement of electrons, through a conductive media. In the lower-frequency ranges it is comparatively obvious that pipes and waveguides would not be used to simulate circuitry; the physical length of these objects would prohibit their use. However, as the frequency increases and the wavelengths become shorter, the electrical characteristics become changed to the extent that the impedance is dominated or controlled by different circuit elements. As an illustrative example:

$$R = \frac{E}{I} \quad \text{(pure resistance)}$$

where:

R = Resistance

E = Voltage

I = Current

This is simply Ohm's law for dc circuits containing zero frequency elements. However, as the frequency increases above zero, reactive opposition will begin to retard, or limit, the current flow, in addition to the resistive element limitation. This is true because of the following mathematical facts:

$$X_L = 2\pi FL$$

$$X_C = \frac{1}{2\pi FC}$$

$$Z = \sqrt{R^2 + (X_L - X_C)^2}$$

where:

Z = Impedance (ohms), for series circuit

X_L = Inductive reactance (ohms)

X_C = Capacitive reactance (ohms)

R = Resistance (ohms)

$2\pi = 6.28$

F = Frequency

From the above formulas it can be proven that as the frequency increases X_L will increase and X_C will decrease in direct proportion; the resistance value will not change. The impedance (Z) is a combination of all impedances within any complex circuit. Theoretically, if the frequency is made low enough, the capacitive reactance will become very large, and the inductive reactance will become very low. Conversely, if the frequency is made large enough, the capacitive reactance will become small and the inductive reactance will become large.

12-269. Considering that capacitive or inductive reactance elements are in opposition to current flow, when one of these reactive elements becomes extremely low, it does not oppose the current flow to any great extent. Since electrical current will always seek the path of least opposition, in a parallel LC circuit current will follow the capacitive-reactance path at the higher frequencies. This characteristic is used to filter out undesirable currents at the higher frequencies by providing a low-reactance path to a different part of the circuit.

12-270. From the discussions of the last few paragraphs, the following facts emerge: Impedance (Z) represents the total opposition within any electrical or electronic circuit; impedance is composed of inductive reactance, capacitive reactance, and resistance. At higher frequencies, impedance is composed primarily of inductive-reactance elements as the opposition factor. Therefore, in a parallel circuit, the higher frequency currents will flow through a low capacitive-reactance path in preference to any other path. The general mathematical formula for the impedance (Z) of a series circuit is changed at the high frequencies because X_C approaches zero; therefore, only the inductive reactance and the resistance becomes important. A further fact emerges from this discussion—you must eliminate or compensate for any remaining capacitive paths, as you increase frequency, to prevent the loss of high-frequency current, unless it is your deliberate intention to lose or filter out these currents.

12-271. DISTRIBUTED CONSTANTS.

12-272. RESISTANCE. The value of a resistor will not change with a change in frequency; therefore, this constant value can be used as a reference point when discussing the changing-frequency versus changing-impedance values.

12-273. CAPACITANCE. As previously discussed, when the frequency increases, the capacitive reactance decreases. However, there is much more to discuss than this one point. At the frequency increases, many more capacitive reactive sources become important. This is true because at low frequencies the capacitive reactance is so large that small sources of capacitive reactance can be ignored as being too small to contribute significantly to the total impedance. However, at higher frequencies, these sources of small capacitive-reactance values constitute almost all of the circuit capacitive reactance, and reduces the total impedance value by directly canceling part of the inductive reactance. These small capacitive elements, which play such an important part in high-frequency circuits, come from such sources as the following: two wires close together, each representing the plate of a capacitor; one wire close to chassis ground, the wire forming one capacitor plate, and the chassis ground the other; elements within a tube form plates of a capacitor and provide high-frequency current feedback paths; even the electrons traveling through a wire each form one plate of a capacitor. As you can see, the elimination of all capacitive sources is virtually impossible. However, these capacitive sources definitely limit our progress into the elevated planes of frequencies above those used today. Therefore, advancement into higher-frequency ranges, being limited by the existing capacitive sources, cannot be accomplished unless we first devise the means of eliminating or decreasing the reactive value of these sources.

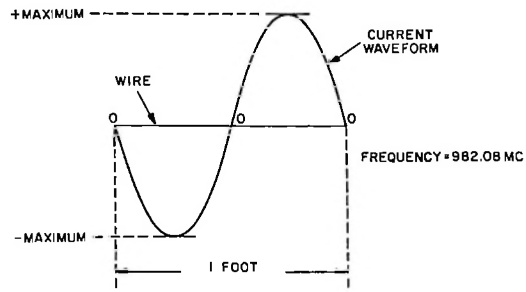


Figure 12-103. Wavelength and Conductor of Equal Length

12-274. INDUCTANCE. As the frequency increases, the inductive reactance increases in a direct proportion. Therefore, the total impedance of a series LC circuit at higher frequencies is comprised almost entirely of inductive reactance; the capacitive reactance electrically decreases, and the circuit resistance remains constant but becomes smaller in comparison to the increasing value of the inductive reactance.

12-275. WAVELENGTH VERSUS IMPEDANCE.

12-276. As the frequency increases, the wavelength becomes shorter, until eventually it becomes short enough to be measured along a short piece of wire contained in some electrical or electronic apparatus. Suppose, for example, that the frequency is exactly 982.08 megacycles, making the wavelength exactly one foot long. As illustrated in figure 12-103, there are three points along this one foot line length which represent zero circuit current. Also, there is one point representing maximum positive current and one point representing maximum negative current. Between the zero and the maximum current points are current amplitudes greater than zero but less than the maximum. These points are of considerable significance because, if the line were cut off at any one of these points and thereby became shorter than 1 foot,

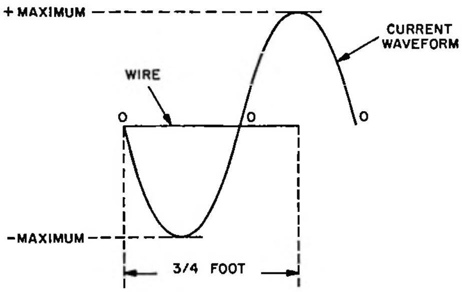


Figure 12-104. Wavelength and Conductor of Unequal Length

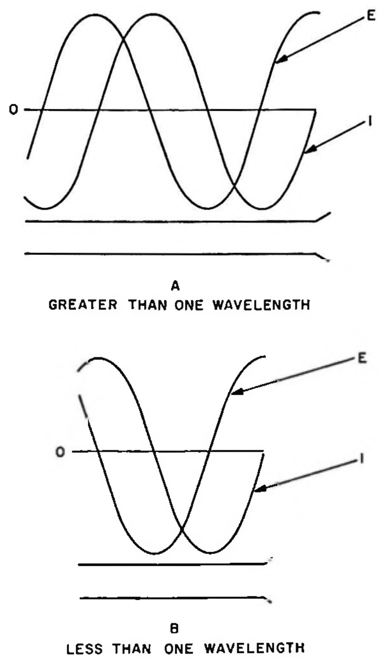


Figure 12-105. Open-Circuited Lines

some part of the one foot long wavelength would be left hanging over and extend beyond the present shortened wire length, as shown in figure 12-104. This shortened wire length, which contains some current at its extreme end, can actually represent an electrical component such as a capacitor or even a resonant or nonresonant circuit.

12-277. The discussion of paragraph 12-276 presents a true impression of what happens to a current waveform as measured along a conductive line, but you can readily realize that, without pressure, force, or some motivating factor, there can be no circuit current. Therefore, along this same length of wire we must represent this current driving force as to amplitude, phase, and frequency relationship with the current waveform. Secondly, any current or voltage amplitude must be represented with respect to some other amplitude or reference point. Therefore, two wires of equal length should be shown in parallel, with the various waveforms existing on one wire with respect to the other. Figures 12-105 and 12-106 are representative of this situation.

12-278. In figure 12-105, both lines are shown as open-ended. A resistance measurement across the ends of these lines will show an infinite impedance. An infinitely large impedance will limit the current at that point to zero. However, you know that the larger the impedance, the more force (voltage) will be required in an attempt to force current through that point. In other words, the open-circuit point will show, for all practical purposes, infinite impedance, the total applied voltage, and zero current. In figure 12-106, just the opposite situation holds true. These lines are shorted to each other at the end points. A short represents a theoretical impedance of zero, and, with no opposition, the maximum current can be pushed through this point with a minimum of force. These examples are not concerned with the current lead or lag with respect to voltage, but only the current amplitude with respect to the amplitude voltage; figure 12-107 illustrates this point.

12-279. INFINITE LENGTH TWO-WIRE LINE.

12-280. GENERAL.

12-281. In normal circuits the inductive,

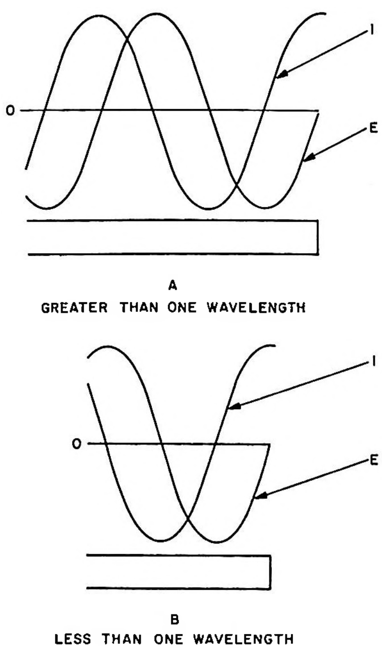


Figure 12-106. Short-Circuited Lines

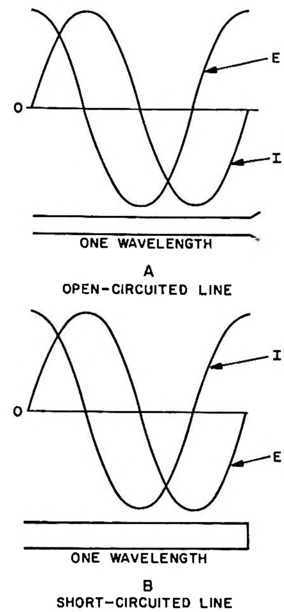


Figure 12-107. Voltage and Current Waveforms Versus Open and Shorted Lines

resistive, and capacitive components are present in the form of physical "lumps." These individual capacitors, coils, and resistors can be changed, replaced, measured, etc. Unfortunately, in two-wire lines, these quantities are distributed throughout the length of the line and cannot be separated into individual components for measurement, replacement, etc.

12-282. CHARACTERISTIC IMPEDANCE.

12-283. BASIC CONSIDERATIONS. The characteristic (surge) impedance of an infinitely long line is the impedance of the line in ohms, at the operating frequency, as seen by the source generator at the input terminals of the line. Therefore, the characteristic impedance of the line is actually present at the line input. This known value of impedance at the input has many valuable uses; no matter how long the line is, if you connect a load equal to the characteristic impedance to the output end of the line, the impedance reflected back to the input end of the line will still equal the original value of the characteristic impedance. However, for a given type and size of line, only one impedance value can be used for a load at the output without affecting the input. Any line has the property of inductance, capacitance, and resistance. The inductive value exists because magnetic flux linkages are established within the wire when any current flows through the wire. For example, two open-ended number 12 lines spaced 6 inches apart have an inductance of approximately 0.6 microhenry per running foot. The capacitance value exists because the two wires each become the plate of a capacitor with air or other material between the wires as a dielectric. For example, the capacitance of the above two open-ended number 12 lines spaced 6 inches apart is approximately 1.7 picofarads per running foot. The resistance

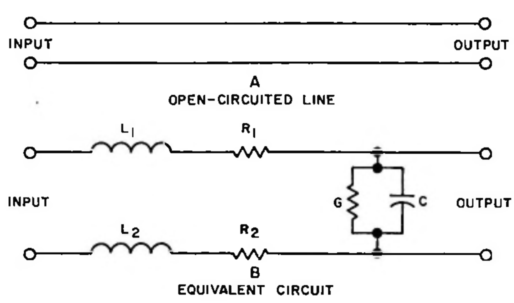


Figure 12-108. Two-Wire Line of Definite Length

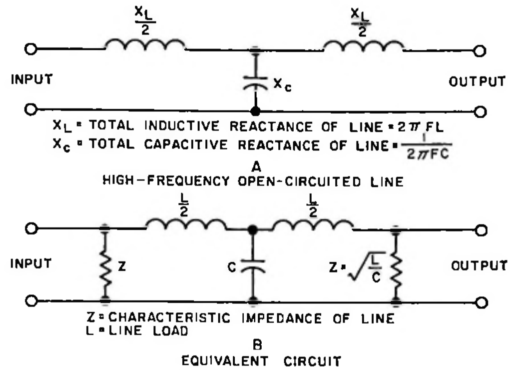


Figure 12-109. High-Frequency Equivalent Circuit of a Two-Wire Line of Definite Length Terminated in Its Characteristic Impedance

of the above two-wire line is of two different types: the actual resistance of the wire to the flow of current through the wire and the leakage resistance from the wire through the insulating (dielectric) material separating the wires. The resistance of the wire to current flow (no material is a perfect conductor) varies in direct proportion to the length and inversely with the cross-sectional area of the wire. The leakage resistance occurring as a result of current flow between the wires (through the insulating material separating the wires) is made possible because no substance is a perfect insulator. The leakage resistance is normally expressed in terms of conductance (reciprocal of resistance). The conductance value is normally in the range of a few picomhos per running foot of wire. These various quantities comprising the characteristic impedance of a line can be replaced or represented by lumped physical components, as shown in figure 12-108. If a voltage is applied across the input terminals of figure 12-108, a current will flow through the characteristic impedance of the line. This impedance can be calculated as follows:

$$Z = \frac{E}{I}$$

where:

Z = Characteristic (surge) impedance (ohms)

E = Applied input force (volts)

I = Current flow (amperes)

12-284. At the higher frequencies, the conductor resistance and leakage resistance are small in comparison to the inductive reactance of the line. Therefore, as shown in figure 12-109, an open-circuited two-wire line of definite length can be represented by the lumped constants of L and C. The open-circuited line shown in part A of figure 12-109 can be terminated in the characteristic impedance represented by an infinite number of "T" sections of L and C.

12-285. In the high-frequency region, the value of the characteristic impedance depends almost entirely on the distributed inductance and capacitance of the line. However, these qualities will change with any physical change of the two-wire line. For example, increasing the separation between the wires will increase the inductance and decrease the capacitance of the two-wire line. The increase in inductance value occurs because of the increased magnetic flux between the two wires, and also because the current traveling in one direction on one wire is not close enough to the current traveling in the opposite direction on the other wire for the resultant fluxes surrounding the wires to cancel. The capacitance effects are lowered because you have effectively provided greater separation between the capacitor plates when you move the wires farther apart. The characteristic impedance of the line will also be changed if you change the cross-sectional area of the wires; if you decrease the diameter of the wire, you effectively decrease the area of the capacitor plates represented by the wires, and thus decrease the capacitance value of the line. Changing the type of dielectric material contained between the wires of the two-wire line will change the line capacitance, resulting in a new characteristic impedance of the line. Therefore, if you change the dielectric material in such a fashion that the capacitance between the wires will increase, the characteristic impedance of the line will decrease. The characteristic impedance of a two-wire line, using air as a dielectric between the wires, can be determined by the following formula:

$$Z = 276 \log_{10} \frac{b}{a}$$

where:

a = Radius of one conductor

b = Spacing between centers of conductors

The characteristic impedance of a concentric line or coaxial line can be calculated by the following formula:

$$Z = 138 \log_{10} \frac{b}{a}$$

where:

a = Outer diameter of inner conductor

b = Inner diameter of outer conductor

12-286. WAVE MOTION ON LINES OF INFINITE LENGTH. On a two-wire line of infinite length, when a voltage (force) is applied to the input of the line, a current waveform will travel on the line. Figure 12-110 illustrates this point where all waveforms have been stopped when the applied voltage reached zero. The resulting waves still exist because it takes time to propagate them along the line.

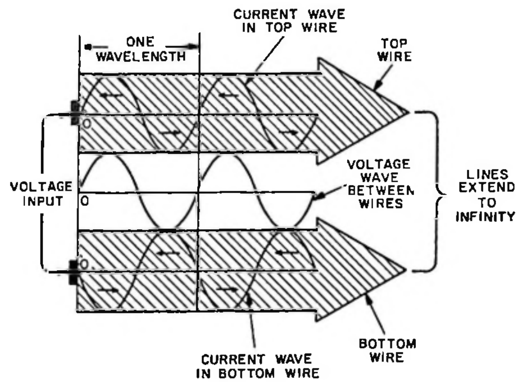


Figure 12-110. Traveling Waves of Current and Voltage on a Two-Wire Line Extending to Infinity

12-287. Important facts pertaining to figure 12-110 can be summarized as follows:

a. The applied voltage and the resultant current are in phase throughout the length of the infinitely long two-wire line.

b. The characteristic impedance, represented by the ratio of the voltage to the current, is constant over the entire length of the line.

c. The input impedance of the infinitely long two-wire line is always equal to the characteristic impedance.

d. The infinitely long two-wire line operates at maximum efficiency because the voltage and current are in phase along its entire length.

e. Any length of line can be made to appear as an infinite line by terminating the line with a load equal to the characteristic impedance of the line.

12-288. LINE REFLECTIONS.

12-289. GENERAL. When a two-wire line is infinitely long or is terminated in its characteristic impedance, there are no reflections, or so-called standing waves, on the line. However, if an open or short circuit occurs any place along the line, maximum reflection will occur between that point and the source. Added impedance along the line, such as a poorly made splice, will cause a reflection with an amplitude which depends on the value of impedance introduced by the splice. The following discussion is based only on the two extremes—open-ended and shorted two-wire transmission lines.

12-290. OPEN-ENDED LINES. Initially an open-ended two-wire line acts as an infinitely long line to the traveling current waveform because the traveling wave cannot "see" the end until it arrives at that point. Both the voltage and current waveforms will remain in phase while the current travels along the line unless the two wires develop a difference in impedance at some point along the line. If this occurs, reflection will be produced on the line between that point and the source. If no added impedance points occur along the line, the current waveform will travel to the open terminals. At the open terminals the current will collapse to zero because current cannot exist across an open circuit. However, when the current collapses, its magnetic field surrounding the end of the line also collapses. This collapsing field cuts the conductor at the output end and thus induces an additional voltage at that point. The original voltage waveform existing across the line in phase with the current to the open end terminal point will be maximum across the open circuit. This maximum voltage plus or minus the induced voltage caused by the collapsing current will act as a voltage source generator. This generator action will cause a new current wave to be set in motion back along the line toward the original input source.

12-291. Figure 12-111 illustrates both the original (incident) traveling wave and the reflected wave for both the voltage and cur-

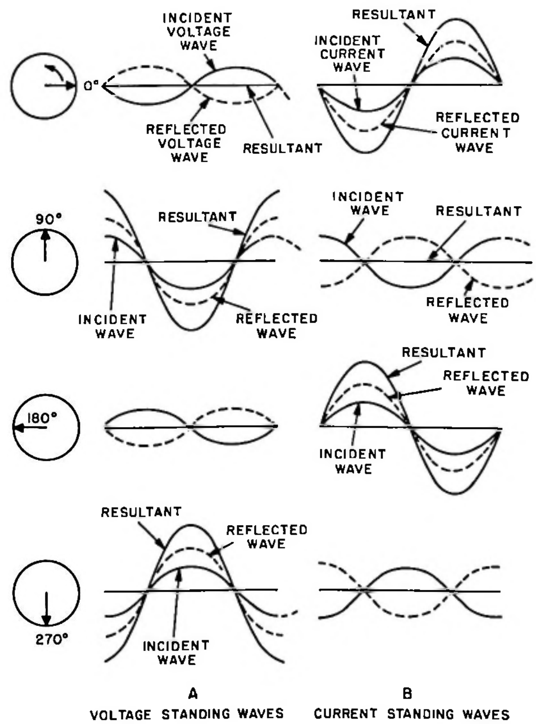


Figure 12-111. Formation of Standing Waves on a Two-Wire Open-End Line One Wavelength Long

rent waveforms on a two-wire open-ended line one wavelength long. Both the voltage and the current incident waveforms are in phase (no additional impedances occurring along the line length) when they leave the input source generator. The standing waves, as represented along the two-wire line, are instantaneous functions along the entire line during the same instant of time and do not represent elapsed time from the source to a particular point along the horizontal (X) axis. Only four positions of the generator voltage vector are shown (0, 90, 180, and 270 degrees). Other voltage vectors starting at points such as 45, 135, 225, and 315 degrees could be shown to provide a more complete picture. Figure 12-111 also shows each waveform as having stopped at the exact angle as specified by the symbol representing the generator. Therefore, at least one complete wave has already traveled to the end of the line and has been reflected back to the source, and the generator is prepared to start another wavefront. The voltage waveforms are illustrated on one plot, and the current waveforms are illustrated on another. However, both the voltage and the current waveforms actually exist simultaneously on one plot representing the original two-wire line.

12-292. When the voltage waveform originates from either a zero or a 180-degree source generator position, the incident wave and reflected wave can be considered separately because no prior waves exist on the line. However, when the generator is stopped at some other angular position, such as 45, 90, or 135 degrees, etc, the incident wave and reflected wave (result of a previous incident wave) are combined together along the line prior to forming a resultant.

12-293. The current incident waveform arriving at the end of the two-wire line does not act in the same fashion as the voltage waveform at this point. The current waveform is not reflected back at the same angle as the arriving incident waveform. This is true because the current waveform arriving at the end of an open-circuited line will collapse to zero amplitude and leave or bounce back from the open end starting from a zero-amplitude point.

12-294. CLOSED-END LINES. The current incident waveform arriving at the end of a shorted line will act the same as a voltage incident waveform arriving at the end of an open-ended line. In other words, the current waveform will always have maximum amplitude across the end of a shorted line because of the minimum resistance at that point. Therefore, the angle of arrival of an incident current waveform at the shorted end of such a line will equal the angle of departure of the reflected current waveform, with respect to a line drawn perpendicular (at right angles) to the terminal end of the length of the two-wire shorted line. The voltage incident waveform arriving at the end of a shorted line will react the same as a current incident waveform arriving at the end of an open-ended two-wire line. This condition can be explained with the aid of figure 12-111. Part A of the figure illustrates voltage standing waves, while part B illustrates current standing waves for an open-ended two-wire transmission line. For a shorted transmission line, the positions of part A and part B of figure 12-111 are reversed. In part A the waveforms will represent current rather than voltage, and in part B the waveforms will represent voltage rather than current.

12-295. The incident and reflected waves add algebraically to form the resultant waveform. Only the resultant waveform would be seen on an oscilloscope. The voltage waveform traveling toward the output end of the line can be considered as moving toward a solid object. This simply means that when the end of the waveform hits the end of the line it will be bounced back just

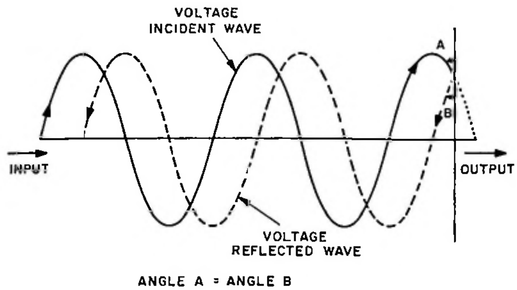


Figure 12-112. Voltage Incident Angle Versus Reflection Angle

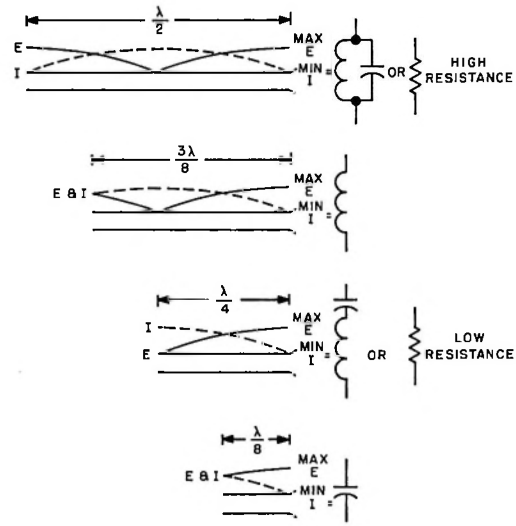


Figure 12-113. Impedance Characteristics of Open-End Resonant Lines

like a rubber ball. In addition, at whatever angle the waveform arrives, it will bounce back from the line end at exactly the same angle (with respect to a line drawn perpendicular to the two-wire transmission line). This is illustrated in figure 12-112, where angle A equals angle B.

12-296. RESONANT LINES.

12-297. GENERAL. A resonant line is defined as a two-wire line of finite length which is not terminated at its output end with a load equal to the line's characteristic impedance. The line may be closed (shorted) or open-ended. The resonant line will contain standing waves of both voltage and current. A resonant line acts the same as a tuned circuit containing a physical coil and capacitor. The resonant line will have a high or low resistive characteristic impedance at multiples of a quarter wavelength from the input. Whether this impedance is high or low at multiples of a quarter wavelength will depend on whether the line is open- or short-circuited.

12-298. At any point along a line that is not an exact multiple of a quarter wavelength, the line will act as either a coil or a capacitor. The series resonant line, in common with a lumped inductance and capacitance circuit, will contain a voltage rise across the individual reactive circuit elements, but will have a low impedance and consequent high current across the entire resonant circuit. A parallel resonant line, in common with a lumped inductance and capacitance circuit, will contain a current rise across the individual internal reactive circuit elements and a current decrease with respect to the over-all parallel circuit. As seen externally, the parallel circuit will have a high impedance and consequent high voltage across the entire resonant circuit.

12-299. OPEN-ENDED LINES. The open-ended resonant line may be better understood by viewing figure 12-113. Only one-half wavelength is represented in the figure because the next half wavelength will represent the exact same impedance values along its length as does the first half wavelength. All succeeding wavelengths of this same frequency will be exact replicas of the first wavelength. Therefore, the study of any part of an open-ended resonant line which is one-half wavelength long will supply the same data as the study of the entire line. In figure 12-113 the lines are represented as having no resistive losses. The voltage is shown as zero across a minimum voltage node (representative short) and as maximum amplitude across a maximum voltage loop (representative open); the current is shown as zero across a minimum current node (representative open) and as maximum across a maximum current loop (representative short). In a practical line, rather than in the theoretical lines illustrated in figure 12-113 and discussed in the following paragraphs, you will always have line losses, because no line is a perfect conductor and the dielectric material contained between the wires is never a perfect insulator.

12-300. In the half-wavelength open-ended line illustrated in figure 12-113, the voltage and current amplitudes will exhibit definite characteristics at each quarter-wavelength, as measured back from the open-ended output terminals of the line. At all odd quarter-wavelengths ($\lambda/4$, $3\lambda/4$, $5\lambda/4$, etc), the voltage and impedance are minimum and the current is maximum. In addition, there is always a resonant rise of voltage from the odd-quarter-wavelength point toward the output (open end) of the line. This voltage gradually increases as the output (open end) of the line is approached, and it becomes maximum across the open end. Therefore, from this discussion, you can deduce that, at all odd quarter-wavelengths from the output end of a two-wire open-ended transmission line, the line will act as a series resonant circuit. Conversely, at all even multiples of a quarter-wavelength ($\lambda/2$, λ, $3\lambda/2$, etc) from the output (open end) of the transmission line, the line will act as a parallel resonant circuit. At these points the impedance and voltage will be maximum and the current will be minimum.

12-301. The discussion concerning an open-ended resonant line described the points along the line where a series or parallel resonant circuit condition existed. However, as illustrated in figure 12-113, between these points representing a resonant circuit (quarter-wavelength points), the line represents either an inductance or a capacitance value. The line may actually be used to replace an equal value of lumped physical inductance (coil) or capacitance (capacitor). Any open-ended line of less than a quarter-wavelength acts as a capacitance; between one-quarter and one-half wavelengths, as an inductance; between one-half and three-quarter wavelengths, as a capacitance; between three-quarter and one full wavelength, as an inductance; then the entire reactance cycle is duplicated for the next full wavelength; and so forth.

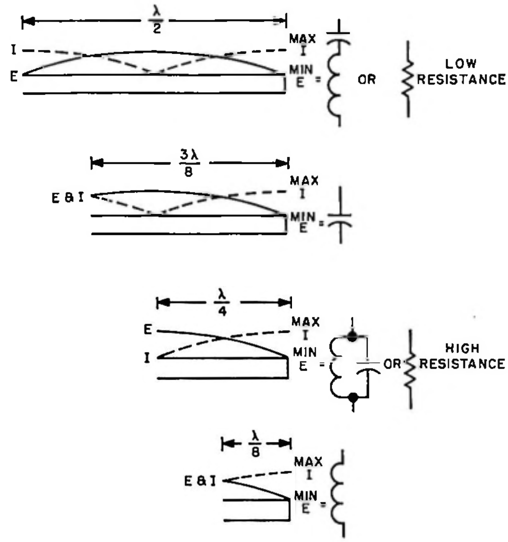


Figure 12-114. Impedance Characteristics of Shorted Resonant Lines

12-302. CLOSED-END LINES. The closed (shorted) end lines shown in figure 12-114 may be studied to better understand the operation of this type line. Again, only one-half wavelength is illustrated because all other half-wavelengths of a given frequency will exhibit the same impedance characteristics. This fact is discussed in greater detail in paragraph 12-299. The voltage and impedance across the output shorted end of the line will naturally be extremely low, and the current across this point will be maximum. At odd quarter-wavelengths, back toward the input source, from the shorted end, the voltage and impedance will be maximum and the current minimum. As this is the condition of a parallel resonant circuit composed of lumped reactance elements, the line at odd quarter-wavelengths from the shorted end will act as a parallel resonant circuit. A shorted two-wire transmission line composed of even multiples of a quarter-wavelength will act as a series resonant circuit because the voltage and impedance will be minimum and the current maximum at these points. When the line is not exactly an even multiple of a quarter wavelength, it will act like an inductance (coil) or a capacitance. The differences between an open-ended line and a closed-end line result from the fact that the standing waves of voltage and current switch places; at those points along the line where the voltage is maximum and the current minimum, the current will become maximum and the voltage minimum as you change from the concept of open-circuited to short-circuited line theory, or vice versa.

12-303. RESISTANCE-TERMINATED LINE. A two-wire transmission line with a pure resistance load, equal to the characteristic impedance of the line, connected across the output terminals of the line, will not produce reflections; therefore, no standing waves will be present on that line.

12-304. REACTANCE-TERMINATED LINE. When a two-wire line is terminated in a reactance, whether equal or not equal to the characteristic impedance of the line, there will be reflections and, consequently, standing waves on the line. When the line is terminated in a capacitance load, the line has the characteristics of an open-end line except that the voltage and current nodes (minimum points) are shifted closer and closer to the output end of the line as the

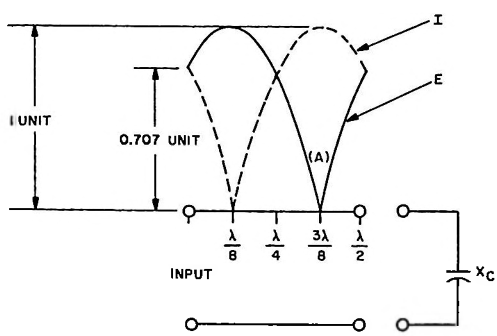


Figure 12-115. Two-Wire Line Terminated in X_C Equal to Characteristic Impedance of Line

capacitive reactance decreases. This fact is illustrated in figure 12-115. As you can see at point (A), the point of maximum current (I) and the point of minimum voltage (E) do not coincide with $\lambda/4$, as it would for an open-ended line. When a two-wire transmission line is terminated in an inductance, as illustrated in figure 12-116, the line has the characteristics of a closed-end line, except that the current and voltage nodes are shifted closer and closer to the output end of the line as the inductive reactance increases. As you can see in figure 12-116, point (A) does not coincide with the $\lambda/4$ point along the line.

12-305. STANDING WAVE RATIO. The standing wave ratio (swr) of a two-wire transmission line represents the ratio of the characteristic impedance of the line to the impedance of the load, or vice versa. If the line is terminated with a pure resistive load which equals the characteristic impedance, there will be no reflected (standing) waves on the line; the swr would be unity (1) because all of the energy sent from the input source through the line would be absorbed by the resistive load at the output. However, when the line is not terminated in a resistance value which exactly matches the characteristic impedance of the line, the

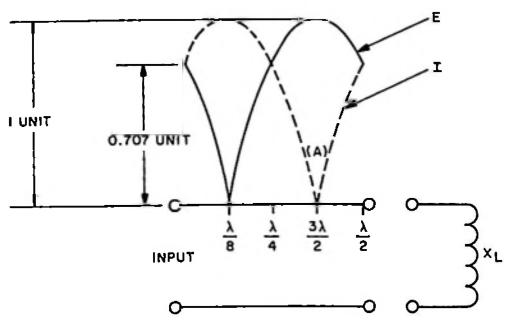


Figure 12-116. Two-Wire Line Terminated in X_L Equal to Characteristic Impedance of Line

line is said to be mismatched. When a mismatch is present, a standing wave ratio will exist. The swr can be calculated by dividing the amplitude of the standing wave representing minimum current into that representing maximum current (voltage amplitude may also be used); the swr can also be calculated by dividing the value of the terminating load into the value of the characteristic impedance of the line. Considering that the higher the swr the greater will be the mismatch between line and load, for proper operation of most transmission lines you must decrease the swr and, consequently, the mismatch. A knowledge of the exact position of the voltage or current loops (maximum amplitude point) and the voltage or current nodes (minimum amplitude point), with respect to the output end of the line, is necessary to determine whether the load is greater or less than the characteristic impedance of the line. For example, if the output load impedance is greater than the characteristic impedance of the line, an rf measuring device will indicate a voltage amplitude which is greater than the voltage nodes. In addition, the current reading at the load will measure less than the current measured at a current loop. Therefore, you can state that the line is showing the characteristics of an open-circuited line, with maximum voltage and minimum current at the load. The converse is true if the output impedance of the line is less than the characteristic impedance of the line.

12-306. When the load is greater than the characteristic impedance and the line exhibits the appearance of an open-circuited line, you know that the load is primarily capacitive. Therefore, you can shorten the line to decrease the capacitive load, or you can add series inductance to the load, in the form of a physical coil, to cancel part or all of the capacitance effects. When the load is less than the characteristic impedance and the line exhibits the appearance of a shorted line, you know that the line is primarily inductive. Therefore, you can lengthen the line to decrease the inductive reactance, or you can add series capacitance to the load, in the form of a physical capacitor, to cancel the inductance effects.

12-307. TRANSMISSION LINE TYPES

12-308. There are five general types of transmission lines: (a) the parallel two-wire line, (b) the twisted pair, (c) the shielded pair, (d) the concentric (coaxial) line, and (e) the waveguide. The use of a particular type of line depends primarily on the frequency and the power to be transmitted and on the type of installation.

12-309. PARALLEL TWO-WIRE LINE. The parallel two-wire line, because of the simplicity of its construction and behavior, is normally used to illustrate and explain the impedance characteristics and distributed constants of transmission lines at high frequencies. However, because of its high radiation loss, especially in the vicinity of metallic (conductive) objects, and because the spacing between the wires must be only a small fraction of one wavelength, the parallel two-wire transmission line is rarely used at microwave frequencies.

12-310. RIBBON TWO-WIRE LINE. The two-wire ribbon type transmission line is so called because the two wires are imbedded in a solid low-loss dielectric throughout the length of the line. This is done to maintain constant spacing between the two wires and to exclude moisture or dirt from between the lines. The ribbon type line is largely used between a television receiver and its antenna terminals. It is commonly manufactured to have a characteristic impedance of either 75 or 300 ohms. As you would expect, the spacing between the wires in the 75-ohm line is closer than that in the 300-ohm line. The impedance will remain the same for any length of line provided that the line is terminated in a load equal to its characteristic impedance.

12-311. TWISTED-PAIR LINE. The twisted-pair line, as the name implies, is so called because the wires are twisted around each other. It is normally used as an untuned line because on a tuned line the insulation may be punctured by the high voltage occurring at the voltage loops (maximum voltage points along the line). The twisted-pair line is not used at the higher frequencies because of the large losses occurring in the rubber insulation. These losses increase when the insulation is wet. The characteristic impedance is normally around 100 ohms, depending on the type of insulation used.

12-312. SHIELDED-PAIR LINE. The shielded-pair line consists of two parallel conductors separated from each other and surrounded by a solid dielectric. These conductors are contained within a copper-braid tubing that acts as a shield. The entire assembly is covered with a rubber of flexible composition coating, to protect the line against moisture and friction. Outwardly, the line looks like an ordinary power cord for an electric motor. The copper shield is connected to ground or a reference potential; thereby, the entire line has a uniform potential from shield to the reference point. The conducting lines now have equal capacitance between each wire and the copper shield and is said to be balanced. Another advantage of the shielded-pair line is that pickup from stray fields is shielded from the conducting wires.

12-313. AIR COAXIAL LINE. The air coaxial (concentric) line is used at the ultrahigh frequencies. The air coaxial line has a wire mounted inside of, and coaxially with, a tubular outer conductor; the inner wire conductor may also be tubular. The inner conductor is insulated from the outer conductor by insulating spacers or beads, located at regular intervals. These spacers or beads are generally composed of pyrex, polystyrene, or other material with good insulating qualities and having a low loss at high frequencies. The primary advantage of the coaxial line is its ability to hold down radiation losses. Unlike the two-wire parallel transmission line with its electric and magnetic fields extending into space for relatively great distances, the coaxial line permits no electric or magnetic fields to extend outside the outer conductor, to pick up stray fields from other lines or objects. Therefore, the coaxial line, by confining its complete field in the space between the two conductors, acts as a perfectly shielded line. The disadvantages of the air coaxial line are that it must be kept short at extremely high frequencies to prevent excessive losses, and it must be kept dry to prevent leakage between the conducting wires. To prevent condensation of moisture, the line may be filled, in certain applications, with dry nitrogen under pressure. The nitrogen is used to dry the line when it is first installed, and a pressure of 3 to 35 pounds per square inch is maintained thereafter to ensure that any leakage will be outward.

12-314. SOLID COAXIAL LINE. The concentric (coaxial) line can also be made with the inner conductor of flexible wire, insulated from the outer conductor by a solid and continuous insulating material. This "solid" coaxial line can be made fairly flexible if the outer conductor is made of metal braid, but the losses in this type of line are relatively high.

12-315. WAVEGUIDES. The term waveguide is applicable to all types of transmission lines in the sense that they are used to direct or guide the energy from one point to another. When the term is used in this sense, the line could be a single conductor, two or more conductors, a coaxial line, a hollow metal tube, or a dielectric rod. However, common usage has limited the meaning of the word waveguide to mean a hollow metal tube or a dielectric transmission line. Waveguides may be classed according to cross section (rectangular, elliptical, or circular) or according to material (metallic or dielectric). Dielectric waveguides are rarely used, because the losses for all presently known solid dielectric materials are too great for efficient transmission.

12-316. In the classification of hollow tube waveguides by cross section there were three types: rectangular, elliptical, and circular. Of these three, the rectangular cross-section type is most commonly used. Elliptical waveguides are not normally used because of the difficulty in fabrication, joining, and bending. Circular waveguides are rarely used because it is difficult to control the plane of polarization and the mode of operation. Circular waveguides also involve the difficulty of joining curved surfaces when a junction is required. However, they do find usage in rotating joints because of their circular symmetry, both physically and electrically.

12-317. The advantages of hollow waveguides are numerous. A hollow waveguide has lower losses in its practical frequency range than any other type of transmission line. Transmission lines, other than the hollow waveguide, have three kinds of losses: radiation, dielectric, and direct resistance. The hollow waveguide is a perfectly shielded line and therefore has no radiation loss. It is filled with air and thereby has reduced dielectric loss. Since the current at high frequencies travels on the surface skin of the conductor, a hollow waveguide provides a large inner surface area for the current to travel through. Therefore, the copper losses of the other transmission lines are greatly reduced in the hollow waveguide type line. The hollow waveguide is simpler in construction than the coaxial line because the inner conductor and its supports are eliminated. Because there is no inner conductor which may be displaced or broken by vibration or shock, the hollow waveguide is much more rugged than the coaxial line.

12-318. The primary disadvantages of hollow waveguides are their minimum size requirements and difficult installation. The minimum size of the waveguide that can be used to transmit a particular frequency is proportional to the wavelength at that frequency. This proportionality depends on the shape of the waveguide and the manner in which the electromagnetic fields are set up within the pipe. In all cases, there is a minimum frequency that can be transmitted. The lowest cutoff frequency is determined by the dimensions of the hollow inside portion of the waveguide. The wide dimension across the inside open end of the waveguide determines the cutoff frequency wavelength. The wavelength corresponding to the cutoff frequency is equal to twice this wide dimension. Higher frequencies can be transmitted, and the width of the guide for these frequencies is greater than their corresponding free-space full-wavelengths. The narrow dimension measurement across the hollow inside face of the open end of the waveguide is not critical with respect to

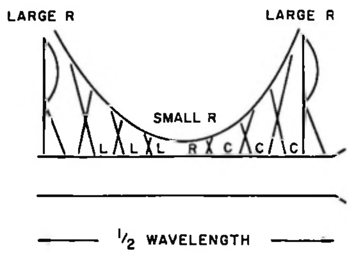


Figure 12-117. Impedance Characteristics of an Open-Ended Line

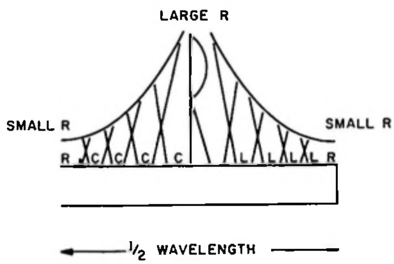


Figure 12-118. Impedance Characteristics of a Shorted Line

frequency. However, this distance determines the voltage level at which the waveguide arcs over. Therefore, for large power and voltage, this narrow dimension must be large. Because the wavelength of the cutoff frequency is only twice the length of the wide dimension of the inside of the waveguide, you normally will not use a hollow waveguide at frequencies below 3000 megacycles (10 centimeters). At lower frequencies, the guide would be too wide for practical application. For example, for 10-meter waves, the waveguide would be 23 feet wide.

12-319. The installation of hollow waveguides is more difficult than the installation of other types of transmission lines. The radius of bends in the hollow waveguide must be greater than two wavelengths to avoid excessive attenuation, and the cross section in both the wide and narrow dimension must be maintained uniform around the bend (elbow). These difficulties hamper installation in small spaces, and, if the guide becomes dented or if solder is permitted to run inside the joints, the attenuation of the line is greatly increased. In addition to the increased attenuation and the resulting standing waves, dents and beads of solder also reduce the breakdown voltage at the waveguide.

12-320. IMPEDANCE-MATCHING.

12-321. OPEN-ENDED LINE. The impedance of a half-wave section of transmission line, open at one end, varies widely over its length, as shown in figure 12-117. At the open end the resistance and voltage are high and the current is low. One quarter-wavelength back from the open end, the resistance and voltage are low and the current is large. One half-wavelength back from the open end, the resistance and voltage are again high and the current is low. However, as shown in figure 12-117, resistive opposition to the flow of current only occurs at exactly the point where a quarter-wave begins or ends; between these two points the opposition to the flow of current is composed of reactive components. The impedance curve of figure 12-117 illustrates that, from the open end of the line, capacitive reactance provides the current opposition back to a quarter-wavelength point. From a quarter-wavelength point back from the open end to a half-wavelength back from the open end, the opposition to the current is primarily inductive reactance.

12-322. CLOSED-END LINE. The impedance characteristics of a half-wavelength of transmission line shorted at one end also vary widely over its length, as shown in figure 12-118. The impedance at the end points of the line and at exact quarter-wave multiples is still composed of resistance, the same as explained in paragraph 12-321 for open-ended lines. However, the re-

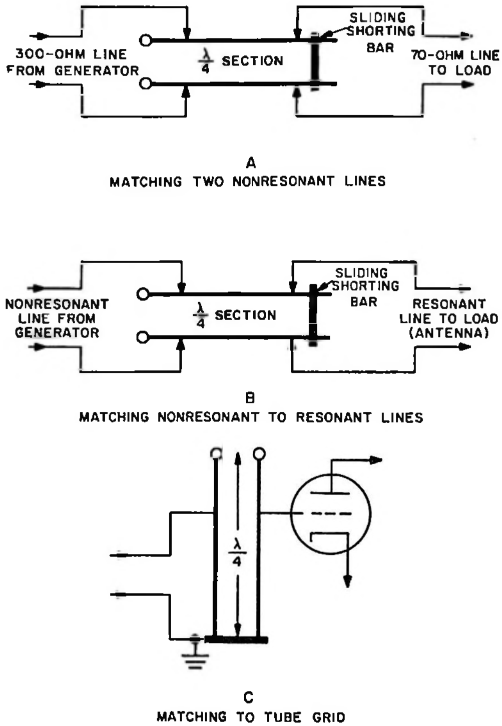


Figure 12-119. Transmission Line as an Impedance-Matching Device

sistance and the voltage are low and the current is high at the shorted end of the line. As you progress back from the shorted end of the line toward the first quarter-wavelength point from the shorted end, the impedance to the current is inductive reactance. Progressing further back the line produces capacitive reactance to the flow of current back to a half-wavelength from the shorted end of the line.

12-323. When either the open- or closed-end line is excited by rf energy, it is possible to match almost any impedance somewhere along the line. For example, a 300-ohm line can be matched to a 70-ohm line without producing reflections (standing waves) on either line. Figure 12-119 shows several methods of matching impedances.

12-324. A half-wave section of line shorted at both ends can be used as an impedance-matching device, because both ends of the line will have low voltage and impedance and high current whereas the center of the line, one quarter-wavelength from either end, will have maximum voltage and impedance and the minimum current. Therefore, the line can be tapped at various energy and impedance levels along its length. An open-ended resonant line can be used as an impedance-matching device, as shown in figure 12-120. The line must have the correct characteristic impedance if it is to match two dissimilar impedances. The necessary characteristic impedance, Z, of a quarter wave matching section is $Z = \sqrt{Z_1 Z_2}$.

For example, to match line 1 of figure 12-120 to line 2 when line 1 is 300 ohms and line 2 is 70 ohms, you take the square root of the product of both line impedances

$$(\sqrt{300 \times 70} = 145).$$

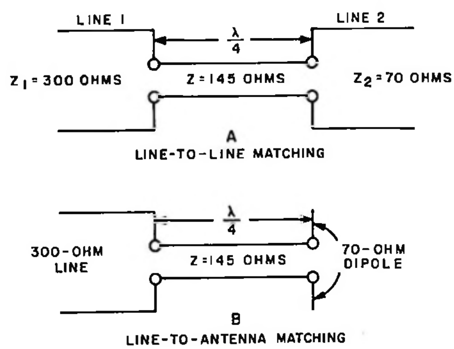


Figure 12-120. Impedance Matching with Open Quarter-Wave Line

12-325. MEASUREMENTS.

12-326. GENERAL.

12-327. The determination of the exact components required to replace a microwave structure at a particular frequency depends on whether the structure is of the termination or the transmission type. The termination type normally has a fixed input impedance which can be represented by a two-terminal equivalent circuit. The transmission type normally consists of multi-terminal microwave structures which must be reduced to successive measurements on four terminals (two-terminal pairs). To limit or minimize the attendant possible error of a single measurement on multi-terminal structures, you should make three input measurements, corresponding to three different known output terminations. However, you can take a number of different input-output measurements to analyze as a set in order to determine the equivalent circuit components required for replacement. To adequately present the subject of measurements of microwave frequencies, three-point, semi-precision, and precision methods of both measurement and analysis will be discussed in the following paragraphs.

12-328. TWO-TERMINAL STRUCTURES.

12-329. The properties of a two-terminal structure, with a dominant mode waveguide input, can be determined from a single-point measurement. You can obtain the impedance value of the termination directly if you use the impedance bridge method of measurement. However, if you employ a standing wave detector to measure the swr and find the first point of minimum voltage from the input terminals, and measure the distance from this minimum voltage point to the reference plane at the input terminals, you can calculate the impedance value from the measured quantities. The two-terminal circuit can also be specified by either its admittance or reflection coefficient. Generally, the equivalent circuit is determined from either the impedance or admittance characteristics because of the similarity and consequent familiarity with low-frequency circuits. The reflection coefficient is normally used to specify the characteristics of the two-terminal circuit only when you desire to change from one reference plane to another; in this case, only the phase of the reflection coefficient at the input would be changed.

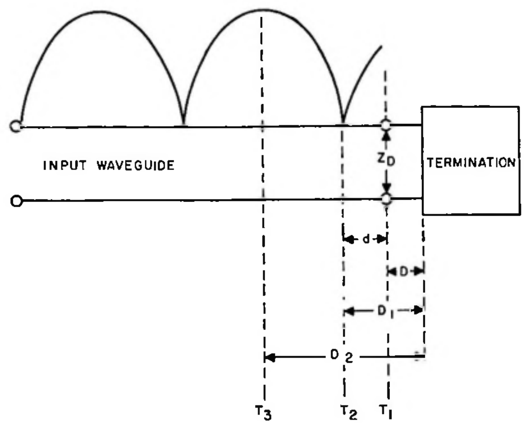


Figure 12-121. Standing Wave Pattern Symbol Location

12-330. The following definitions of symbols will aid your understanding of the text associated with impedance and distributed constants measurements at microwave frequencies; figure 12-121 illustrates the use of some of these symbols.

SWR = Standing wave ratio (voltage)

D = Location of input reference plane (T_1)

D_1 = Location of voltage minimum

D_2 = Location of voltage maximum

$d = D_1 - D$

Γ = Reflection coefficient at D

ϕ = Phase of Γ at D

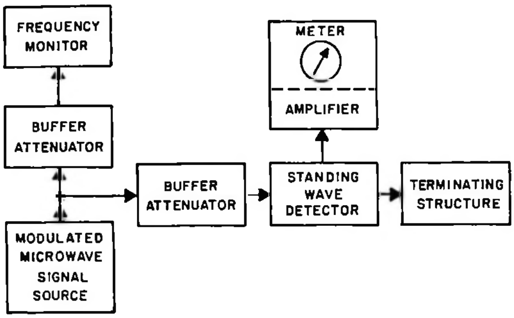


Figure 12-122. Slotted-Section Method of Measuring Input Impedance of a Two-Terminal Structure, Test Equipment Arrangement

Y_1 = Characteristic admittance of input guide

$Z_1 = 1/Y_1$ = Characteristic impedance of input guide

Z = Absolute input impedance

$Z_D = Z/Z_1$ = Relative input impedance at D

R = Absolute input resistance

$R_D = R/Z_1$ = Relative input resistance at D

X = Absolute input reactance

$X_D = X/Z_1$ = Relative input reactance at D

12-331. SLOTTED-SECTION METHOD. Figure 12-122 illustrates an arrangement using a movable-probe standing wave detector for impedance measurements. The square-wave modulated signal source is isolated from the waveguide by a buffer attenuator circuit which isolates the detector from excessive output power and prevents frequency shifting with changes in loading. The detector output is applied to an amplifier calibrated for the range of swr readings required. The frequency monitor can be a transmission type cavity tuned to the measurement frequency; the cavity output is applied through a crystal detector to a galvanometer set for maximum deflection at the appropriate frequency. This test method is performed as follows:

a. Find the length of a half-wavelength on the guide ($\lambda g/2$), by moving the standing wave detector probe back and forth along the guide until two successive points of minimum voltage are located. The physical distance between these two minimum voltage points equals one half-wavelength.

b. Determine the input reference plane (T_1), by terminating the output end of the slotted waveguide section with a shorting bar, and find this point of minimum voltage with the detector probe. This location is known by the symbol D and should be recorded.

c. With the output remaining as a short, determine both the swr value (discussed in paragraph 12-305) and the first point of minimum voltage (D_1) from the input reference plane (D); see figure 12-122. In an accurate measurement, all directly read values of D and swr must be corrected, by using a calibration curve which takes into account residual discontinuities in the detector (for example, the end of the slot in the slotted section).

12-332. RESONANCE METHOD. The resonance method of determining the input impedance of a two-terminal structure is normally employed with a coaxial waveguide, and uses a fixed probe in a variable length of line or a fixed line length with a variable frequency. The test equipment arrangement to perform this measurement is shown in figure 12-123. A fixed short-circuit plunger is slid along the line, and the resultant power changes on the line are picked off by the fixed probe; this changing power traces out a resonance curve whose properties can be analyzed the same as for a

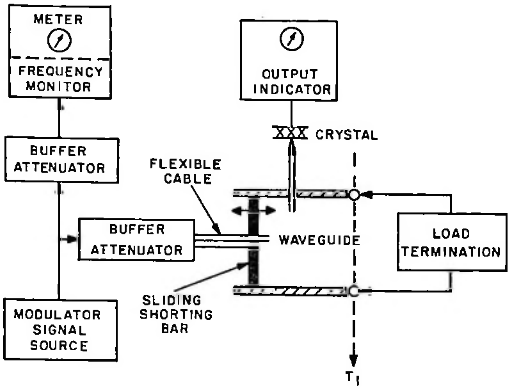


Figure 12-123. Resonance Method of Measuring Input Impedance of a Two-Terminal Structure, Equipment Arrangement

standing wave curve. With a constant voltage amplitude input supplied by the modulated signal source, the impedance of the termination can be obtained from the ratio of maximum to minimum power amplitudes and the voltage minimum (D_1).

12-333. The procedure for determining the input impedance of a two-terminal structure by the resonance method is as follows:

a. While the shorting bar is connected across the end terminals of the guide, at the position denoted as T_1 on figure 12-123, the plunger (detector probe) position is varied along the guide, to determine the maximum and minimum amplitudes of the resonance curve (P_{max} and P_{min}). The first minimum amplitude back from the shorting bar is specified by the symbol D_1 (L_1).

b. Replace the shorting bar with the device being measured, and again slide the probe along the guide to obtain P_{min} and P_{max}. The new distance from the shorting bar to the first minimum amplitude is also specified by the symbol D_1 (L_2). The swr can be calculated from the minimum and maximum amplitudes, as was discussed in paragraph 12-305, and the value of d represents $L_1 - L_2$, the distance from the short to the first minimum amplitude minus the distance from the device being measured to the first minimum amplitude. This data will permit the calculation of the input impedance.

12-334. FOUR-TERMINAL STRUCTURES.

12-335. The four-terminal structure, unlike the two-terminal structure which has two input terminals but contains a load or short across the output, has both its input and output terminals accessible. The input and output electrical qualities of the four-terminal structure may be calculated by many different but equivalent representations. However, this discussion will center around only four of the possible representations: impedance, admittance, scattering, and the tangent relationship.

12-336. IMPEDANCE REPRESENTATION. The impedance representation relates the input and output voltages with the input and output currents, to obtain the input and output impedance values. Figure 12-124 illustrates a four-terminal tee circuit impedance representation. The input and output impedances are calculated as follows:

$$V_1 = Z_{11}I_1 + Z_{12}I_2$$

$$V_2 = Z_{12}I_1 + Z_{22}I_2$$

$$Z_{out} = -\frac{V_2}{I_2}$$

$$Z_{in} = \frac{V_1}{I_1} = Z_{11} - \frac{Z_{12}^2}{Z_{22} + Z_{out}}$$

where:

V_1 = Voltage input across one set of terminals

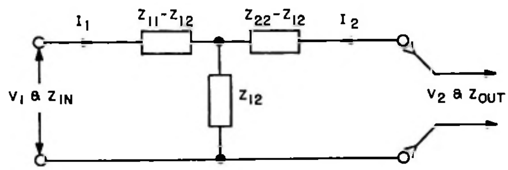


Figure 12-124. Four-Terminal Tee Circuit Impedance Representation

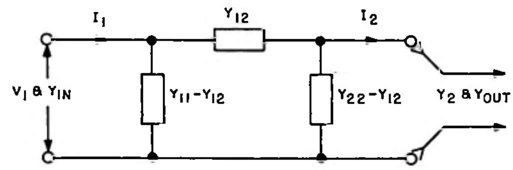


Figure 12-125. Four-Terminal Pi Circuit Admittance Representation

I_1 = Current caused as a result of V_1

V_2 = Voltage input across other set of terminals

I_2 = Current caused as a result of V2

Z_{in} = Input impedance

Z_{out} = Output impedance

Z_{11}, Z_{12}, and Z_{22} = Four-terminal structural impedances

Z_{11} = Impedance at input with output open

Z_{12} = Ratio of receiving voltage at input with input open, to driving current at output

Z_{22} = Impedance at output with input open

12-337. ADMITTANCE REPRESENTATION. The relations for the admittance representation are shown in figure 12-125. The admittance is calculated in exactly the same fashion as the impedance representation calculation of paragraph 12-336.

$$I_1 = Y_{11}V_1 + Y_{12}V_2$$

$$I_2 = Y_{12}V_1 + Y_{22}V_2$$

$$Y_{out} = -\frac{I_2}{V_2}$$

$$Y_{in} = \frac{I_1}{V_1} = Y_{11} - \frac{Y_{12}^2}{Y_{22} + Y_{out}}$$

where all symbols are defined in paragraph 12-336, except the following:

Y_1, Y_2, and Y_3 = Four-terminal structure admittances

Y_1 = Admittance at input with output shorted

Y_{12} = Ratio of receiving current at input, with input shorted, to driving voltage at output

Y_{22} = Admittance at output with input shorted

12-338. SCATTERING REPRESENTATION. The voltage-scattering method relates the incident and reflected parts of the total input and output voltages. Figure 12-126 illustrates the scattering circuit for the four-terminal impedance representation.

$$\left| S_{11} \right|^2 = \left| S_{22} \right|^2 = 1 - \left| S_{12} \right|^2 \quad \text{(For lossless structure only)}$$

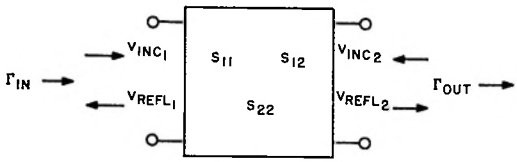


Figure 12-126. Four-Terminal Circuit Scattering Representation

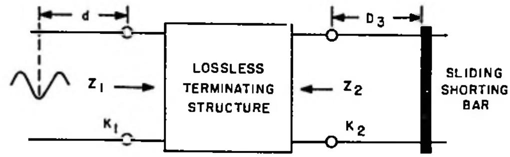


Figure 12-127. Four-Terminal Circuit Tangent Relation Representation

$$V_{refl_1} = S_{11}(V_{inc_1}) + S_{12}(V_{inc_2})$$

$$V_{refl_2} = S_{12}(V_{inc_1}) + S_{22}(V_{inc_2})$$

$$\Gamma_{out} = \frac{V_{inc_2}}{V_{refl_2}}$$

$$\Gamma_{in} = \frac{V_{refl_1}}{V_{inc_1}} = S_{11} - \frac{S_{12}^2}{S_{22} - \dfrac{1}{\Gamma_{out}}}$$

where:

V_{refl_1} = Reflected voltage wave at first set of terminals

V_{inc_1} = Incident voltage wave at first set of terminals

V_{refl_2} = Reflected voltage wave at second set of terminals

V_{inc_2} = Incident voltage wave at second set of terminals

Γ_{in} = Input reflection coefficient

Γ_{out} = Output reflection coefficient

$| \; |$ = Vertical lines on each side of symbol means absolute magnitude of that symbol; plus or minus signs in connection with absolute magnitude have no meaning and are not used.

12-339. TANGENT RELATION REPRESENTATION. In many waveguide measurements, the input and output characteristics are not directly available for measurement; they must be computed from the value of the swr and the location of the first voltage minimum point from the output termination. For this discussion, let us assume that a lossless termination consists of a lossless waveguide terminated with a movable sliding shorting bar. A circuit based on these assumptions is illustrated by figure 12-127; D_3 represents the distance between the output terminals and the shorting bar, and d, as defined in paragraph 12-330, represents the distance from the input terminals to the first point of minimum voltage on the line. Z_1 and Z_2 represent the characteristic impedance of their respective lines, and symbols K_1 and K_2 are equal to 2π divided by the wavelength on the input and output line, respectively.

12-340. MEASUREMENT UTILITY.

12-341. The two-terminal circuits discussed in paragraphs 12-329 through 12-333 and the four-terminal circuit discussed in paragraphs 12-335 through 12-339 are represented both illustratively and mathematically in the text. The impedance (tee) and admittance (pi) circuit illustration and the mathematical concepts discussed in the text are desirable representations because of both their familiarity to you and their relative ease of calculation and measurement. However, for lossless structures, the circuits and formulas of figure 12-128 are just as convenient. The total impedance of the four-terminal structure is denoted by Z, and the total admittance by Y.

12-342. Sometimes it is desirable to shift the reference plane to one of the other representations. If you shift the input and output reference planes of a known circuit by given amounts, the scattering coefficients of the circuit, illustrated in figure 12-128, between the shifted reference planes are related, by a phase shift, to the original circuit scattering coefficients. The scattering representation is also useful for a circuit terminating in a load which matches its characteristic impedance. When a match is effected, the output reflection coefficient (Γ_{out}) is reduced to zero because the input reflection coefficient (Γ_{in}) is now equal to the scattering coefficient (S_1). The input impedance (Z_{in}) can be determined from the simplified formula:

$$Z_{in} = \frac{Z_1(1 + \Gamma_{in})}{(1 - \Gamma_{in})}$$

The insertion swr then becomes

$$swr = \frac{1 + |S_{11}|}{1 - |S_{11}|}$$

In addition, the insertion loss can be directly determined from S_2, as follows:

$$\text{Insertion Loss} = - 20 \text{ Log}_{10} |S_{12}| \quad \text{(in db)}$$

12-343. The tangent relation representation is particularly useful if circuit representations at different reference planes are to be obtained for lossless four-terminal structures. For example, in the circuit of figure 12-129, the matched load on one pair of terminals causes the input impedance (Z_{in}) on the opposite pair of terminals to equal Z_1 times the insertion swr of the input guide. These tangent relations hold true for the measured characteristic impedance representations if $Z_1 = Z_2$.

12-344. TECHNIQUES—LOSS STRUCTURES.

12-345. FOUR-TERMINAL STRUCTURES. As shown in figure 12-130, the input end of the four-terminal structure is connected to a standing wave detector, and the output end of the structure is connected to the terminating impedance. In addition, an impedance bridge as illustrated in figure 12-131 can also be used. One method used for measuring this general four-terminal structure is as follows:

a. Locate the input reference plane (T_1) by placing a shorting bar across the output of the standing wave detector and recording the distance (D) from the shorting bar to the first point of minimum voltage. Record the value of this input guide half-wavelength ($\lambda/2$).

b. Connect the output of the detector to the input of the four-terminal structure to be measured, and connect the shorting bar across the output of this structure. You can now measure the short-circuit value of the voltage minimum (D_1) and the swr; from this data the short-circuit output impedance (Z_{out}) can be calculated by the mathematical formulas contained in paragraph 12-336 because $d = D_1 - D$. If a bridge arm is used rather than a detector, Z_{out} can be measured directly.

c. Next, replace the shorting bar termination with an open-circuit termination, and repeat step b above to obtain Z_{out} for the open-circuit connection.

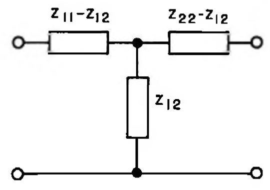
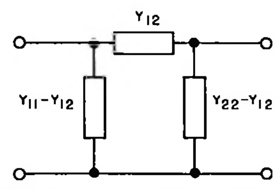
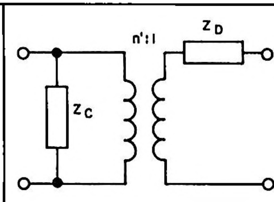
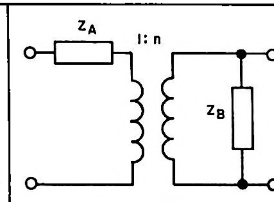
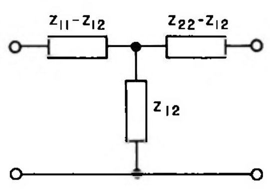
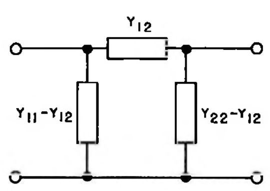
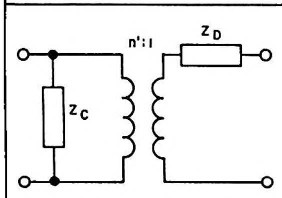
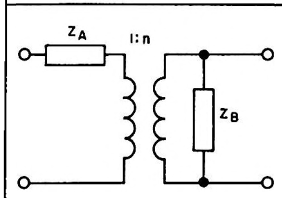
				
	$(Z) = Z_{11} Z_{22} - Z^2_{12}$	$Z_{11} = \frac{Y_{22}}{(Y)}$ $Z_{22} = \frac{Y_{11}}{(Y)}$ $Z_{12} = \frac{Y_{12}}{(Y)}$	$Z_{11} = Z_C$ $Z_{22} = Z_D + \frac{1}{n'^2} Z_C$ $Z_{12} = \frac{Z_C}{n'}$	$Z_{11} = Z_A + \frac{1}{n^2} Z_B$ $Z_{22} = Z_B$ $Z_{12} = \frac{Z_B}{n}$
	$Y_{11} = \frac{Z_{22}}{(Z)}$ $Y_{22} = \frac{Z_{11}}{(Z)}$ $Y_{12} = \frac{Z_{12}}{(Z)}$	$(Y) = Y_{11} Y_{22} - Y^2_{12}$	$Y_{11} = Y_C + \frac{1}{n'^2} Y_D$ $Y_{22} = Y_D$ $Y_{12} = \frac{Y_D}{n'}$	$Y_{11} = Y_A$ $Y_{22} = Y_B + \frac{1}{n^2} Y_A$ $Y_{12} = \frac{Y_A}{n}$
	$Z_C = Z_{11}$ $Z_D = \frac{(Z)}{Z_{11}}$ $n' = \frac{Z_{11}}{Z_{12}}$	$Z_C = \frac{Y_{22}}{(Y)}$ $Z_D = \frac{1}{Y_{22}}$ $n' = \frac{Y_{22}}{Y_{12}}$	$Z_C = \frac{1}{Y_C}$ $Z_D = \frac{1}{Y_D}$	$Z_C = Z_A + \frac{1}{n^2} Z_B$ $Y_D = Y_B + \frac{1}{n^2} Y_A$ $n' = \frac{Y_B + \frac{1}{n^2} Y_A}{Y_A} n$
	$Z_A = \frac{(Z)}{Z_{22}}$ $Z_B = Z_{22}$ $n = \frac{Z_{22}}{Z_{12}}$	$Z_A = \frac{1}{Y_{11}}$ $Z_B = \frac{Y_{11}}{(Y)}$ $n = \frac{Y_{11}}{Y_{12}}$	$Z_A = \frac{Z_C Z_D}{Z_D + \frac{1}{n'^2} Z_C}$ $Z_B = Z_D + \frac{1}{n'^2} Z_C$ $n = \frac{Z_D + \frac{1}{n'^2} Z_C}{Z_C} n'$	$Z_A = \frac{1}{Y_A}$ $Z_B = \frac{1}{Y_B}$

Figure 12-128. Determination of Impedance and Admittance Representations Between Two Pairs of Terminals

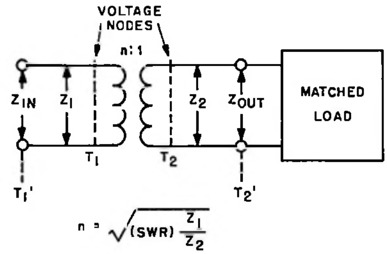


Figure 12-129. Matched Load Versus Reference Planes

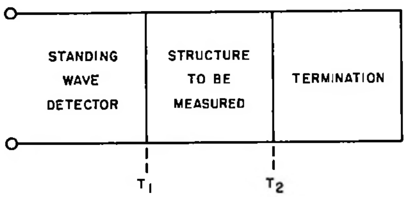


Figure 12-130. Measurement of a Four-Terminal Structure Using a Standing Wave Detector, Equipment Arrangement

d. Next, replace the open-circuit termination with a termination having a known value, and repeat step b above to determine the input impedance (Z_{in}).

NOTE

Rather than using a short circuit, an open circuit, and an actual impedance termination, all in sequence, to obtain the data required in steps a through d above, you can use a single variable shorting plunger. The plunger can be positioned at the output terminals or an even number of half-wavelengths away to represent a short circuit; it can be placed a quarter-wavelength back from the terminating line end to represent an open circuit; and it can be positioned between a quarter and a half-wavelength from the terminating end to represent an impedance termination.

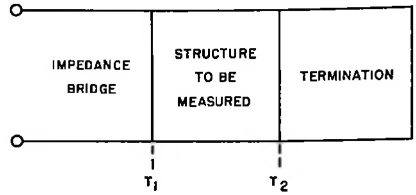


Figure 12-131. Measurement of a Four-Terminal Structure Using an Impedance Bridge, Equipment Arrangement

12-346. TEE REPRESENTATION. The values of Z_{out} (shorted), Z_{out} (open), and Z_{in}, obtained in paragraph 12-345 on a four-terminal structure, can be substituted in the equations contained in paragraph 12-336 to evaluate the equivalent Tee parameters.

12-347. PI REPRESENTATION. The values obtained in paragraph 12-345 can be substituted, the same as was done in paragraph 12-346, in the admittance equations contained in paragraph 12-337. However, the impedance values of paragraph 12-345 will have to be converted to admittance values—Y_{out} (shorted) and Y_{out} (open).

12-348. SCATTERING REPRESENTATION. If a scattering representation is desired, the same general procedures of paragraphs 12-345 and 12-346 must be followed; reflection coefficients would replace impedances and admittances representing short or open circuits. Further, the termination representing a known impedance, used to obtain Z_{in}, must now match the characteristic impedance (Z_2).

12-349. TECHNIQUES—LOSSLESS STRUCTURES.

12-350. FOUR-TERMINAL STRUCTURE. The tangent relation representation can be calculated as follows:

a. The value of D is determined as in paragraph 12-345, step a.

b. Connect a matched termination ($Z_{match} = Z_2$) to the output terminals of the structure to be measured, and determine the value of the swr and D_1 (L_1).

c. Next, reverse the structure and apply the matched load to Z_1; determine the new value of D_1 (L_2).

d. The remaining procedure is the same as was applied in paragraphs 12-345 and 12-346.

12-351. SEMI-PRECISION IMPEDANCE MEASUREMENT. Semi-precision measurements require the analysis of a complete set of data points simultaneously. The technique involved consists of the derivation of the reactance transformation diagram, tangent relationship method, dissipative Weissfloch circuits, distance invariant circuit, etc. In addition, if an error analysis is also carried out, the semi-precision method becomes a precision method. Considering that all complicated circuitry can be reduced to four-terminal structures, with the attendant ease of impedance calculation and measurement, the semi-precision and precision methods of impedance measurements are complex, cumbersome, and require a high degree of mathematical background. Therefore, these methods are beyond the scope of this publication.

12-352. TEST EQUIPMENT.

12-353. ADMITTANCE COMPARATOR. The admittance meter circuit shown in figure 12-132 uses a single voltage source to feed three separate coaxial lines; an individually adjustable wire loop, coupled to the magnetic field in each line, samples the current through that line. The outputs from the three loops are connected in parallel and applied to a detector. By adjusting the loops in each line until the paralleled outputs are balanced, a null (zero point) will be obtained on the indicator of the detector. This instrument can best be used if two of the three coaxial lines are terminated in known values and the third line is terminated with the unknown; a reading scale can be provided on the indicator portion of the detector to directly indicate the conductance and susceptance value of the unknown admittance. For example, in figure 12-132, one line contains the input voltage source; the second line is terminated with a resistor equal in value to the characteristic impedance of that line; the third line is terminated with a stub equal to 1/8 wavelength of the measurement frequency; the fourth line is terminated with the unknown; and the fifth line contains the detector-indicator.

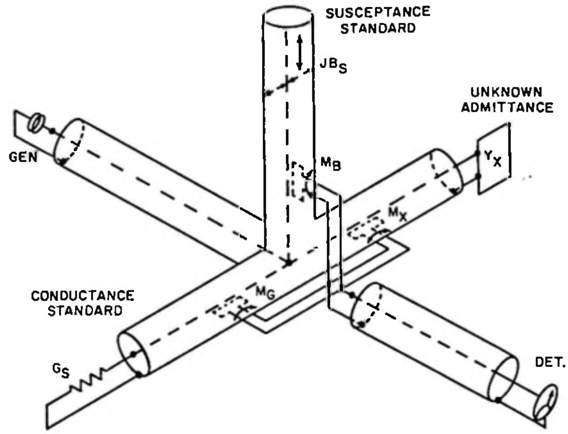


Figure 12-132. Admittance Meter Circuit

12-354. Mathematically speaking, the input current to each branch line depends on the input admittance of that line. Therefore, for the unknown line, the admittance can be mathematically represented as follows:

$$Y_X = \frac{M_b B_S - M_g G_S}{M_X}$$

where:

Y_X = Input admittance of unknown line

M_b = Mutual inductance of the 1/8 wavelength terminated line

M_g = Mutual inductance of the characteristic impedance terminated line

M_X = Mutual inductance of the unknown admittance line

B_S = Susceptance of the 1/8 wavelength terminated line

G_S = Resistance equal to the characteristic impedance of the line terminated in the conductance standard

For the 1/8 wavelength terminated line, the admittance is JB_S, which actually is $-J/Z_0$; for the line terminated with its characteristic impedance, the admittance is equal to the reciprocal of that impedance ($1/Z_0$). The conditions of paragraph 12-354 can be established only if the vector sum of the loop voltages is zero and the null condition is achieved.

12-355. In a practical measurement case, using the metering circuit of figure 12-132, the series inductances of the line lengths will introduce an error at the higher frequencies. This error can be eliminated by physically adding a shunt capacitance of the proper value to each line; the characteristic impedance (Z_0) will then equal the square root of the quotient of the inductance (L) and the shunt capacitance (C).

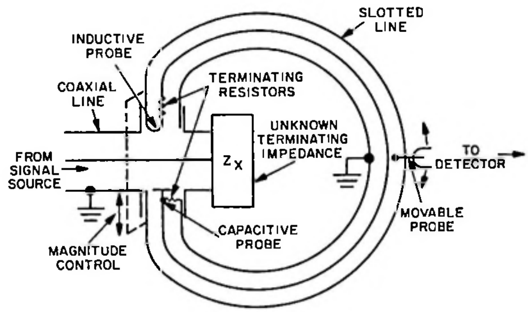


Figure 12-133. Byrne Bridge Used for Impedance Measurement in UHF Range

12-356. BYRNE BRIDGE. The Byrne bridge is a null-indicating device used to measure impedance in the uhf range. The bridge, formed with coaxial lines, is shown in figure 12-133. This null device uses an oscillator type source voltage input to a line which is terminated in the unknown impedance. At some point, a short distance back from the terminating end, a capacitive probe is used to sample the voltage, and an inductive probe (loop) is used to sample the current. These probes are located on opposite sides of the center conductor in such a fashion that, when one probe moves toward the center conductor, the other probe moves away. The terminating resistor at each probe matches the probe with the slotted line. Therefore, a wave started at one probe will be absorbed by the terminating resistor at the other probe. The indicating detector probe operates in the slotted line, to sample the energy at a balanced point.

12-357. The unknown impedance is measured by the Byrne bridge if you first adjust the capacitive probe for a null reading at the detector. The detector indicator dial can then be calibrated to read the unknown im-

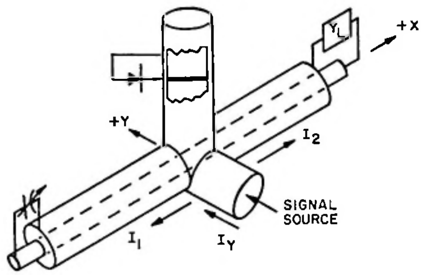


Figure 12-134. Standing Wave Indicator Used To Indirectly Measure Impedance at Microwave Frequencies

pedance magnitude directly, and is independent of frequency. However, the phase angle of the unknown impedance must be known. If the unknown impedance is a pure resistance, the detector probe should be midway between the current and voltage pick-off probes. However, for a reactive load, the null will occur to one side or the other of this mid position, and the distance of the null from the mid position will be directly proportional to the frequency of measurement. The phase control dial can be calibrated directly in phase angle for any given frequency. Then, if some other frequency is applied, a calculation would provide the new phase angle. For example, if the dial were calibrated for a null at 50 megacycles and you applied some other frequency (f), the phase angle would be calculated as follows:

$$\theta = \frac{f}{50 \text{ mc}} \text{ times (dial reading for null)}$$

12-358. STANDING WAVE INDICATOR. The standing wave indicator shown in figure 12-134 provides a direct reading in swr, r (the swr when the generator losses are negligible), and the phase angle of the reflection coefficient (Γ). With this data, provided by direct measurement, the equivalent impedance or admittance can be calculated by the formulas contained in paragraphs 12-335 through 12-343. The indicator shown in figure 12-134 consists of a coaxial tee junction with an oscillator source voltage applied to the short arm of the tee, the unknown admittance connected to one end of the cross arm of the tee, a standard capacitor connected to the opposite end of the tee cross arm, and a detector indicator connected into the tee through a short cutoff coaxial tube.

12-359. In figure 12-134, at the junction of the tee, the total current from the generator headed in the Y direction will consist of I_1 plus I_2. However, the individual current headed in the X direction will be the current called I_1; this is the current left over after I_2 is subtracted from the total. Therefore, the following equations hold true:

$$I_{total} = I_y = I_1 + I_2$$

$$I_x = I_y - I_1$$

12-360. The capacitor arm of the tee in figure 12-134 should be adjusted to obtain a susceptance of unity at the junction, as indicated on the detector meter. Then, if the detector probe is moved from a minimum to a maximum voltage point, the ratio between these two voltages will be the swr value. Next, if a line is drawn between the generator arm and the unknown impedance arm (45-degree angle from each), the angle of a counterclockwise rotation from the 45-degree reference line to the first point of minimum voltage is equal to the angle of the current reflection coefficient (Γ). The angle can therefore be read directly if the detector indicator dial is properly calibrated.

12-361. COMPLEX IMPEDANCE PLOTTER. This instrument, shown in figure 12-135, plots the unknown resistance and reactance directly on a calibrated chart.

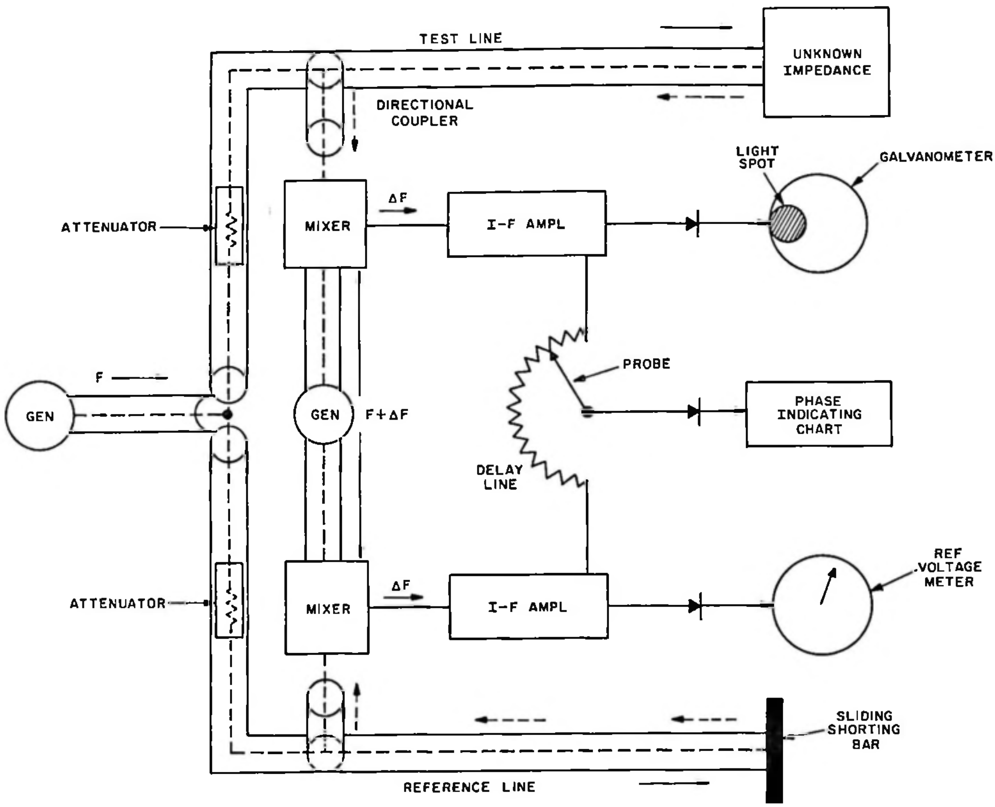


Figure 12-135. Complex Plotter Used for Impedance Measurement in Microwave Range

12-362. The impedance plotter illustrated in figure 12-135 functions as follows:

a. The test signal from the generator is applied through separate attenuators, to a test line terminated in the unknown impedance and to a reference line terminated in a short to produce total reflection. The incident waves traveling along both lines toward their respective terminations are equal. The reflection coefficient will be unity for the reference line, and this, in turn, causes the ratio of the amplitude of both reflected signals to be equal to the reflection coefficient at the termination of the test line. The reflected signals of both lines are coupled to the mixers, and the difference frequency (ΔF) is amplified. The amplitude of this difference frequency is displayed on the reference voltage meter. The reflected signal is also applied to a light spot galvanometer; the screen of the meter consists of a translucent impedance chart, and the light spot can be radially deflected from the center of this chart.

b. Adjust the input generator signal level and obtain a pre-determined setting on the reference voltage meter; since the light spot indicator is calibrated directly in terms of the reflection coefficient because the reflected signals are equal in amplitude, you

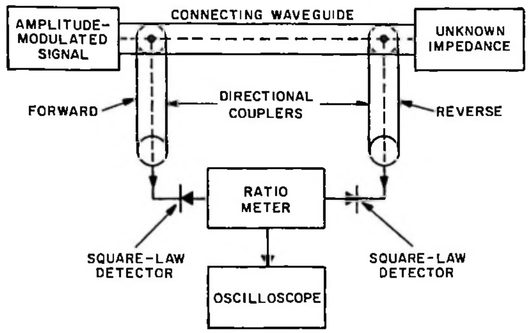


Figure 12-136. Reflectometer Equipment Arrangement for Impedance Measurement in Microwave Range

can read the swr directly from the calibrated chart.

c. Rotate the capacitance probe along the circular delay line until you reach the point where the two reflected signals (one from each amplifier) cancel and produce a null point. The phase-indicating chart is geared directly to the movement of the probe so that the phase angle can be read directly. If the indicating chart is of the Smith type, you can read the resistance and reactance directly.

12-363. REFLECTOMETER. The reflectometer measures the amplitude ratios of the incident signal versus the reflected signal. As shown in figure 12-136, an amplitude-modulated signal source, operating on a frequency acceptable to the ratio meter applies a signal to the unknown load; also, it applies a signal through a forward directional coupler and through a detector to the ratio meter. This source can be a swept generator if you supply an output from the generator to the horizontal sweep circuits of the oscilloscope. The majority of the incident wave power travels down the connecting guide to the unknown load as only a small part of the original power is lost to the forward coupler. A small part of the power, reflected from the unknown impedance load is picked off by the reverse directional coupler, detected, and applied to the ratio meter. If the detectors are barretter or silicon crystal types operated at low levels, they have square-law characteristics. If the ratio meter is calibrated on a square-law basis, it will permit direct readings. This impedance measurement method makes no allowance for the phase angle of the reflection coefficient; only magnitude is measured.

SECTION VIII

GAIN MEASUREMENT

12-364. GENERAL.

12-365. The term gain has a number of different, although related, meanings when it is used to describe the operation of electronic circuits. For example, it is used somewhat indiscriminately to describe ratios of power, complex ratios of transmission quantities (such as voltage or current), and the ratios of the magnitudes of these quantities. The resulting ambiguity leaves the term almost meaningless in a general discussion of measurements, unless it is carefully qualified or unless reference is made to agreed standards. Various professional groups have laid a foundation for resolving this confusion of meaning by setting standard definitions to the term gain and to certain alternate terms used to describe related concepts. The trend of these efforts is to define gain and the opposite term, loss, so as to restrict their application to power ratios only. The corollary terms, amplification and attenuation, which are frequently to be found in the literature used synonymously with gain and loss, are then defined as the numerical ratio of any two magnitudes of any quantity, such as voltage or current. By qualifying these four basic terms when necessary, various useful concepts may be given precise definitions. This greatly simplifies the specification of measurements of the quantities involved and also the standardization of the measurements. In the interest of clarity, the distinctions in meaning that have been outlined are observed consistently in the material that follows.

12-366. Gain and amplification are dimensionless quantities expressed by single numbers giving the power ratio or the magnitude ratio at a single frequency. For example, if the input or output power consists of more than one component, such as a multi-frequency signal or noise, the particular components used and their weighting must be specified. When these numbers apply at specified band-center frequencies or at specially designated reference frequencies, they are commonly referred to as absolute values or mid-band values. In the measurement of low-pass systems and components, it is customary to specify an arbitrary reference frequency at which absolute measurements shall be made. For example, in audio-frequency practice, the reference is commonly taken to be either 400 or 1000 cps.

12-367. Gain and loss are usually, although not necessarily, expressed in decibels by multiplying the common logarithm of the appropriate power ratio by ten. When so expressed, the result may be either positive or negative. Gain is the negative of loss, and, if the gain is positive, it is usually preferred to loss as a measure of the transmission. Amplification and attenuation of voltages are often desirably expressed in logarithmic form. By a special extension of the term decibel, which is a concession to firmly established usage, the magnitude ratio is expressed in decibels by multiplying its common logarithm by twenty.

12-368. Relative gain and relative amplification are frequency functions and express the respective ratio relative to an absolute reference value over the useful frequency range of the equipment or component under consideration.

12-369. GAIN AND GAIN-RELATED PROPERTIES.

12-370. Gain, in accordance with paragraph 12-365, describes a ratio of two amounts of power. The two power quantities may be those measured at two different points in a transmission equipment, or alternatively, at the same point, but under two different conditions of power transfer. In general, gain is the ratio of the power at the output of an equipment or component to a reference power, the reference being the chief distinguishing factor in the various concepts of gain which are described in the following paragraphs.

12-371. GAIN DEFINITIONS.

12-372. POWER GAIN. When not otherwise specified, the reference power is taken as the power at the input of the particular equipment under consideration. In a general discussion of this kind, there is a definite need for a specific term suitable for describing any one of a number of related devices such as circuits, transmission equipment, or components of equipment. The term transducer serves this purpose very well, being defined as a device capable of being actuated by waves from one or more transmission equipments or media and of supplying related waves to one or more transmission equipments or media. The term waves covers such general concepts as signal, power, energy, response, stimulus, voltage, current, etc. The waves in the input and output of a transducer need not be alike. Therefore, power gain when not otherwise modified, refers to the ratio of the power that a transducer delivers to a specified load under specified operating conditions to the power absorbed by its input circuit.

12-373. BRIDGING GAIN. An amplifier or other transducer is sometimes shunted, or bridged, across a part of an existing transmission path for the purpose of forming a branch path. In this case, interest centers in comparing the power at two different points in the equipment, but the reference is not the power absorbed by the input circuit of the transducer; instead, it is the power in the impedance across which the input is bridged. Bridging gain is therefore defined as the ratio of the power a transducer delivers to a specified load impedance under specified operating conditions to the power dissipated in the reference impedance across which the input of the transducer is bridged.

12-374. INSERTION GAIN. When an amplifier is inserted between a source and a load for the purpose of increasing the power in the load over that which the source alone could deliver, there are various ways of specifying the gain. In each case, the relevant concept is that of comparing the power in the load first in the reference condition, with the source alone connected to it, and then with the amplifier inserted. Several different conditions may exist depending upon the impedance relations both in the reference condition and at the input and output of the amplifier when it is inserted. In the general case, when no particular impedance relations are implied, the power ratio is simply called insertion gain. It is formally defined as the ratio of the power delivered to that part of the equipment following the transducer (which has been inserted in a transmission equipment) to the power delivered to that same part before insertion. Insertion gain is not the unique property of a transducer. It depends not only upon the transducer characteristics but also upon the source and load impedances between which the transducer is inserted. These must be specified when insertion gain is specified.

12-375. POWER AND THE TRANSDUCER. The amount of power which a source can deliver to a specified load depends upon the relation existing between the load impedance and the internal impedance of the source. It is a maximum when the two impedances are conjugate, in which case the impedances are said to be matched on a maximum-power-transfer basis (or, simply, matched). Under these conditions, the power is given approximately by the quotient of the mean square of the open-circuit voltage of the source and four times the resistive component of the source impedance. This power is independent of the load impedance, and is the maximum amount of power available from the source, regardless of whether or not the load can absorb it. The concept of available power finds many useful applications in connection with source ratings and particularly in signal-to-noise considerations, as has been explained previously in this chapter.

12-376. A useful concept in connection with available power is that of the ideal transducer. This is a hypothetical passive transducer, connecting a specified source to a specified load, which transfers the maximum possible power from the source to the load. Thus the ideal transducer dissipates no energy and when connected between a specified source and load presents to each its conjugate impedance. It should be noted that this concept differs from that of the ideal transformer.

12-377. TRANSDUCER GAIN. When the reference condition for gain is taken as the available power of the source, a special form of insertion gain results, called <u>transducer gain</u>. Transducer gain may be thought of as a comparison of the power in a specified load when the transducer under test is inserted to that when an ideal transducer is substituted for the transducer under test.

12-378. AVAILABLE GAIN. When the load impedance of an amplifier is matched to the amplifier output circuit on a conjugate basis, the amplifier delivers its available output power to the load. The ratio of the available power of the source is called the <u>available gain</u>. There is a still more restrictive form of insertion gain, but it remains a function of the impedance relation existing between the source and the amplifier input circuit. It is necessary therefore to include this condition in the definition. Available gain is then the ratio of the available power from the output terminals of the transducer <u>under specified input conditions</u> to the available power from the source. When the source is matched to the input circuit of the transducer, and the load is likewise matched to the output circuit, the available gain assumes its greatest value. This is sometimes called the <u>matched power gain</u>, because it is common practice to operate microwave equipments on a completely matched basis. When this is the case, the distinctions between available power, impressed power, and absorbed power vanish, as do the distinctions between gain concepts. At the higher radio frequencies, noise generated within the receiver determines the weakest usable signal, and both sensitivity and noise figure become interdependent measures of the weakest signal that can be received.

12-379. VOLTAGE AMPLIFICATION. <u>Voltage amplification</u> is defined as the ratio of the magnitude of the voltage across a specified load impedance connected to a transducer to the magnitude of the voltage across the input of the transducer. It is generally dependent upon the source and load impedances. Voltage amplification has a particular significance in connection with certain types of amplifiers in which both the load and the input impedances are effectively equivalent to open circuits. In this instance, the concept of power gain has no useful significance. Amplifiers of this kind, which are usually characterized as voltage amplifiers, commonly operate into the grid circuits of vacuum tubes where the grid-input impedances are effectively infinite.

12-380. Relative voltage amplification (amplitude-frequency response) is one of several important steady-state properties of amplifiers. Its importance is due, in part to the fact that it can be measured with relative ease compared with other steady-state properties. Among these others, which are sometimes more significant measures of amplifier performance, are phase shift, time delay, and the frequency derivative of phase shift. They are difficult to measure, particularly when broad bands of frequency must be spanned.

12-381. CURRENT AMPLIFICATION. Current amplification is defined as the ratio of the magnitude of the current in a specified load impedance connected to a transducer to the magnitude of the current in the input circuit of the transducer.

12-382. SENSITIVITY, POWER.

12-383. In certain applications, a significant property of a transducer is the strength of the input wave required to cause a specified amount of power to be delivered to the load to which it is connected. Sensitivity is an important property of radio receivers. In the case of receivers used for entertainment purposes, this property and its measurement are fully standardized.

12-384. RECEIVER SENSITIVITY. Receiver sensitivity is defined broadly as the input carrier voltage, with a specified standard modulation, that must be supplied by a signal generator in order to develop a standard value of test output with all gain controls set at maximum. Thus, sensitivity in a receiver is a measure of the ability to receive weak signals.

12-385. STANDARDS. Inasmuch as sensitivity measurements imply a standard test output, sensitivity is frequently expressed in terms of either the input signal magnitude or the input signal power. In the case of broadcast receivers operating at low and moderate radio frequencies, sensitivity is commonly expressed in microvolts. On the other hand, at microwave frequencies, where power quantities are more conveniently measured, sensitivity is commonly expressed in microwatts of input power.

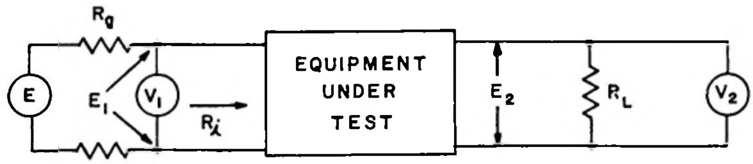
12-386. In amplifier applications where standardization does not yet exist, sensitivity may be expressed in terms of any convenient measure of the input wave, with a standard output power implied and with gain controls set at maximum.

12-387. BASIC MEASUREMENTS.

12-388. The various concepts defined in paragraphs 12-369 through 12-386 may be interpreted in terms of basic measurement circuits which are valid in principle at all frequencies. In these paragraphs, a circuit suitable for the measurement of each gain-related property is illustrated, together with formulas for computing the result from the measured data. In figures 12-137 through 12-143 the formulas are included with the measurement circuit illustrations. Specific comments relating to the figures are given below. It is to be noted that all voltages and currents used in the formulas are root-mean-square values.

12-389. GAIN.

12-390. POWER. In figure 12-137, it has



R_g AND R_L ARE THE RATED SOURCE AND LOAD IMPEDANCES, ASSUMED TO BE RESISTIVE.

R_i IS THE INPUT RESISTANCE OF THE EQUIPMENT UNDER TEST

$$\text{POWER GAIN} = \frac{P_L}{P_{REF}}$$

$$P_L = \frac{(E_2^2)}{R_L}, \quad P_{REF} = \frac{(E_1^2)}{R_i}$$

$$\text{POWER GAIN (db)} = 20 \text{ LOG}_{10} \frac{E_2}{E_1} + 10 \text{ LOG}_{10} \frac{R_i}{R_L}$$

Figure 12-137. Power Gain Measurement

been assumed that the source and load impedances are essentially resistive. This is commonly the case in practice and is highly desirable in making measurements on a comparable basis. It has further been assumed that the input impedance to the equipment under test is also resistive. When the impedances are complex, it becomes necessary in computing the gain, to (a) measure the magnitudes and power factors of the impedances concerned or (b) measure their real and reactive parts. If, for example, the computation is based upon impedances Z_{Load} and Z_{in} which are to be regarded as complex, this gain is expressed as follows:

$$\text{Power Gain (db)} = 20 \log_{10} \frac{E_2}{E_1}$$

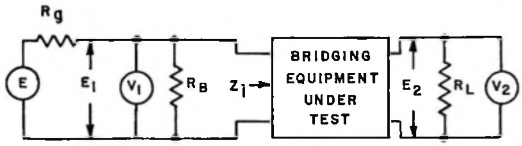
$$+ 10 \log_{10} \frac{Z_{in}}{Z_L}$$

$$+ 10 \log_{10} \frac{(pf)_L}{(pf)_{in}}$$

where E_1 and E_2 are defined in figure 12-137 and the pf terms represent the power factors (cosines) of the designated impedance angles.

12-391. BRIDGING. It is usual for the input impedance (Z_i) of the bridging equipment to be high, relative to the reference load (R_B) across which it is bridged. The presence of Z_i and its properties do not influence the computation of bridging gain as given in figure 12-138. However, when a direct measurement of E_1 is not feasible and computation of the impressed input power must be made indirectly, it is necessary to evaluate carefully the effect of the bridged impedance.

12-392. INSERTION AND TRANSDUCER. What has been said in connection with impedances other than those which can be regarded as essentially resistive applies to the computation of insertion and transducer gain. Compare figure 12-139 with figure 12-140. For the reference condition in the latter measurement, it is only necessary to know that the resistive component of the



$$P_{REF} = \frac{(E_1)^2}{R_B} \qquad P_L = \frac{(E_2)^2}{R_L}$$

$$\text{BRIDGING GAIN (db)} = 10 \text{ LOG}_{10} \frac{P_L}{P_{REF}}$$

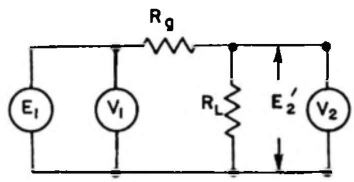
$$= 10 \text{ LOG}_{10} \left(\frac{E_2}{E_1}\right)^2 \frac{R_B}{R_L}$$

Figure 12-138. Bridging Gain Measurement

ideal transducer will cause a maximum-power transfer match under normal circumstances. When the output impedance of the equipment under test and the load impedance are both complex, the measurement requires that they be matched on a conjugate basis. This can be accomplished by means of an appropriate impedance-matching network; see figure 12-141.

12-393. MEASUREMENT MODIFICATIONS.

12-394. The basic measurement circuits which have been described require some modifications in actual practice, in order to achieve a satisfactory degree of accuracy

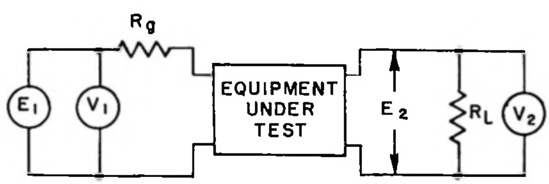


$$P_{ref} = \frac{(E_2')^2}{R_L}$$

R_g AND R_L ARE THE RATED SOURCE AND LOAD RESISTANCES RESPECTIVELY

A

THE REFERENCE CONDITION



$$P_L = \frac{(E_2)^2}{R_L}$$

B

WITH THE EQUIPMENT UNDER TEST INSERTED IN THE CIRCUIT

$$\text{INSERTION GAIN (db)} = 10 \log_{10} \frac{P_L}{P_{ref}}$$

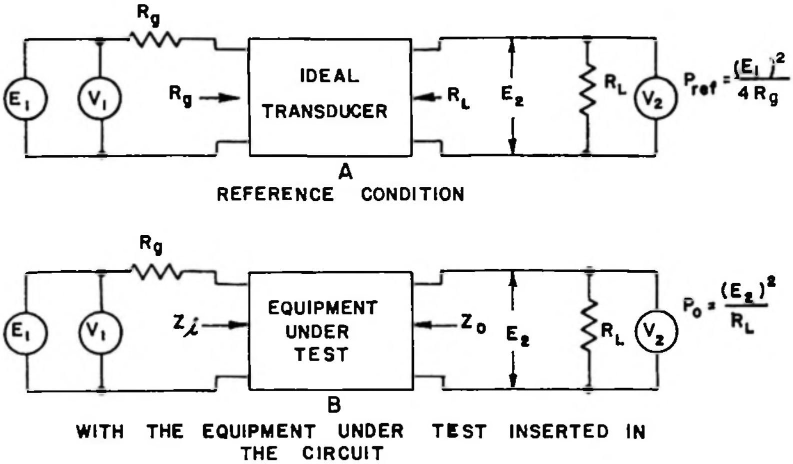
$$= 20 \log_{10} \frac{E_2}{E_2'}$$

IN TERMS OF E_1, ASSUMED TO REMAIN CONSTANT,

$$P_{ref} = \frac{E_1^2}{R_g + R_L} \cdot \frac{R_L}{R_g + R_L} = E_1^2 \frac{R_L}{(R_g + R_L)^2}$$

$$\text{INSERTION GAIN (db)} = 20 \log_{10} \left[\left(\frac{E_2}{E_1}\right) \frac{(R_g + R_L)}{R_L} \right]$$

Figure 12-139. Insertion Gain Measurement

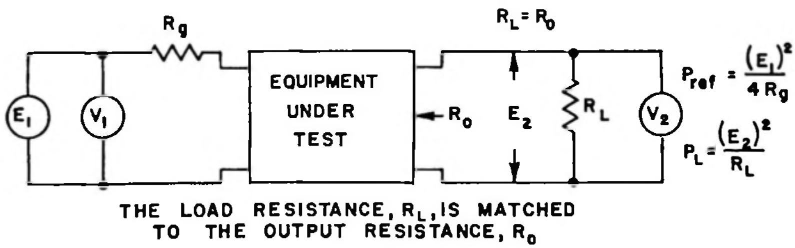


TRANSDUCER GAIN (db) $= 10 \log_{10} P_L/P_{ref}$

$$= 10 \log_{10} \left[\frac{4R_g}{R_L} \left(\frac{E_2}{E_1} \right)^2 \right]$$

$$= 6 + 20 \log_{10} \frac{E_2}{E_1} + 10 \log_{10} \frac{R_g}{R_L}$$

Figure 12-140. Transducer Gain Measurement



AVAILABLE GAIN (db) $= 10 \log_{10} \frac{P}{P_{ref}}$

$$= 10 \log_{10} \left[\frac{4R_g}{R_L} \left(\frac{E_2}{E_1} \right)^2 \right]$$

Figure 12-141. Available Gain Measurement

and to afford certain operational conveniences. Specific modifications depend not only upon the particular frequency range, but more broadly upon such considerations as the gain of the device under test, the normal signal levels of operation, the degree of accuracy required, and the available instrumentation.

12-395. Modifications which are almost always required, regardless of frequency range, include the use of calibrated adjustable attenuators and impedance-matching networks. In the measurement of high-gain equipments, the normal input signal may be of such an exceedingly low order of magnitude as to make direct measurement with the required accuracy extremely difficult using normally available instruments. Suitable levels are achieved by using a calibrated adjustable attenuator in conjunction with the input meter. The attenuator is sometimes an integral part of the signal generator serving as the source of power for the measurements; in other cases, it is a separate component of the measurement circuit.

12-396. It is also convenient, although not essential, to use a second calibrated attenuator in conjunction with the output meter so that the scale deflection of the output meter may be maintained at a constant and favorably placed position. When these techniques are used, it is possible to control the scale deflections of both meters, even to the point of having the two deflections identical. Variations in gain are then determined directly by the input and output attenuators. There are a number of advantages to this method of operation. The meter scale range, the absolute sensitivity, and possible variations in accuracy with deflection are no longer critical factors in the selection of a meter. It is also possible, in certain applications, for you to use uncalibrated indicators; for example, crystal detector arrangements and cathode-ray oscilloscopes. This method also permits the maximum flexibility to be incorporated in a single test equipment, since a wide range of signal levels and gain and loss measurements can easily be accommodated.

12-397. The inclusion of calibrated attenuators in the measuring equipment usually requires the use of impedance-matching networks. In making gain measurements, the equipment or component under test must be terminated in the impedance with which it is designed to be used. Available attenuators frequently do not meet these impedance requirements. It then becomes necessary for you to use some kind of impedance-matching network interposed between the input and output of the equipment under test and the adjacent calibrated attenuator that requires impedance correction. Impedance-matching networks take several forms, the most common being a fixed-loss pad. At the lower frequencies transformers are used occasionally for this purpose, if their frequency characteristics are satisfactory; and at microwave frequencies tuners are commonly used.

12-398. <u>PRACTICAL CIRCUIT AT AUDIO FREQUENCIES.</u>

12-399. GENERAL.

12-400. Audio-frequency measurements, as developed to fit the needs of complex equipments such as those used in radio broadcasting transmission, have, of necessity, attained a high order of refinement; in fact, many of the procedures have become standardized. These paragraphs present the various techniques used in making gain and amplification measurements, including specific circuits and methods of computing the results. These techniques are described in considerable detail, for they apply equally to the measurement of facilities or to equipments of facilities. Furthermore, many of them can be applied in principle regardless of the frequency range. They will therefore serve to suggest sound practices to be followed in newer fields where measurement procedures have not yet become as well known.

12-401. ABSOLUTE AND RELATIVE GAIN MEASUREMENTS.

12-402. Figure 12-142 presents, in block

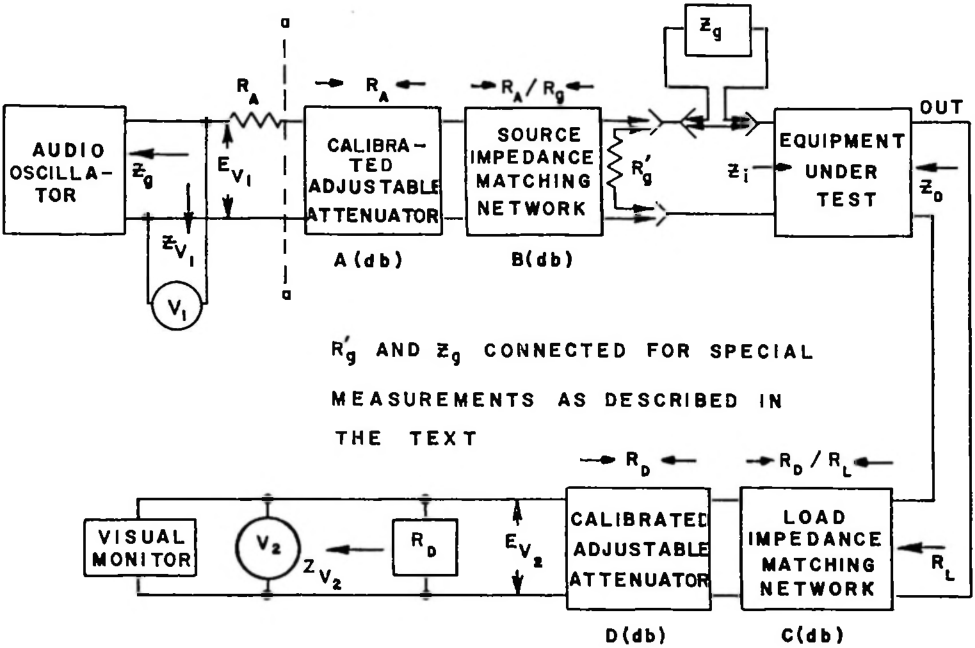


Figure 12-142. Measurement of Absolute and Relative Gain, Equipment Arrangement

diagram form, a versatile measuring circuit suitable for the measurement of gain and amplification in connection with facilities or equipments. The circuit arrangement affords maximum flexibility in the types of measurement that can be performed and in the range of values that can be measured.

12-403. TERMINATION. Provision is made for the equipment under test to be terminated in the source and load impedances with which it is intended to operate. These terminations, which are designated R_g and R_L in the figures, are usually chosen to be pure resistances so that comparable results can be assured. The source and load impedance-matching networks, B and C, provide means, when necessary, for matching the input and output adjustable attenuators, A and D, to the impedances, R_g and R_L. In audio-frequency practice, these networks ordinarily take the form of simple fixed-loss pads. The calibrated adjustable attenuators, together with the fixed matching pads, provide sufficient flexibility to permit the voltmeters, V_1 and V_2, to be operated at any desired scale deflection.

12-404. SOURCE TYPES. The input section to the left of the junction marked a-a in figure 12-142 simulates the type of source shown in figures 12-137 through 12-141. The output of the audio oscillator is adjusted to maintain constant the voltage E_{V_1} as read on meter V_1. Thus, the network to the left of junction a-a is equivalent to an open circuit voltage (E_{V_1}), in series with an impedance (R_A), having the value which the input attenuator (A) is designed to face. The nature of impedance Z_g is then immaterial insofar as the accuracy of measurement is concerned. The output attenuator termination, indicated as R_D in the block diagram, is actually chosen so that, in combination with the meter V_2 and any other shunting equipment, it presents to the attenuator output the correct resistance value. The impedances R_g and Z_g can be connected into the circuit at the points shown, when required. These additional elements are required in connection with specific measurements, and their use is fully described in the following paragraphs. It should be noted that the input circuit facing the input terminals of the equipment under test can be reduced to a simple series combination of a voltage source and an impedance by employing Thevenin's theorem. Similarly, the circuit facing the output terminals of the equipment under test can be reduced to a simple resistance (R_L).

12-405. TECHNIQUES. The oscillator output level and the settings of the calibrated attenuators should be adjusted until the equipment to be tested is supplied with an input signal, the level of which is safely below the point of overdrive and yet is well above the hum and noise level. The overdrive level can be checked by observation of the output signal on a cathode-ray oscilloscope, preferably over the entire frequency range.

12-406. When you use a single voltmeter and alternately connect it between the input and output positions, the calibrated attenuators should be adjusted until the voltmeter deflection is identical in the two positions. It is important to insure, moreover, that the position of the voltmeter does not affect the signal levels in the equipment.

12-407. When you use separate meters, the output of the oscillator and the settings of the attenuators should be adjusted so that the voltage readings are made on the scales where the calibration accuracy and ease of reading are a maximum for the particular instruments involved.

12-408. You can check the stability of the measuring circuit by making a known change in the level of the signal from the oscillator and at the same time introducing a compensating loss in the input attenuator. No change of level should be observed in the output voltmeter reading. This test should be made at several frequencies, including some near the upper part of the frequency range to be covered. Prior to making measurements, you should operate the equipment for a preliminary period to stabilize the internal temperature. Ambient temperature may have an effect upon the gain of the equipment under test and should be controlled if necessary.

12-409. INSERTION GAIN MEASUREMENT. In figure 12-142, you must assume that the values of resistances R_A, R_g, R_L, and R_D are known and that the fixed pad losses are known. The insertion gain may then be computed from these data and the readings of the voltmeters and adjustable attenuators using the following relationships:

$$\text{Insertion Gain (db)} = A_{db} + B_{db} + C_{db}$$

$$+ D_{db} + 10 \log_{10} \left[\frac{4E_{V_2}{}^2}{E_{V_1}{}^2} \cdot \frac{R_A}{R_D} \right]$$

$$+ 10 \log_{10} \left[\frac{4R_g R_L}{(R_g + R_L)^2} \right]$$

12-410. You can derive the relationships of paragraph 12-409 as follows: The reference condition for the measurement is with the equipment under test removed from the circuit and the output of pad B connected to the input of pad C. Actually, of course, it is not necessary for you to make separate measurements of reference and load power, since a single measurement suffices to yield all quantities needed for the gain computation. At each of the junctions, except where

B and C are joined, the impedances are matched on a maximum-power-transfer basis. The available power at the output terminals of pad B is the available power of the source $\left[\left(E_{V_1}\right)^2 / 4R_A\right]$, less the power dissipated in the input attenuator and the fixed pad (B). The power transferred to the input of pad C (or what is equivalent, to the termination R_L), is a function of the mismatch existing between R_g and R_L. The fraction of the available power that is transferred across the junction of pads B and C is called the mismatch factor or coupling factor, and is shown in this case to be:

$$\text{Coupling Factor} = \frac{4R_gR_L}{(R_g + R_L)^2}$$

The above equation is seen to reduce to unity when $R_g = R_L$. The reference power, expressed in logarithmic form, is found to be:

$$10 \log_{10} P_{ref} = 10 \log_{10} \frac{\left(E_{V_1}\right)^2}{4R_A} - 10 \log_{10} \left[\frac{4R_gR_L}{(R_g + R_L)^2}\right] - A_{db} - B_{db}$$

12-411. With the equipment under test replaced in the circuit, and with E_{V_1} and the input attenuator setting unchanged, you can compute the power in R_L as follows:

$$10 \log_{10} P_L = 10 \log_{10} \frac{\left(E_{V_2}\right)^2}{R_D} + C_{db} + D_{db}$$

Subtracting this equation from the second equation of paragraph 12-410 gives the original expression contained in paragraph 12-409. If the attenuators are adjusted so that E_{V_1} and E_{V_2} are identical, the computation of the insertion gain is further simplified.

12-412. TRANSDUCER GAIN MEASUREMENT. Referring to figure 12-142 again, the only distinction between this measurement and that of insertion gain lies in the computation. Since the reference power is the available power of the effective source facing the input terminals of the equipment under test, no mismatch factor is involved, and the expression for the gain reduces to:

$$\text{Transducer Gain (db)} = A_{db} + B_{db} + C_{db} + D_{db} + 10 \log_{10} \left[\frac{\left(E_{V_2}\right)^2}{\left(E_{V_1}\right)^2} \cdot \frac{4R_A}{R_D}\right]$$

12-413. BRIDGING GAIN MEASUREMENT. For this measurement, a resistance, equal in value to the reference load and shown as R_g' in figure 12-142, is shunted across the output of fixed pad B. The pad is designed to match R_g' to the output impedance of attenuator A. Thus, with the equipment under test disconnected, the input circuit is completely matched and terminated in the reference load. The input of the equipment to be tested is then connected in parallel with the reference load. Equipment intended for bridging operation ordinarily has an input impedance that is high relative to that of the circuit on which it is to be bridged. Nevertheless, it may be necessary for you to take into account the power absorbed by the input circuit of the bridging equipment. The reference power is the power in the reference load with the bridging equipment shunting it. Assuming that Z_i is much larger than R_g' the power in R_g' is found to be:

$$10 \log_{10} P_{ref} = 10 \log_{10} \frac{\left(E_{V_1}\right)^2}{4R_A} - A_{db} - B_{db}$$

The power in the load, R_L, is:

$$10 \log_{10} P_L = 10 \log_{10} \frac{\left(E_{V_2}\right)^2}{R_D} + C_{db} + D_{db}$$

The bridging gain is then found by subtraction to be:

$$\text{Bridging Gain (db)} = A_{db} + B_{db} + C_{db} + D_{db} + 10 \log_{10} \left[\frac{\left(E_{V_2}\right)^2}{\left(E_{V_1}\right)^2} \cdot \frac{4R_A}{R_D} \right]$$

12-414. This last expression has the same form as that describing the transducer gain in paragraph 12-412; however, this does not imply that the bridging and transducer gains of a given facility or equipment are identical. The presence of the reference load (R_g') in the equipment makes the bridging gain of that equipment approximately 6 db lower than the transducer gain. If the equipment under test absorbs appreciable power in its input circuit, the reduction of the power in R_g' must be considered.

12-415. THE MEASUREMENT OF VOLTAGE AMPLIFICATION. In the zero-impedance source method (see figure 12-143), the resistance R_g' is placed across the output of matching network B. Impedance relations are chosen so that R_g' is very much smaller than $|Z_i|$. Under these conditions, the voltage across resistance R_g' approximates an open-circuit voltage with respect to the input of the equipment under test. The voltage amplification, for the conditions stated, is given by the expression:

$$\text{Voltage Amplification (db)} = A_{db} + B_{db} + C_{db} + D_{db} + 20 \log_{10} \left(\frac{2E_{V_2}}{E_{V_1}} \right)$$

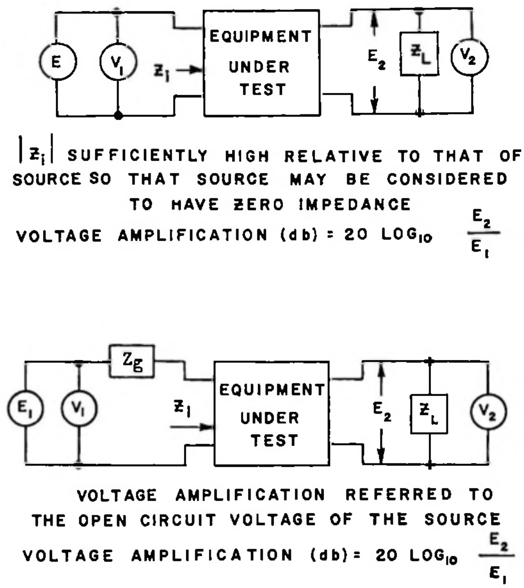


Figure 12-143. Voltage Amplification Measurement

12-416. In the finite impedance source method, resistance R_g' is again required, and, in addition impedance Z_g is inserted in series to simulate the specified impedance of a driving source. Impedance relations must be such that R_g' is very much smaller than $|Z_g|$. Thus the voltage across R_g' approximates a constant voltage source. In typical practice, R_g' should have a value not exceeding 1% of $|Z_g|$. The voltage amplification, for the conditions stated, is given by the expression:

$$\text{Voltage Amplification (db)} = A_{db} + B_{db} + C_{db} + D_{db} + 20 \log_{10}\left(\frac{2E_{V_2}}{E_{V_1}}\right)$$

This second method of expressing voltage amplification is particularly useful when the input impedance of the equipment or component under test cannot be regarded as effectively equivalent to an open circuit, and when the source impedance or the input impedance or both are complex.

12-417. RELATIVE GAIN MEASUREMENTS.

12-418. The basic circuits for the measurement of absolute gain and amplification are suitable in all cases for relative measurements. In particular, the comprehensive measurement circuit of figure 12-142 may be used, with the following additional factors taken into account.

12-419. REFERENCE FREQUENCY. The reference frequency for relative gain measurements is usually selected in the middle (reckoned on a logarithmic frequency scale) of the usable pass band. Where the response of the device is reasonably uniform in the middle of the audio range, the reference is usually taken to be either 400 or 1000 cps.

12-420. MEASUREMENT METHODS. The frequency of the signal source is varied in increments over the useful range of frequencies of the device under test. It is often a good choice to use equal intervals on a logarithmic frequency scale (octaves, for example). The spacing of the frequency intervals is dictated largely by the smoothness, or lack of it, in the frequency response. For this reason, it is advisable to plot the data as they are taken, rather than merely to record data for future presentation. Relative gain may be measured relative to a constant input signal level, or a constant output level, at each frequency.

12-421. In the constant output level method of measurement, the readings of both meters are maintained at some convenient deflection. The output attenuator (D) is left at the setting established in the reference frequency measurement. The audio oscillator is adjusted, when necessary, to maintain constant the reading of meter V_1 at the reference deflection. The input calibrated attenuator is adjusted at each frequency to restore the output meter indication to the reference deflection. Gain is computed in accordance with the appropriate formulas provided previously.

12-422. In the constant input level method of measurement, both meter deflections remain constant at the reference frequency value. The output level of the audio oscillator is adjusted, as necessary, to maintain the reference reading on meter V_1. The input attenuator remains set at the reference value. Changes in gain with frequency are offset by changes in the output attenuator setting to maintain the reference deflection of meter V_2.

12-423. Which of the two methods is preferred depends upon the relative convenience which each offers. In connection with the method of constant output level described above, precautions must be taken to insure that the input signal does not overdrive the device under test at low values of relative gain. The latter method sets the input signal to the device at a level in the reference condition which is known to be safely below the overdrive limit.

12-424. In connection with relative gain measurements, they can be made without the use of attenuators by making the necessary changes in the output level of the audio oscillator and maintaining the reading of either the input or the output voltmeter. The difference in voltmeter readings can then be used for the computation of the gain. It is generally preferable, however, to utilize attenuators so that meter errors are held at an absolute minimum.

12-425. MEASUREMENT CIRCUIT COMPONENT REQUIREMENTS. Although the requirements and characteristics of the measuring equipment components discussed in this section apply ostensibly at audio frequencies, many of the requirements are general in nature and apply regardless of the frequency range.

12-426. AUDIO-FREQUENCY SOURCES. The source of power for absolute and relative gain measurements is ordinarily an audio-frequency oscillator. A list of the characteristics that are desirable, either in the interests of accuracy or as a practical matter of operating convenience, is given below:

a. Adjustable frequency - preferably continuously - at least throughout the useful range of the equipment under test.

b. Output level continuously adjustable at least over a moderate range of values.

c. Absolute frequency stability on the order of 2% maximum. (This may not be sufficiently small when making measurements of frequency-selective devices.)

d. Freedom from output voltage drift with time and with reasonable line voltage variations.

e. Spurious components, such as harmonic distortion and hum, should ordinarily not exceed 10% rms of the output voltage. However, this tolerance may be much too broad when making measurements of devices having sharp cutoff characteristics.

f. Output voltage stability with changes in frequency.

12-427. Some oscillators are critical as to loading, in which case it may become necessary to isolate the load on the oscillator. Several different useful arrangements are illustrated in figure 12-144. Oscillators critical as to load impedances are prone to exhibit an increase in harmonic distortion in their output with departure from isolation loading. In any of the arrangements of figure 12-144, if the voltage E_1 is maintained constant, the source simulates an open-circuit voltage E_1 in series with an impedance R_g.

12-428. METERS. The input voltmeter may be of the vacuum-tube, rectifier or thermocouple type, or it may be a standard volume indicator. The meter may provide rms, average, or peak reading. The stability should be such that its readings are not influenced by extraneous factors, eg, tube power supply.

12-429. If a rectifier meter is used, certain precautions must be taken. The loss introduced in the transmission circuit by meters of this type must be taken into account in computing the gain, especially if the same physical instrument is used for the input and output measurements. This effect can be compensated by substituting for the meter a resistance equal to its input impedance when the meter is disconnected. Standard volume indicators which are a form of rectifier instrument should be arranged to have an impedance magnitude greater than 10 times the impedance across which they are bridged. The transmission loss caused by the bridging of a meter (or other passive impedance that is essentially resistive) across a transmission equipment can be computed from the data given in figure 12-145.

12-430. A thermocouple meter may be used provided that its heater resistance is properly accounted for. These meters are well suited to audio-frequency measurements,

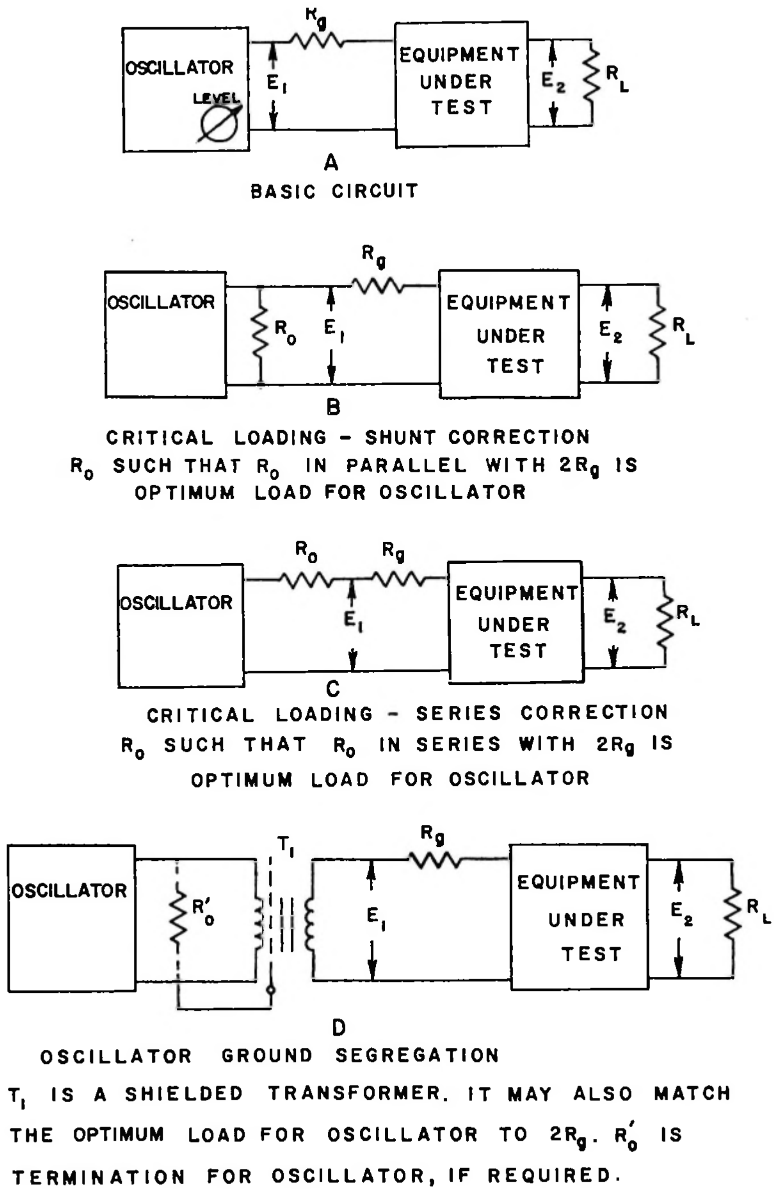
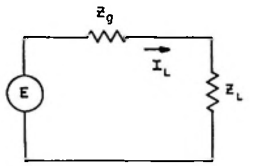


Figure 12-144. Optimizing Oscillator Load

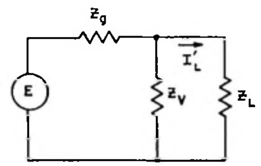
their principal limitation being their liability to damage on overload.

12-431. The output meter may be similar to the input meter, or may actually be the same meter switched alternately between the two positions. If a single meter is used, it is desirable to arrange the signal levels



$$I_L = \frac{E}{Z_L + Z_g}$$

A
REFERENCE CONDITION



$$I_L' = \frac{E Z_V}{Z_g(Z_L + Z_V) + Z_L Z_V}$$

B
WITH IMPEDANCE Z_V BRIDGED

$$\text{BRIDGING LOSS} = 20\log \frac{I_L}{I_L'} = \left| \frac{Z_g(Z_L+Z_V)+Z_L Z_V}{Z_V(Z_L+Z_g)} \right|$$

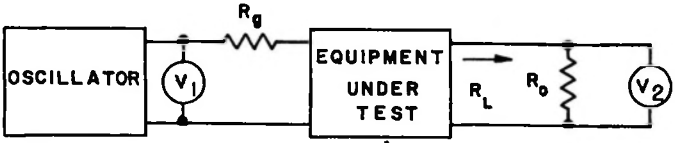
LOSS CAUSED BY A VOLUME INDICATOR:
IF $Z_V = 7500$ (THE IMPEDANCE OF A STANDARD VI) AND
IF $Z_g = R = 600$ AND ALL ARE ESSENTIALLY RESISTIVE
$\text{VI LOSS} = 20 \log_{10} 1.04 = 0.34$ db

Figure 12-145. Loss Caused by a Bridging Meter or Impedance

so that both readings are identical. This not only makes for convenience but yields the advantages that accrue when the meter is reduced to a comparator. Care must be taken in switching a meter between two points of a high-gain equipment to avoid introducing parasitic couplings into the equipment. The meters should desirably not exhibit any appreciable frequency error. If this is not the case, a calibration curve must be resorted to. In addition, the scale spread and the pointer structure must be such that deviations on the instrument with changes in gain are discernible to an accuracy commensurate with the least order of accuracy for the measurement. Figure 12-146 shows some possible variations in meter circuit arrangements that may be employed.

12-432. When separate meters are used, the absolute values of the voltages enter into the computations of the gain, and it is necessary that the calibrations be sufficiently accurate over the frequency range to insure the required gain measurement accuracy. Frequency characteristics of the two meters should preferably be identical so that errors will not be introduced when making frequency response measurements.

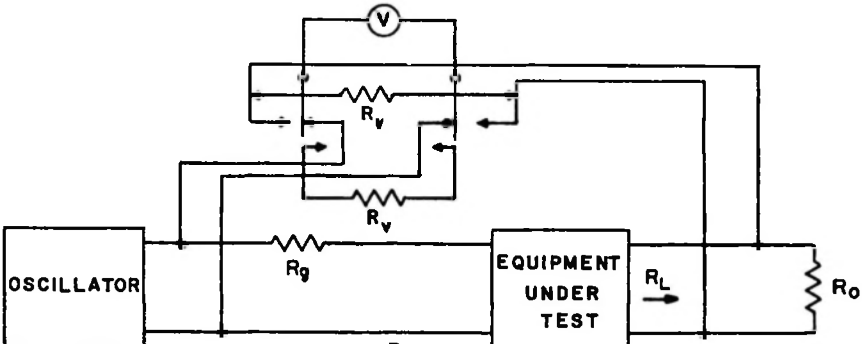
12-433. ADJUSTABLE ATTENUATORS. The adjustable attenuators included in figure 12-142 are not essential to the measurement, except insofar as they permit operational convenience and provide the means for avoiding certain meter errors. The accuracy of calibration and the increments of attenuation available, together with the accuracy with which the voltmeters can be read, govern the accuracy with which the absolute gain can be determined. Relative gain (frequency response) measurements require absolute accuracy so that relative gains are indicated within the desired limits of precision. A total measuring error of $\pm$ 0.2 db from all causes is the order of precision required for relative gain measurements on equipment for broadcast transmission. This limit includes frequency response errors in the attenuators, which, if excessive, can be accounted for by calibration. Not only is proper attenuator design a necessity, but careful maintenance is also required to avoid errors resulting from switch contacts which are in bad condition. The adjustable attenuators should be mounted in shielded enclosures. This precaution helps to avoid errors caused by parasitic capacitance to external surroundings. These stray couplings become increasingly troublesome with increasing frequency. The total range of attenuation should at least equal the over-all gain to be measured. However, the matching pads or other fixed-loss pads can be used to supplement the limited range of available attenuators.



A

THE BASIC CIRCUIT

RESISTANCE OF R_o AND V_2 IN PARALLEL MUST EQUAL R_L

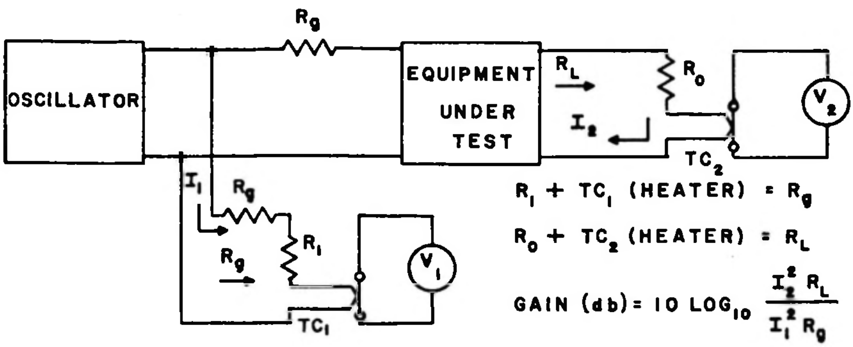


B

THE USE OF A SINGLE METER

RESISTANCE OF R_o AND R_v IN PARALLEL MUST EQUAL R_L

R_v = METER RESISTANCE



R_1 + TC_1 (HEATER) = R_g

R_o + TC_2 (HEATER) = R_L

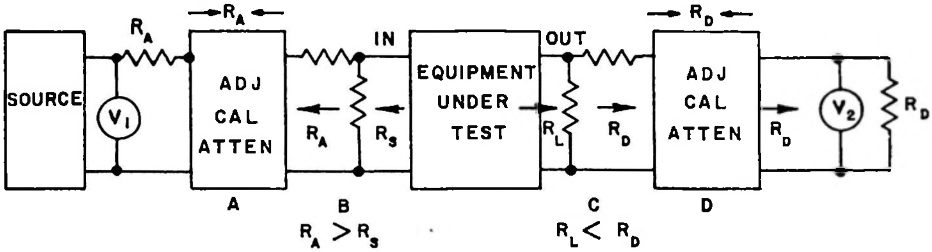
$$\text{GAIN (db)} = 10 \log_{10} \frac{I_2^2 R_L}{I_1^2 R_g}$$

$$= 20 \log_{10} \frac{I_2}{I_1} \sqrt{\frac{R_L}{R_g}}$$

C

TYPICAL USE OF THEROCOUPLE METERS

Figure 12-146. Alternate Meter Connections



B AND C ARE MINIMUM LOSS "L" PADS, MATCHING R_A TO R_S AND R_L TO R_D RESPECTIVELY.

IF Z IS THE LARGER AND z IS THE SMALLER OF TWO IMPEDANCES BEING MATCHED BY EACH MATCHING PAD, THEN

$$\text{LOSS (db)} = 20 \log_{10} \left(\sqrt{\frac{Z}{z}} + \sqrt{\frac{Z-z}{z}} \right)$$

NOTE THAT THE SERIES ARM ALWAYS FACES THE HIGHER IMPEDANCE.

Figure 12-147. Matching Pads

12-434. IMPEDANCE MATCHING. As explained earlier, if the available attenuators do not have impedance values that are correct terminations for the equipment under test, it is necessary to use impedance-matching circuits. These can ordinarily take one of several forms of lumped-resistance attenuators which introduce a fixed loss. These are called fixed-loss pads. The various useful attenuator configurations and the determination of the component values are to be found in other sections of this publication. The most common type for impedance matching is an "L" configuration, known as a minimum-loss taper pad. Figure 12-147 illustrates the use of minimum-loss pads in a typical measurement circuit.

12-435. Pads and terminations should be constructed using noninductive precision resistors (1% or better accuracy as dictated by the accuracy required in the measurements). The resistors should be temperature and frequency stable and should be power rated in accordance with the greatest expected dissipation requirements.

12-436. PRACTICAL CIRCUIT MEASUREMENT AT RADIO AND MICROWAVE FREQUENCIES.

12-437. GENERAL.

12-438. Measurement of gain in the radio-frequency ranges, lying above the audio-frequency range and extending into the microwave range, can be performed using the basic circuits of paragraphs 12-390 through 12-416 with but minor modifications. As the upper end of the frequency spectrum is approached, it becomes more feasible to measure power directly than to measure voltage across a known impedance. Power can be measured using a variety of devices including bolometer-wattmeters, directional couplers, and standing-wave detectors. Crystal rectifiers, in conjunction with microammeters, are used sometimes as square-law devices, but more frequently as constant level indicators. They are also used to respond to the envelope of modulation when the signal source is amplitude-modulated. This latter arrangement has the advantage that

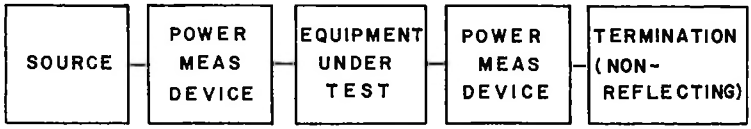


Figure 12-148. Direct Measurement of Power Gain, Block Diagram

the recovered audio-frequency envelope voltage is readily amplified and displayed on an oscilloscope. Signal generators are available to you in a variety sufficient to cover substantially any desired portion of the frequency spectrum. The majority of these are equipped with integrally mounted attenuators, are carefully shielded, and can be operated on a continuous-wave basis or can be modulated. It is fairly common practice at the higher frequencies to operate transmission equipments on a matched basis; at least gain measurements are usually made on that basis since tuners of various kinds make this relatively easy to achieve. When completely matched circuits are used, the distinctions among gain, insertion gain, and available gain disappear.

12-439. It would be virtually impossible to catalog all of the possible variations in the basic circuits that can be used throughout the radio-frequency spectrum. In this section, therefore, only a few typical modifications of the basic circuits are discussed, and special techniques not previously considered are dealt with.

12-440. POWER GAIN MEASUREMENT.

12-441. The block diagram of figure 12-148 presents the essential features necessary for the direct measurement of power gain at frequencies where power-measuring equipment is feasibly used. The power-measuring devices are arranged to measure power in the forward-going waves only. Gain is then computed simply from the ratio of the measured values of output to input power. For this measurement to correspond with the defining condition of paragraphs 12-370 through 12-386, it is necessary that the termination be non-reflecting and that the input of the equipment under test be matched to the input line so as to absorb completely the impressed power. Measurements made under mismatch conditions must be properly interpreted but are otherwise valid.

12-442. The circuit can be modified in various ways so as to incorporate in it some of the lower-frequency techniques. It is assumed that the signal source contains a calibrated attenuator; a second attenuator could be used following the equipment under test so as to operate the equipment and the measuring devices at any desired level. This particular arrangement lends itself to the use of directional couplers, but other power-measuring instruments could be substituted.

12-443. AVAILABLE GAIN MEASUREMENT. In figure 12-149, it is not implied that the input of the equipment under test is matched to the source. It is assumed, however, that the available power of the source is known, because some signal generators are calibrated directly in available power, or the available power can be evaluated from a knowledge of the resistive component of the source impedance and the open-circuit voltage. When you make the measurement, adjust the output of the signal generator until the desired power level is attained in the load. The ratio of this power to the source power is the available gain.

12-444. TYPICAL UHF CIRCUIT GAIN MEASUREMENT. Figure 12-150 illustrates several adaptations of basic techniques of gain measurement for use in the uhf range. A feature of the circuit is the use of a standing-wave detector for power measurements. In this circuit, no attempt is made to match the input of the equipment under test, because the slotted section permits the power absorbed by the input to be calculated from the standing wave ratio. The signal generator is modulated at 1000 cps, and the slotted-line crystal responds to the envelope of the audio-frequency modulation. The amplitude of the envelope at any point is indicated by the setting of a calibrated attenuator, in conjunction with an oscilloscope used as a level comparator. The attenuator is calibrated in db's so that the standing-wave ratio in decibels can be measured directly. The swr in db is proportional to A_{max} minus A_{min}, which are the respective attenuator settings for constant indicator reading as the probe is traveled through the slotted section. The crystal is operated in the square-law region; therefore, the swr in decibels is given by:

$$S_{db} = \frac{A_{max} - A_{min}}{2}$$

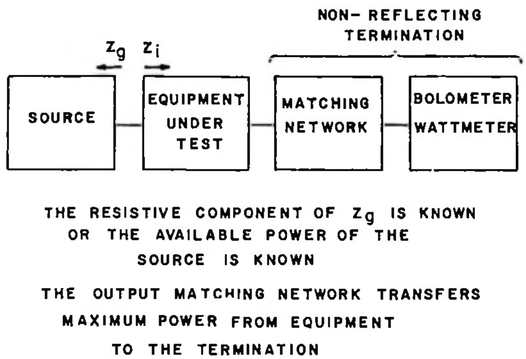


Figure 12-149. Measurement of Available Gain, Block Diagram

12-445. The output of the equipment under test is matched to its termination by means of an impedance-matching network comprised of a line stretcher and a stub tuner. A crystal rectifier and microammeter serve as an rf load-voltage indicator. The technique of measurement is to adjust the input power level until a convenient deflection is obtained on the load microammeter. The

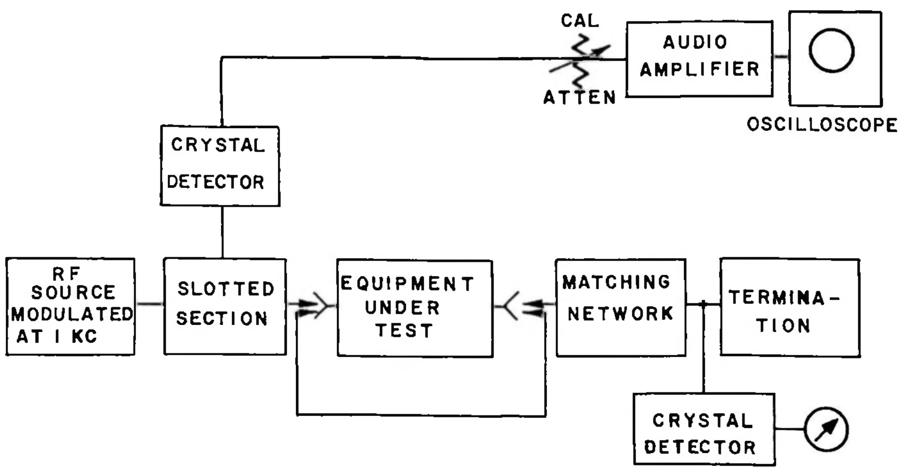


Figure 12-150. UHF Gain Measurement Using Slotted Line, Block Diagram

input power is then measured by means of the standing wave detector. After removing the equipment under test from the circuit, the load and its associated crystal detector are connected directly to the slotted line. The signal generator is then readjusted to give the same indication on the crystal detector as before. It is then furnishing to the load the same power which had been delivered by the equipment under test. This power is then measured. The ratio of the power delivered to the load to the power absorbed by the input of the equipment under test is the gain. In terms of the attenuator settings, it is given by:

$$\text{Power Gain (db)} = \frac{1}{4}\left[A_{max} + A_{min}\right] \text{ with load alone} - \frac{1}{4}\left[A_{max} + A_{min}\right] \text{ with equipment inserted}$$

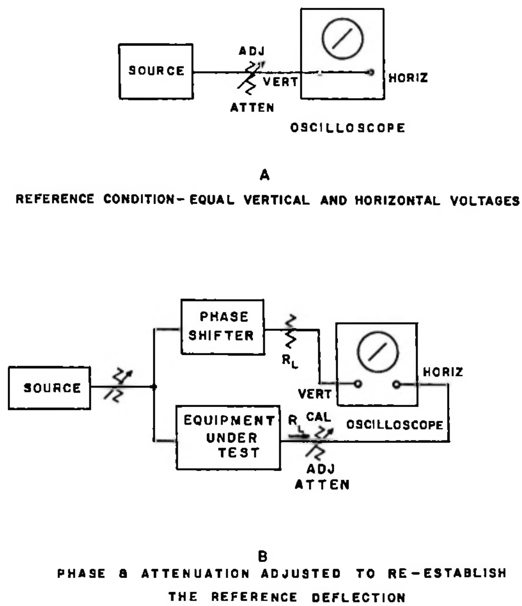


Figure 12-151. Basic Gain Measurement, Using Null Method

12-446. SPECIAL METHODS FOR GAIN MEASUREMENT.

12-447. GENERAL.

12-448. The design, line-up, and maintenance requirements of complex transmission equipments have led to the development of an important group of modern transmission measurements. Inasmuch as a continuous and preferably instantaneous record of various transmission properties is needed, fast-acting automatic devices are used that are capable of scanning the equipment with respect to frequency, time, or amplitude. A common feature of these equipments is the use of a comparator technique for the correlation of output-to-input or output-to-reference on an almost instantaneous basis. Closely related to the comparator technique is the conventional null technique of indicating a bridge balance, which may also be used to advantage in certain transmission measurements. As an introduction to a discussion of scanning techniques, some examples are given of static comparator and null techniques.

12-449. NULL METHOD OF GAIN MEASUREMENT.

12-450. AUDIO AND LOW RADIO FREQUENCIES. You can determine the gain or amplification by using the basic arrangement illustrated in figure 12-151. Establish a reference condition, as in part A of the figure, by applying equal voltages to the vertical and horizontal channels of an oscilloscope, and marking the deflection slope. Connect the equipment under test, as in part B, and adjust the phase-shifter and calibrated attenuator until the reference slope is re-established. The gain can then be computed from the reading of the calibrated attenuator, giving due regard to the matter of terminations. The phase shifter is required only when the phase shift of the device under test is sufficient to cause the os-

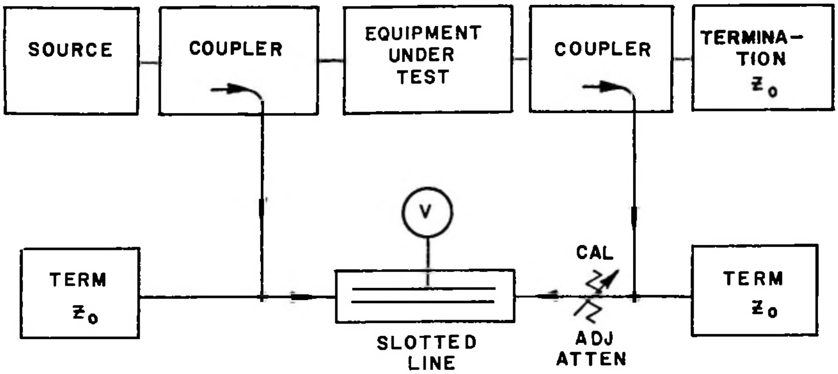


Figure 12-152. Basic Principle of a Waveguide Bridge, Block Diagram

cilloscope figure to depart appreciably from a straight line. The use of a calibrated phase shifter would, of course, provide means whereby the phase response could be measured.

12-451. This basic principle has wide applicability. Passive networks and components can be measured for loss and attenuation by placing the calibrated attenuator in the arm of the bridge containing the phase shifter. In place of the oscilloscope, other arrangements can be used to indicate a null when the bridge is balanced. Null techniques have the advantage of high accuracies without the need for elaborate instrumentation. In connection with measurements of frequency-selective networks, the null balance removes the need for special precautions concerning oscillator harmonic content. Accuracies of better than 0.5 db have been reported at the peak of attenuation in filter measurements.

12-452. MICROWAVE FREQUENCIES. The null technique previously described can be adapted for use at microwave frequencies. Figure 12-152 shows the basic components of a waveguide bridge used for gain measurements. Two independent waves are impressed upon the opposite ends of a standing wave slotted section. One of these, proportional to the power impressed upon the input to the device under test, travels from left to right in the figure. The other, proportional to the output power, travels in the opposite direction. When the amplitudes of the two waves are equalized by adjustment of the calibrated attenuator, a standing wave pattern is set up in the slotted line. If the sliding probe is now placed at one of the nodes, an exact null can be achieved by critical adjustment of the attenuator. The attenuator reading is then an indication of the ratio of the output power to the input power, if it can be assumed that the directional couplers are identical and that the system is matched so that the impressed and observed powers are equivalent.

12-453. TYPICAL EQUIPMENT CONFIGURATIONS.

12-454. GAIN FREQUENCY SCANNER TECHNIQUE. Although equipments may be scanned relative to frequency, time, or amplitude, and measurements made of a wide variety of properties, this discussion is largely related to what are sometimes called gain-frequency scanners. These display gain or amplification as a function of frequency. The scan may take place slowly enough to permit a record to be made by a

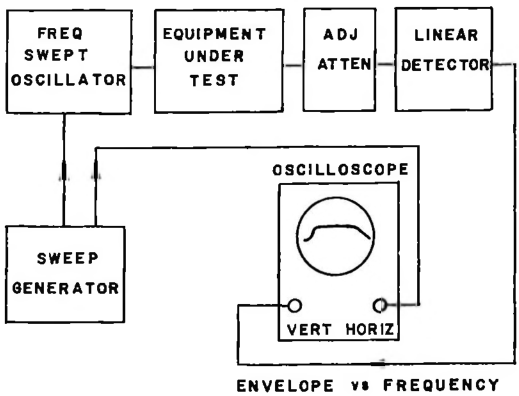


Figure 12-153. Basic Requirements of a Gain-Frequency Scanner, Block Diagram

mechanical pen on a paper strip, or faster scans may be used and the response displayed in the form of a cathode-ray trace. For permanent records in the latter case, photographs are readily made. Although virtually a necessity in connection with equipment measurements, the techniques are applicable to the measurement of a wide variety of components. Nor are these methods limited to any particular portion of the frequency spectrum. Certain precautions with respect to scanning rates must be taken, particularly over broad frequency bands. As a rough guide, to avoid distortion of the response because of too rapid a scanning rate, the rate of sweeping should not exceed the square of the bandwidth.

12-455. The block diagram of figure 12-153 presents the essential features of a gain-frequency scanner for the production of a qualitative record of equipment performance. Various modifications may be incorporated in the basic arrangement to permit quantitative measurements to be made. The power source is an oscillator capable of having its frequency swept linearly over a specified range with the output level remaining reasonably constant. The resulting frequency-modulated signal is passed through the equipment under test and, after suitable attenuation, is rectified in a linear detector to recover a voltage proportional to the amplitude-frequency variations of the output of the equipment. This voltage, after suitable amplification, is displayed as a cathode-ray trace. The oscilloscope deflection and the frequency variation are arranged to be approximately proportional.

12-456. A number of problems arise when this basic circuit is adapted for making accurate quantitative measurements. The equipment under test requires an input signal which is constant in amplitude but variable in frequency over a specified band. You must take precautions, therefore, to insure constant output from the oscillator as it is swept through the frequency range. This problem is usually solved by using what amounts to a fast-acting automatic gain control arrangement, the details of which vary considerably with the frequency range. Another problem arises in connection with sweeping the oscillator. In order to avoid a distorted display, a proportionality is required between the oscilloscope deflection and the instantaneous frequency. If the oscillator can be swept linearly and if the same linear sweep is applied to the oscilloscope, you should encounter no difficulty. However, it is frequently difficult to produce a linear frequency sweep, particularly when mechanical tuning arrangements are used. When this is the case, you can derive a suitable deflection voltage from the frequency variation itself, by utilizing a linear frequency-modulation detector (an fm receiver, for example). Thus, the frequency may vary at an arbitrary rate with the attendant advantage that devices such as rotating air-variable capacitors may be used. When there are no hysteresis effects, it is unnecessary to suppress the return trace, as would be the case were a linear sawtooth scan used. The fm detector may sample the output of the oscillator directly, or alternatively, the output of the equipment under test. The latter arrangement may be preferable when an appreciable time delay in transmission is involved.

12-457. In order to obtain quantitative measurements, you will have to establish references for both frequency and amplitude. The latter can be taken care of by a comparator arrangement whereby the response of the equipment under test and a reference signal are switched alternately into the oscilloscope, thus creating two independent traces. The reference is established by adjusting the attenuator, with the equipment under test removed from the circuit, until the two traces coincide. With the equipment returned to the circuit, you can adjust the attenuator until the traces coincide at a reference frequency. Thus the absolute gain of the equipment is established. The reference may be a zero line, or it may be the input signal to the equipment. The switching should take place at a rate which preferably is faster than the persistence of vision, but which allows a full scan in each position. The switching rate should be synchronized to the scan and preferably should be a multiple of the scanning rate. For example, if the scanning rate is 60 cps (a common choice because of its convenience), the switch might be a 30-cps synchronous switch, possibly in the form of a polarized relay.

12-458. You can make absolute frequency measurements by providing a frequency marker from some kind of calibrated marker generator. This can sometimes be accomplished very conveniently by loosely coupling, to the oscillator output, a sharply tuned absorption-type wavementer. This will cause the trace to contain a narrow absorption notch corresponding to the frequency setting of the wavemeter. You can use two wavemeters when it is desired to mark the limits of the band under scrutiny. A gain-frequency scanning equipment incorporating these refinements is shown in figure 12-154.

12-459. It is of interest to note that the gain scanner is readily converted to a gain-slope scanner, displaying the frequency derivative of gain, by differentiating the output of the equipment under test before impressing it on the oscilloscope. This is based on the assumption that if you scan the frequency linearly, the frequency and the time derivatives will be linearly proportional. The frequency derivative of gain is of interest in equipments exhibiting very smooth gain-frequency properties.

12-460. The accuracies that are realizable from scanning measurements are heavily dependent upon the care with which you set up the scanning arrangement. Generally speaking, accuracies comparable to those obtainable from point-by-point methods are possible, and, in the case of very complicated equipments, the inherent accuracy of the scanning technique exceeds that of the point-by-point methods.

12-461. FREQUENCY CONVERTERS IN HIGH-FREQUENCY GAIN MEASUREMENTS. The application of heterodyne frequency conversion to transmission measuring offers the important advantage of operating the standards of the measurement equipment at a fixed frequency. Under ordinary conditions of measurement over broad frequency ranges, you may find it increasingly difficult to realize standards that exhibit constant properties over a broad range as the frequency requirements go higher and higher. Elaborate calibrations become a necessity, but these are limited by the difficulties connected with measurement of the calibration errors.

12-462. Frequency conversion has been successfully employed to overcome the problem of adequate standards. The accuracy of the measurements depends largely upon the linearity of the conversion process. With careful attention to design, you can achieve linearities of 0.01 db and 0.1 degree in the

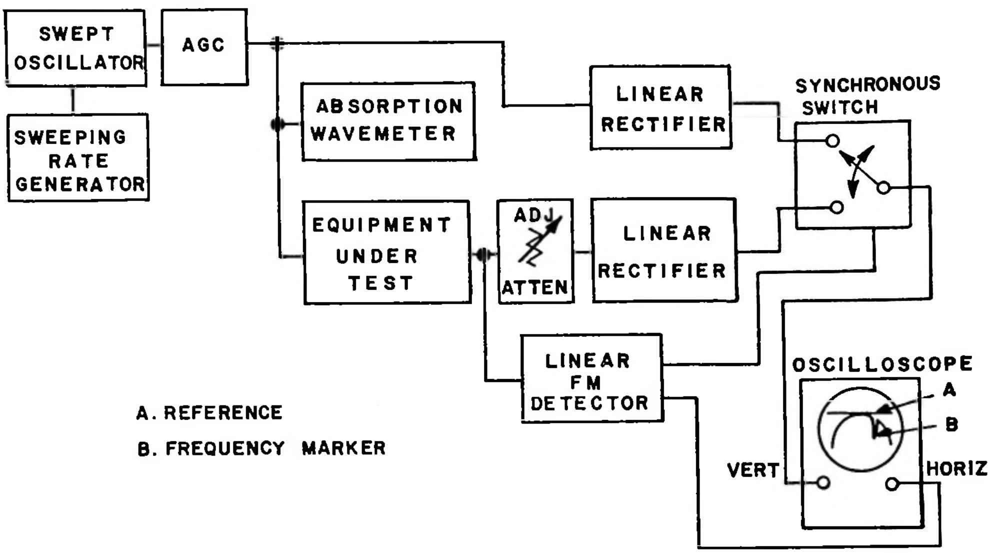


Figure 12-154. Gain-Frequency Scanner for Quantitative Measurement, Block Diagram

frequency range from 0.05 to 80 mc and with a dynamic range of as much as 60 db The conversion principle lends itself to either point-by-point or scanning measurements of gain, phase, delay, and impedance.

SECTION IX

ATTENUATION AND ITS MEASUREMENT

12-463. GENERAL.

12-464. DEFINITION OF TERMS USED IN ATTENUATION MEASUREMENT.

12-465. ATTENUATION. The term attenuation refers to the weakening of a stimulus as it travels through a medium. Generally, it is defined as being proportional to the ratio of the energy measured at a specified location to either a reference value or a reference location. When taken to be the ratio of electrical powers, attenuation becomes a measure of power loss or power reflection of the transmission equipment between two points of observation. These terms will be elaborated upon later. Under proper conditions, attenuation may also be expressed as the ratio of voltages or the ratio of currents.

12-466. For convenience, an electrical equipment may be represented by a variety of four-terminal structures in cascade. Each structure performs a desired function with a certain power efficiency. For example, consider the representative block diagram shown in figure 12-155. Each block represents a four-terminal structure, symbolized by the letter A. The signal source (S) feeds power to the load (L). The power level at each pair of terminals is designated by the symbol P with appropriate subscripts. Referring to figure 12-155, the power efficiency, η, of each individual block can be formulated, therefore, as:

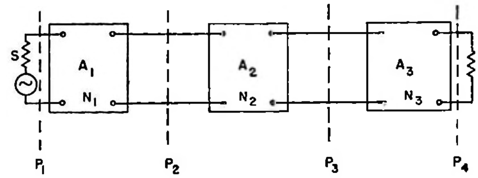


Figure 12-155. Four-Terminal Structures in Cascade, Block Diagram

$$\eta_1 = \frac{P_2}{P_1}, \text{ for block } A_1$$

$$\eta_2 = \frac{P_3}{P_2}, \text{ for block } A_2$$

$$\eta_3 = \frac{P_4}{P_3}, \text{ for block } A_3$$

In addition to the power efficiencies of the four-terminal structures, there is also an over-all power efficiency of the equipment, defined as:

$$\eta_T = \frac{P_4}{P_1}$$

By comparing these equations, you can see that the over-all power efficiency of the equipment is equivalent to the product of the individual block efficiencies. Thus:

$$\eta_T = \eta_1 \times \eta_2 \times \eta_3$$

or

$$\eta_T = \frac{P_2}{P_1} \times \frac{P_3}{P_2} \times \frac{P_4}{P_3} = \frac{P_4}{P_1}$$

12-467. Since the multiplication of numbers is often not very convenient, an additive process can be devised by the use of logarithms. By this expediency, the logarithm of the over-all power efficiency of the equipment becomes simply the addition of the logarithms of the various block efficiencies. Let the logarithms be represented as follows:

$$A_1 = \log \eta_1$$

$$A_2 = \log \eta_2$$

$$A_3 = \log \eta_3$$

The logarithm of the over-all power efficiency, A_T, equates to:

$$A_T = \log \eta_T = A_1 + A_2 + A_3$$

Therefore, it appears that the logarithmic method of measuring power ratios leads to simplification in calculations.

12-468. Another fundamental reason for using logarithms is based upon the Weber-Fechner law of psychology, which states that "the minimum change in stimulus necessary to produce a perceptible change in response is proportional to the stimulus already existing." When interpreted mathematically, this law affirms that sensation is proportional to the logarithm of the stimulus. For example, in order to increase the hearing response of the human ear by 200 percent, it would be necessary to increase a sound intensity by 100 times. This effect is indicated by the logarithmic unit. Since attenuation is related to power ratios, its unit is selected to be logarithmic for the same reasons as described above.

12-469. NEPER. The neper unit is derived from the Naperian system of logarithms invented by John Napier. This system uses the base e, that is, e = 2.7183, and the attenuation in terms of nepers is defined to be:

$$A_n = \frac{1}{2} \log_e \frac{P_o}{P_i} = \frac{1}{2} \ln \frac{P_o}{P_i}$$

where P_o and P_i are the output and input powers, respectively, and ln is equivalent to $\log_e$. When the output and input impedances are identical, this expression reduces to:

$$A_n = \ln \frac{V_o}{V_i} \text{ nepers}$$

and

$$A_n = \ln \frac{I_o}{I_i} \text{ nepers}$$

where V_o and V_i represent the amplitude of the output and input voltages, respectively; likewise, I_o and I_i represent the amplitude of the output and input currents, respectively.

12-470. BEL AND DECIBEL. The bel unit is based upon the Briggsian, or decimal, system of logarithms, invented by Henry Briggs. This logarithm refers to a base of 10, and attenuation in terms of bels is defined to be:

$$A_d = \log \frac{P_o}{P_i} = 2 \log \frac{V_o}{V_i}$$

$$= 2 \log \frac{I_o}{I_i} \text{ bels}$$

where these symbols have the same meaning as described above. The bel unit commemorates Alexander Graham Bell, the inventor of the telephone. Since it is rather large for practical purposes, the decibel unit is used instead of the bel, the decibel being the tenth part of a bel. Thus, the last equation may be re-expressed as:

$$A_d = 10 \log \frac{P_o}{P_i} = 20 \log \frac{V_o}{V_i}$$

$$= 20 \log \frac{I_o}{I_i} \text{ decibels}$$

The decibel unit is abbreviated db.

12-471. Although the decibel represents a unit of power ratio, it may also be used as a relative power unit by comparing measured powers to a standard power level, or a so-called zero level. Thus, a power level of A decibels can be interpreted to signify A db above a zero level. Commonly used zero levels of power are 1 mw, 6 mw, and 12.5 mw. The most popular of the three is 1 mw. Decibels above or below 1 milliwatt are usually abbreviated $\pm$ dbm.

12-472. MILES OF STANDARD CABLE. In referring to the earlier literature in telecommunications, you often come across a transmission loss unit known as miles of standard cable, abbreviated msc. A mile of standard cable is defined as the power loss between the ends of a mile of cable at 796 cps, whose constants per loop mile are as follows:

Weight = 20 lb

R = Resistance = 88 ohms

L = Inductance = 1.0 mh

C = Capacitance = 1.054 μf

G = Leakance = 1.0 μmho

A figure of 10 msc means 10 times the loss of a mile of standard cable.

12-473. CONVERSION CONSTANTS.

12-474. NEPERS VS DECIBELS. To convert nepers into decibels, or vice versa, consider the equations of paragraph 12-469. The first step is to relate the logarithms involving the bases e and 10, as follows:

$$A_n = (0.11513)\, A_d \text{ nepers}$$

and

$$A_d = (8.686)\, A_n \text{ decibels}$$

12-475. THE DIFFERENT POWER ZERO LEVELS. Assume that P represents a measured power level; the resulting attenuation in decibels above two different zero levels given as P_1 and P_2 are shown, respectively by the following equations:

$$A_1 = 10 \log \frac{P}{P_1} \text{ db}$$

and

$$A_2 = 10 \log \frac{P}{P_2} \text{ db}$$

where

$$A_2 = 10 \log \frac{P}{P_1} + 10 \log \frac{P_1}{P_2}$$

and, therefore,

$$A_2 = A_1 + K$$

$$K = 10 \log \frac{P_1}{P_2}$$

It can be seen, therefore, that the equations above transform any power in db referred to a zero level of P_1 to a zero level of P_2. The transformation involves the addition of

K to A_1 in order to realize A_2, as described by the preceding equations. However, K may be either a positive or negative quantity, depending upon the values of P_1 and P_2.

12-476. MSC UNIT. This unit permits an easy method of determining the attenuation constant of a mile of standard cable, as previously explained. For example:

$$1 \text{ msc} = 0.10616 \text{ neper}$$

Conversions between msc units and nepers or decibels are tabulated in table 12-4.

Table 12-4. Conversions Relating MSC, Neper, and Decibel Units

MULTIPLY	BY	TO OBTAIN
Nepers	9.420	msc
Decibels	1.084	msc
msc	0.10616	Nepers
msc	0.9221	Decibels

12-477. SIGNIFICANCE OF SIGN. Since the logarithm of the ratio of two powers may be either positive or negative, it is necessary to interpret the sign properly when discussing attenuation in an equipment composed of cascaded four-terminal structures. When the attenuation involving the ratio of power output to power input of a four-terminal structure is positive, the structure is said to possess an amplification gain. When the sign is negative, the insertion of the structure results in a power attenuation which may be due to either power dissipation, power reflection, or a combination of both.

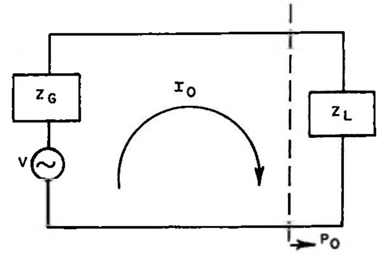


Figure 12-156. Simple Series Circuit of Two Mismatched Impedances

12-478. ATTENUATION AT ALL FREQUENCIES.

12-479. NORMAL FOUR-TERMINAL CIRCUITS.

12-480. When a passive four-terminal circuit is introduced between a signal source and a terminating load, and the power (P_o) delivered to the load is reduced to P_2, the resulting attenuation is defined to be:

$$A = 10 \log \frac{P_o}{P_2} \text{ db}$$

Since the attenuation is caused by the insertion of the four-terminal circuit, it is often referred to as insertion loss. Such attenuations are not only dependent upon the parameters of the inserted circuit, but also upon the load impedance and the impedance of the signal source; as indicated above, they may be entirely dissipative, entirely reflective, or in part dissipative and in part reflective. Consider the simple circuit shown in figure 12-156. The source impedance (Z_G) and the load impedance (Z_L) are complex and mismatched. The current delivered is given by:

$$I_o = \frac{V}{Z_G + Z_L}$$

and the corresponding power delivered is:

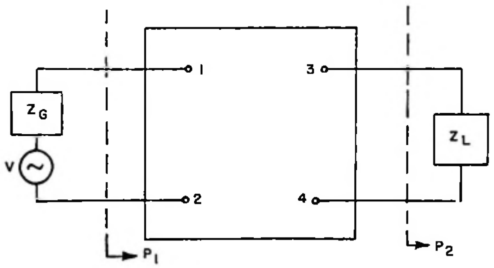


Figure 12-157. Passive Four-Terminal Circuit Inserted Between Two Mismatched Impedances

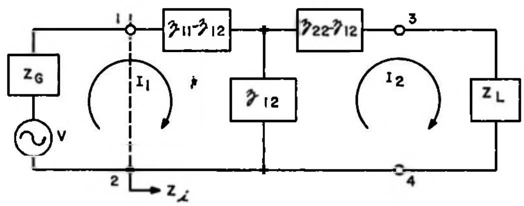


Figure 12-158. Open-Circuit Impedance Equivalence of Passive Four-Terminal Circuit of Figure 12-157

$$P_o = \left|I_o\right|^2 R_L$$

therefore

$$P_o = \frac{\left|V\right|^2}{\left|Z_G + Z_L\right|^2} R_L$$

where R_L represents the real part of the load impedance. Now insert a passive four-terminal circuit into figure 12-156, as illustrated in figure 12-157. Its equivalent circuit in terms of the open-circuit impedances of the four-terminal circuit is given in figure 12-158. These open-circuit impedances are defined as follows:

z_{11} = driving impedance of terminals 1-2 with terminals 3-4 opened.

z_{22} = driving impedance of terminals 3-4 with terminals 1-2 opened.

z_{12} = transfer impedance with terminals 3-4 open and terminals 1-2 excited with a driving voltage.

From figure 12-158, the driving point impedance (Z_i) is computed to be:

$$Z_i = \frac{z_{11}z_{22} - z_{12}^2 + z_{11}Z_L}{z_{22} + Z_L}$$

12-481. LOSSLESS FOUR-TERMINAL CIRCUIT.

12-482. When the passive four-terminal circuit is lossless, its open-circuit impedances are reactive functions. The power output is equal to the power input since no loss of power occurs in the circuit. Thus, from figure 12-158:

$$P_2 = \left|I_2\right|^2 R_L$$

or

$$P_2 = \left|I_2\right|^2 R_L = \left|I_1\right|^2 R_i$$

where R_i is the real part of Z_i contained in the formula of paragraph 12-480, and the current (I_1) is given by the equation:

$$I_1 = \frac{V}{Z_G + Z_i}$$

Therefore, the output power becomes:

$$P_2 = \frac{\left|V\right|^2}{\left|Z_G + Z_i\right|^2} R_i$$

Consequently, the insertion loss (A_r) can be found in terms of circuit impedances by substituting the equations for P_o and P_2 in the following equation:

$$A_r = 10 \log \frac{P_o}{P_2}$$

$$A_r = 10 \log \left\{ \frac{\left|Z_G + Z_i\right|^2}{\left|Z_G + Z_L\right|^2} \cdot \frac{R_L}{R_i} \right\}$$

This last equation represents reflective attenuation, and it may be transformed into a more convenient form by introducing the reflection factor of current (I_1) which is defined by:

$$\rho = \frac{Z_G - Z_i}{Z_G + Z_i}$$

For the usual case where the source and load impedances are purely resistive, that is, $Z_G = R_G$ and $Z_L = R_L$, the equation for insertion loss reduces to:

$$A_r = 10 \log \left\{ \frac{4R_G R_L}{\left|R_G + R_L\right|^2} \quad \frac{1}{1 - \left|\rho\right|^2} \right\}$$

where the reflection coefficient is:

$$\rho = \frac{R_G - Z_i}{R_G + Z_i}$$

Only under matched conditions, where $R_G = R_L$ and $Z_i = R_G$, does the reflective attenuation or reflective insertion loss reduce to zero.

12-483. LOSSY FOUR-TERMINAL CIRCUIT.

12-484. When the four-terminal circuit of figure 12-158 is lossy, the output power (P_2) can be evaluated by knowledge of the output current. It can be shown that this current is equivalent to:

$$I_2 = \frac{I_1 z_{12}}{z_{22} + Z_L}$$

By substituting for I_1, it becomes:

$$I_2 = \frac{V z_{12}}{(Z_G + Z_i)(z_{22} + Z_L)}$$

The insertion loss attenuation can now be represented by:

$$A = 10 \log \frac{\left|(Z_G + Z_i)(z_{22} + Z_L)\right|^2}{\left|Z_G + Z_L\right|^2 \left|z_{12}\right|^2} \text{ db}$$

It is interesting to note that the factors influencing attenuation contribute both dissipative and reflective losses; they can be separated into corresponding dissipative and reflective attenuation terms.

12-485. The dissipative attenuation term is found by matching the impedances. For the usual case, $Z_G = R_G = R_L = Z_L = Z_i$. The equation for the above insertion loss attenuation can be converted into the dissipative attenuation loss as follows:

$$A_{dis} = 10 \log \left| \frac{z_{22} + R_G}{z_{12}} \right|^2$$

The difference between the insertion loss (A) and the dissipative insertion loss (A_{dis}) represents the reflective attenuation, or:

$$A_r = A - A_{dis}$$

$$= 10 \log \left\{ \left| \frac{Z_G + Z_i}{Z_G + Z_L} \right|^2 \left| \frac{z_{22} + Z_L}{z_{22} + R_G} \right|^2 \right\}$$

12-486. A positive insertion loss represents power loss, while a negative insertion loss represents power gain. Power gain can occur in the cases described, since the amount of power delivered is dependent upon the impedance Z_L when the circuit is out, and upon the impedance Z_i when the circuit is in. To assure that an insertion loss is truly a loss, the power delivered to the load before the circuit is inserted must be defined as the maximum power which could be delivered to the load.

12-487. AF AND I-F ATTENUATORS.

12-488. GENERAL.

12-489. An attenuator is a four-terminal circuit whose insertion loss remains constant as the frequency varies. Departure from constant attenuation occurs only when the design limits of the constituent elements are exceeded. Attenuators are used to diminish delivered power or voltage by predetermined amounts, for the purpose of monitoring signals or calibrating communication equipment and components. To perform this basic function, an attenuator is often designed so that its input and output impedances are matched, within a prescribed frequency range, to the resistances between which it operates. Also, these resistances are usually matched to each other. In such cases, the designed attenuator becomes a bilateral device and its calibration holds true, regardless of which end is selected as the input terminal. However, in special cases involving narrow frequency bands, attenuators can be designed to provide a desired insertion loss when introduced between arbitrary impedances. Such designs are infrequent and quite complicated.

12-490. Matching is important for a variety of reasons. Since the oscillator frequency is usually dependent upon the input load provided by the circuit it feeds into, a constant input impedance prevents frequency drifts. For attenuators which require calibration, matching reduces the calibration error to a minimum. When attenuators are designed for known mismatches, a slight change in output load may appreciably alter the insertion loss. Such is not the case when attenuators are designed for match conditions. A slight change in output load results in an insertion loss error which is negligibly small.

12-491. LUMPED RESISTIVE ATTENUATORS.

12-492. Lumped resistive attenuators are constructed with pure resistances in order to realize attenuation which is frequency-insensitive. With the proper precautions, they can be designed to cover a frequency range extending into megacycles. They may be fixed or variable. An extensive treatment of resistive attenuators would cover such types as the T-type, π-type, H-type, balanced H-type, O-type, L-type, V-type, balanced V-type, bridged T-type, bridged H-type, and bridged balanced H-type attenuator circuits. The design of representative types will be elaborated upon in this section.

12-493. Care must be exercised in the interpretation of design equations. Many designs are not based upon <u>insertion loss</u> as defined in paragraphs 12-478 through 12-486, but rather upon the attenuation of power with the attenuator in the circuit, as considered in paragraphs 12-463 through 12-468. However, the difference between such attenuation and insertion loss may be negligible in most cases. Another precaution is to clearly note the matching criteria used by the designer. This, also, may differ in comparing various approaches.

12-494. POTENTIOMETER-TYPE ATTENUATOR. The potentiometer-type attenuators are commonly known as volume controls. They are widely used in electronic circuits, such as radio receivers and audio amplifiers. The precision types are constructed of resistance steps. They are very reliable, and they introduce negligible noise as the contact is moved. A typical arrangement is shown in figure 12-159 where voltage V_i is taken as coming from a constant-

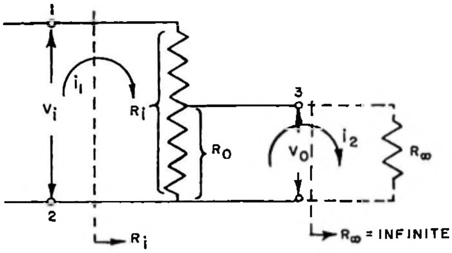


Figure 12-159. Potentiometer-Type Attenuator

voltage source with zero internal impedance; the resistance across the output voltage (V_O) is the input resistance to the grid of a tube, which is practically infinite in value. For the above circuit, the insertion loss may be easily computed by initially assuming that R_∞ is finite. Then the final expression can be found by letting R_∞ approach an infinite value. The power output with the four-terminal circuit removed is:

$$P_O = \frac{V_i^2}{R_\infty}$$

With the four-terminal circuit in place, the currents are quickly computed. Thus, the input current is:

$$i_1 = \frac{V_i}{R_i - R_O + \dfrac{R_O R_\infty}{R_O + R_\infty}}$$

$$= \frac{V_i (R_O + R_\infty)}{R_i R_O - R_O^2 + R_i R_\infty}$$

The output current becomes:

$$i_2 = \frac{i_1 R_O}{R_O + R_\infty}$$

$$= \frac{V_i R_O}{R_i R_O - R_O^2 + R_i R_\infty}$$

The output voltage is:

$$V_O = i_2 R_\infty = \frac{V_i R_O R_\infty}{R_i R_O - R_O^2 + R_i R_\infty}$$

The power output is found to be:

$$P_2 = \frac{V_O^2}{R_\infty} = \frac{V_i^2 R_O^2 R_\infty}{(R_i R_O - R_O^2 + R_i R_\infty)^2}$$

The corresponding insertion loss is:

$$A = 10 \log \frac{P_o}{P_2}$$

$$= 10 \log \left[\frac{\left(\dfrac{R_i R_O - R_O^2}{R_\infty} + R_i \right)^2}{R_O^2} \right]$$

As R_∞ approaches an infinite magnitude, the insertion loss reduces to:

$$A = 10 \log \left(\frac{R_i}{R_O} \right)^2 = 20 \log \left(\frac{R_i}{R_O} \right)$$

However, it is seen that for R_∞ as infinite:

$$\frac{V_i}{V_O} = \frac{R_i}{R_O}$$

Therefore, the insertion loss can be redefined as follows:

$$A = 20 \log \left(\frac{R_i}{R_O} \right) = 20 \log \left(\frac{V_i}{V_O} \right)$$

The potentiometer, therefore, acts as a voltage divider. This is true only when the imposed restrictions (that is, constant input voltage and infinite output impedance) prevail.

12-495. T-TYPE ATTENUATOR. The T-type attenuator consists of series and shunt elements. They are designed to operate be-

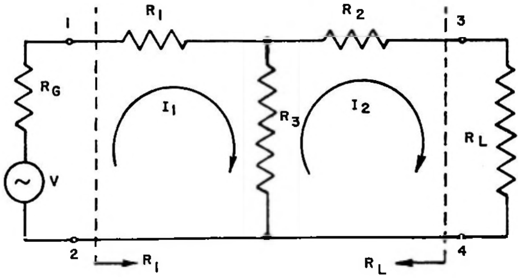


Figure 12-160. T-Type Attenuator

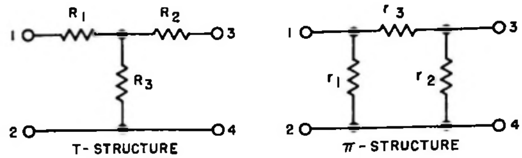


Figure 12-161. T and π Structures

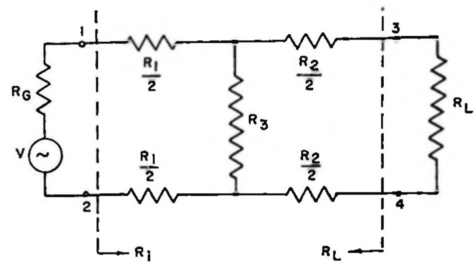


Figure 12-162. H-Type Attenuator

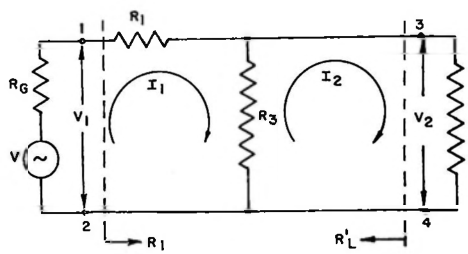


Figure 12-163. L-Type Attenuator, Type A

tween either equal or unequal resistances. Figure 12-160 shows a T-type attenuator connector between two resistances, R_G and R_L. R_i is the equivalent input resistance of the circuit shown across terminals 1 and 2. Its chosen value may vary with the design. R_L represents the equivalent input resistance of the circuit across terminals 3 and 4 with the emf of the signal source short-circuited. It is almost always selected to be equal to the load resistance.

12-496. π-TYPE ATTENUATOR. The configuration of the π-structure in comparison with the T-structure is illustrated in figure 12-161. The two structures can be made equivalent as far as the conditions at the terminals are concerned. These conditions involve equivalences between input resistances, currents, voltages, and power at the respective terminals.

12-497. H-TYPE ATTENUATOR. The H-type attenuator is shown in figure 12-162. It may be designed by dividing the series resistance elements of the T-type attenuator into equal halves and placing each half in a series leg. This type of arrangement is used whenever it is necessary to keep each side of the circuit balanced. The T-type attenuator is applicable where it is required to ground one side of the circuit.

12-498. L-TYPE ATTENUATOR. L-type attenuators are special cases of the T-type attenuators, as shown in figures 12-163 and 12-164.

12-499. L-TYPE DECADE ATTENUATOR. Cascading L-type structures (see figure 12-165) results in one form of variable ladder attenuator. It is designed to offer a constant resistance to one end only. Its chief use is in the output circuit of signal generators, where perfect matching is not important. In designing such attenuators, use is made of the characteristic resistance of the L-section in such a way that its input resistance looking toward the signal source remains constant as sections are added.

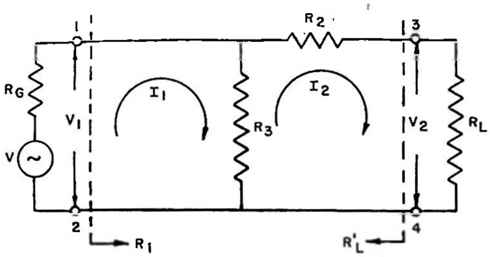


Figure 12-164. L-Type Attenuator, Type B

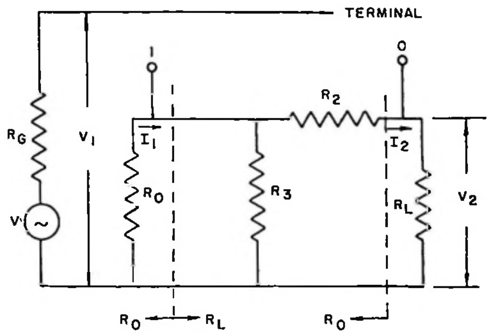


Figure 12-165. Section of L-Type Decade Attenuator

12-500. π-TYPE DECADE ATTENUATOR. An arrangement for cascading π-type structures is illustrated in figure 12-166. The elements of each section are determined from the designated terminal resistances and the required attenuation.

12-501. T-TYPE DECADE ATTENUATOR. The T-type decade attenuator is similar to the π-type in design. The difference is merely that each π structure of figure 12-166 is transformed into an equivalent T-structure.

12-502. MULTIPLE T-TYPE VARIABLE ATTENUATOR. The shunting resistance of a T-type structure decreases for increasing insertion loss. This information is used in the design of a variable attenuator known as the multiple T-type, shown in figure 12-167. It has the advantage of no power loss when the switch is in the position for zero attenuation.

12-503. ELECTRONIC ATTENUATOR. The electronic attenuator employs a cathode follower whose screen voltage, as well as its cathode resistance, may be varied. The circuit is illustrated in figure 12-168. The tube is a sharp cutoff pentode type with fixed grid bias. The output can be made to vary from 4 to about 90 db by adjusting the cathode resistors and the screen voltage. A range of 4 to about 70 db is realized with the cathode resistors, and an extended range of 20 db with the screen voltage control, if the input level is not too high.

12-504. I-F RESISTANCE ATTENUATOR. Much care is needed in constructing i-f resistive attenuators so that their functional

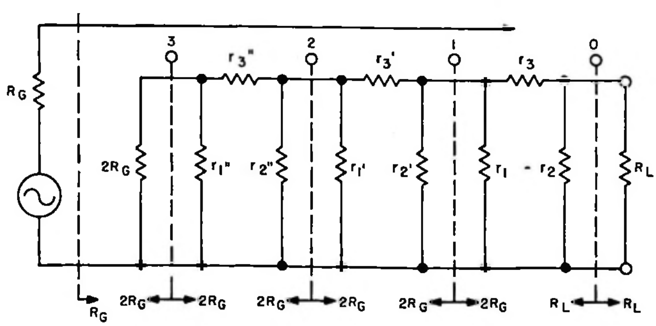


Figure 12-166. π-Type Decade Attenuator

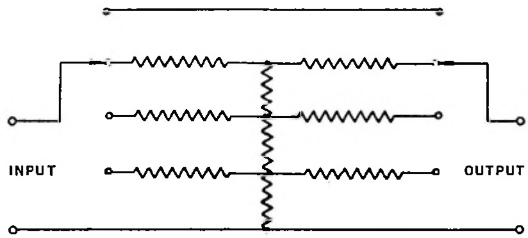


Figure 12-167. Multiple T-Type Attenuator

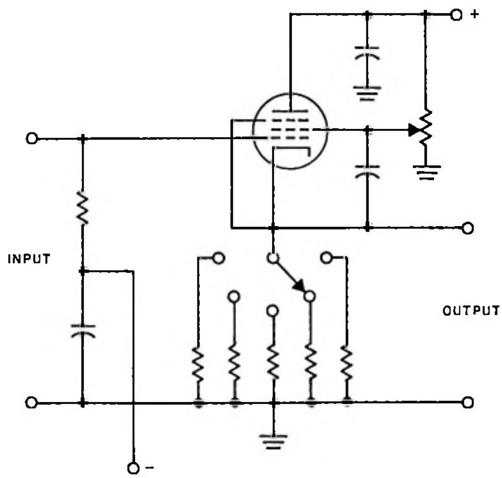


Figure 12-168. Cathode-Follower-Type Electronic Attenuator

attenuations lie within an acceptable range as the frequency is varied. You should use nonreactive resistances at i-f frequencies. You can fabricate such resistances with constant-temperature-coefficient resistance wire, wound on thin mica cards. The Ayrton-Perry method of winding should be employed to reduce self-capacitance and inductance. This method uses two windings wound in the opposite sense in order to create opposing magnetomotive forces. They are connected in parallel at the two ends. Self-capacitance is decreased by having the insulation of the two windings touch each other with a minimum area of contact, and by having the terminals at opposite ends of

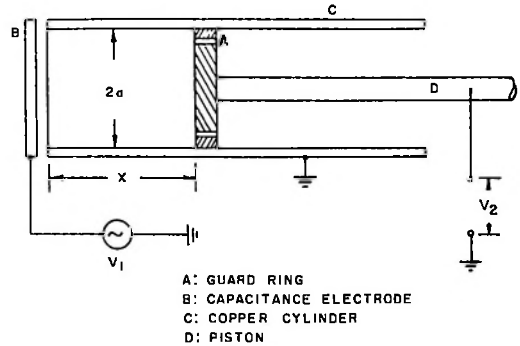


Figure 12-169. Piston-Type Capacitive Attenuator

the cord. In order to reduce the inductive reactance at i-f frequencies, make sure that the wires are very small and quite close together.

12-505. When attenuators are placed in cascade, capacitive effects exist among the various terminals. You can reduce these to a minimum by providing the sections with individual shields tied together to the common side of the equipment. However, each individual unit is still by-passed by the direct capacitance between its input and output switches. You can eliminate this by placing a shield between each pair of switches.

12-506. With increasing frequency, the distributed capacitance of each series resistance causes the attenuation to decrease progressively. However, the distributed capacitance of each shunt resistance produces a rising attenuation characteristic. You can make these two effects cancel. You can also add compensating capacitances to extend the operating frequency range.

12-507. REACTIVE ATTENUATORS.

12-508. CAPACITIVE ATTENUATORS. An example of a capacitive attenuator for the testing of radio receivers is illustrated in figure 12-169. The capacitance of the unit

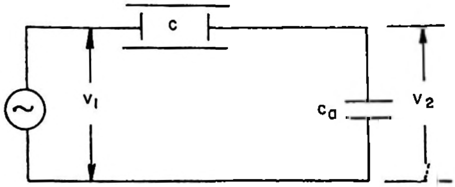


Figure 12-170. Piston-Type Capacitive Attenuator Loaded with Capacitance, Simplified Equivalent Circuit

is controlled by the piston position. When placed between a signal source and a capacitive load, the output voltage in figure 12-170 is given by the expression:

$$V_2 = \frac{V_1 C}{C_a + C}$$

where C_a is the load capacitance and C is the capacitance introduced by the piston capacitor. For C, which is very much smaller than C_a, the voltage ratio becomes:

$$\frac{V_2}{V_1} = \frac{C}{C_a}$$

12-509. Another type of capacitive attenuator, shown in figure 12-171 in a transverse section view, uses two metallic coaxial cylinders attached to a common shaft by means of end plates. The inner cylinder has an axial length which is less than that of the outer cylinder, so that the former is effectively shielded by, and confined within, the outer structure. The inner structure is free to move, while the outer structure is fixed in position. The movement is controlled by means of the rack and pinion shown. Each cylinder is divided into two sections, which are insulated from each other. The important capacitances are indicated by the broken-line construction with respect to the input and output terminals. They are drawn schematically in figure 12-172. Capacitances C_1 and C_2 remain fixed as the inner cylinder moves counterclockwise.

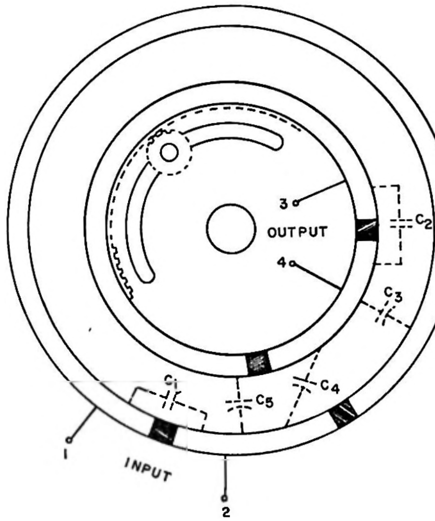


Figure 12-171. Relative-Type Capacitive Attenuator

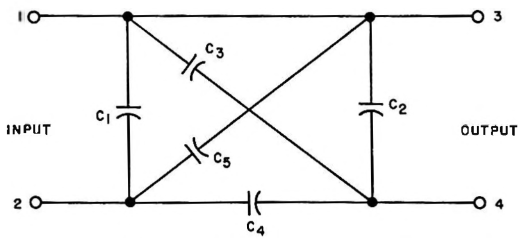


Figure 12-172. Schematic Diagram of Attenuator Shown in Figure 12-171

wise. However, capacitance C_4 decreases while capacitances C_3 and C_5 increase. Consequently, capacitances C_3 and C_5 vary simultaneously in one sense, and capacitance C_4 varies in an opposite sense, while the total capacitance between the two cylinders remains constant. Therefore, the effective input and output capacitances remain substantially constant despite any change in C_4. You can use this arrangement to control the attenuation of voltages.

12-510. INDUCTIVE ATTENUATORS. A mutual-inductive attenuator of the piston

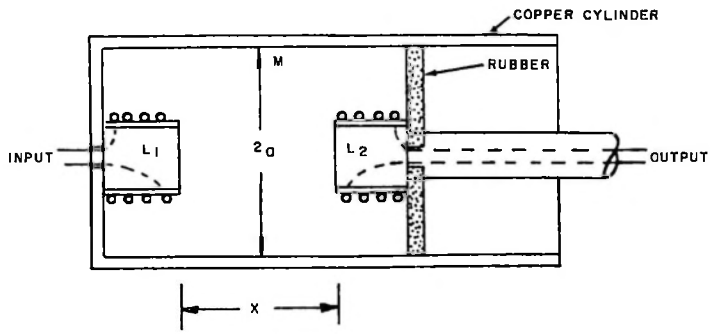


Figure 12-173. Inductive Attenuator

type is shown in figure 12-173. The mutual inductance varies with position x, which is controlled by the piston. The output voltage decreases as x increases.

12-511. Microwave attenuators are of two types: reflective and dissipative. The first type employs a waveguide which is operated below its cutoff frequency. Attenuation is achieved through mismatch, which reduces the power output by reflecting a portion of the incident power. The amount of mismatch is dependent upon the length and size of the waveguide. This technique is also applicable at rf frequencies. In the dissipative type, the difference between the output power levels is absorbed within the transmission system.

12-512. The electrical behavior of a waveguide is described by its phase constant, its attenuation constant, and its characteristic impedance, symbolized, respectively, by β, α, and Z_0. The combination of α, measured in nepers per unit length, and β, measured in radians per unit length, is known as the propagation constant, γ, or:

$$\gamma = \alpha + j\beta$$

where J equals an angle of 90 degrees. For most practical transmission facilities, γ is assumed to be an extremely low loss or zero. Only when α is large does the waveguide attenuate.

12-513. The electromagnetic field generated within a transmission facility is completely described by a summation of modes. These modes are characterized as transverse electromagnetic (tem), transverse electric (te), and transverse magnetic (tm).

12-514. In coaxial lines, propagation is restructed to the tem mode by spacing the conductors so that the distance of separation is much less than the wavelength of the exciting frequency. However, as the frequency increases, higher modes (te and tm) may propagate, since the behavior of these modes is similar to that of a high-pass filter. Each higher mode possesses a cutoff frequency. The higher mode, whose cutoff frequency is the smallest, is referred to as the dominant mode. The tem mode is known as the principal mode. When the propagating mode is entirely tem, lumped parameters per unit length may be defined as:

L = series inductance

R = series resistance

C = shunt capacitance

G = shunt conductance

The propagation constant and the characteristic impedance are given, respectively, for a lossy guide by:

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)}$$

$$= \alpha + j\beta$$

$$Z_O = \sqrt{\frac{R + j\omega L}{G + j\omega C}} = R_O + jX_O$$

For a lossless guide, they simplify to:

$$\gamma = j\,\frac{2\pi}{\lambda}$$

and

$$Z_O = R_O = \sqrt{\frac{L}{C}}$$

where $\omega = 2\pi f$ (f is the operating frequency in cps)

μ = permeability of the medium (henries per meter)

λ = wavelength in the medium

$$= \frac{1}{\sqrt{\mu\epsilon}} \cdot \frac{1}{\mu\epsilon} \text{ meters}$$

In waveguides, since the tem mode cannot exist, propagation is by means of the dominant mode. You prevent all other higher modes from propagating by choosing suitable dimensions. In a lossless guide:

$$\beta = \frac{2\pi}{\lambda_g}$$

where λ_g = wavelength in meters measured in the guide. It is equivalent to:

$$\lambda_g = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{\lambda_c}\right)^2}}$$

where

λ_c = the cutoff wavelength of the mode

λ = the wavelength of the medium with which the guide is filled (for example, air).

12-515. The analytic treatment of both coaxial lines and waveguides is similar. It involves only the dimension along the axis of propagation in terms of the α, β, and Z_O factors. The power transmitted by such components is dependent upon their attenuation constant and the mismatches introduced by the source and load impedances.

12-516. Reflective attenuators are commonly known as cutoff attenuators, or piston attenuators. They are composed of waveguide lines operating well below the cutoff frequencies of the generated modes, so that energy does not propagate. The propagation constant becomes a real quantity in the operating region, or:

$$\gamma = \alpha$$

The characteristic impedance (Z_O) reduces to a pure reactance, or:

$$Z_O = jX_O$$

Each generated mode decays exponentially from the launching end to the receiving end, and the attenuation rate corresponds to its attenuation constant. In such a line, the dominant mode possesses the least rate of attenuation. Since the higher modes decay rather rapidly, the dominant mode tends to take over after the two ends of the cutoff waveguide are separated by a critical spacing. Beyond this critical spacing, the attenuator possesses an attenuation rate which corresponds to the attenuation constant of the dominant mode of the waveguide. This constant is given by:

$$\alpha = \frac{2\pi}{\lambda_c}\sqrt{1 - \left(\frac{\lambda_c}{\lambda}\right)^2} \text{ nepers per unit length}$$

for $\frac{\lambda_c}{\lambda} = 1.05$, $\alpha \approx 2/\lambda_c$; where λ_c is the cutoff wavelength of the mode and λ is the wavelength of the frequency used. This device has the great advantage of yielding absolute attenuation, which is dependent at a given frequency on the cross-section dimensions of the guide, and the separation between input and output. It dissipates practically no power. Considering the various types of available reflective attenuators, the circular waveguide is preferred because of its ease of construction, the dominant mode being either tm or te.

12-517. In the circular attenuator, the tem mode propagates through a coaxial section to the input end of the cutoff waveguide where cutoff guide modes are excited. Assuming that the input and output ends are properly separated, so that all modes higher than the dominant mode will decay to zero ideally, the decaying dominant mode couples to the output end, where power is absorbed in proportion to either the square of the electric field or the square of the magnetic field. Thus, if you vary the distance of separation between the two ends, the output power decreases with respect to the input power, the difference being reflected from the input end. A fixed dissipative attenuator is normally used to absorb this reflected power. In the tm mode design, the launching and receiving ends of the guide consist of circular metal disks. In the te mode design, they are metal loops which terminate the coaxial sections, as illustrated in figure 12-174. The former is a capacitive type, while the latter is an inductive type.

12-518. The dominant decaying mode in these cutoff attenuators is not a wave motion. Consequently, the wavelength is infinitely long and there is no phase change along the axis of decay. The time variation is simultaneous everywhere.

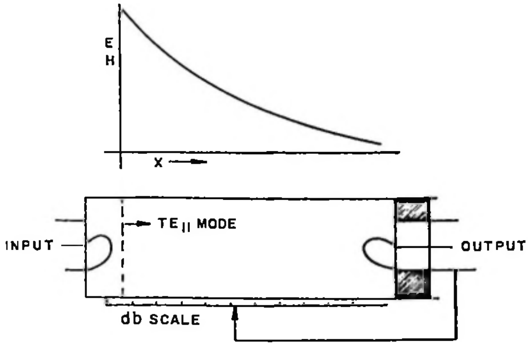


Figure 12-174. TE_{11} Mode Cutoff Attenuator

12-519. The two major design problems are matching and mode purity. Both influence the minimum insertion loss, while the latter distorts the calibration curve. These attenuators are most useful wherever low insertion loss is not required, since the minimum insertion loss, including dissipative matching, approaches 30 db for a well-matched linear attenuator. The linearity of cutoff attenuators is not only a function of the modes present, but also a function of the loading at the launching end by the termination connected to the receiving end.

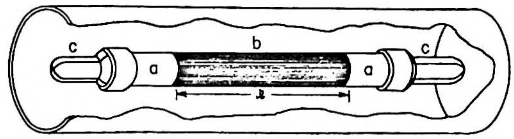
12-520. Coaxial and waveguide dissipative attenuators employ a lossy substance to absorb power. In earlier designs a mixture of carbon was used, either as a lossy dielectric or as a lossy coating. Because of their susceptibility to aging and atmospheric changes, carbon attenuators were not suitable as precision components. Consequently, resistive metallic films were developed for metalized glass attenuators. Metals in the form of very thin films were found to provide a low temperature coefficient and to be insensitive to atmospheric influences and aging. Such attenuators were observed to be electrically and mechanically stable. In the development of metallic film elements, glass was selected as a base material because its melting point is most usable for metalizing techniques, its surface is smooth, thereby reducing the possibility of film breaks, it does not chemically react with the film, it is not hygroscopic, and it will not warp or change shape.

12-521. Two metalizing techniques have been developed in the fabrication of such attenuators. In one process, a metalizing solution containing metal-organic compounds is spread evenly over the surface of the glass. The painted glass is then baked at a high temperature until the organic material decomposes and the resulting metallic film embeds itself into the surface of the glass. In the second process, the metallic film is evaporated upon a glass surface in an evacuated chamber until the desired resistance value is obtained. Baked-on films are composed of platinum and palladium in order to realize a low thermal coefficient of resistance. Evaporated films are of either chromium or nichrome, with a protective film of magnesium fluoride to prevent abrasions.

12-522. Resistive films resulting from either process are extremely thin and appear to be transparent. The film thickness is less than the depth of current penetration, even at the highest microwave frequencies. Hence, its dc resistance is equivalent to its microwave resistance. This facilitates the design and fabrication of lossy elements for coaxial and waveguide attenuators.

12-523. COAXIAL ATTENUATORS.

12-524. FIXED. Coaxial attenuators are so designed as to bilaterally match over a broad frequency range the characteristic impedance of the line in which they are inserted. Their attenuation response remains flat within a narrow tolerance range. The various configurations used are described as the series types, the T-type, and the chimney type. The maximum frequency design limit is approximately 10, 000 mc.

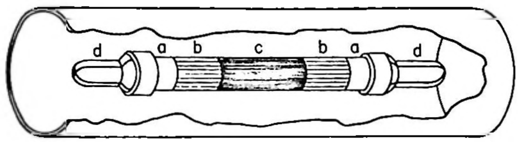


a: GLASS TUBING

b: METALIZED SECTION

c: PLUG-IN BULLET

A
SINGLE SECTION COAXIAL ATTENUATOR



a: GLASS TUBING

b: METALIZED MATCHING SECTION

c: MAIN ATTENUATION SECTION

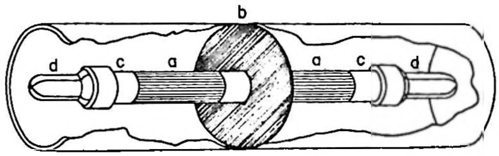
d: PLUG-IN BULLET

B
COMPENSATED COAXIAL ATTENUATOR

Figure 12-175. Series-Type Coaxial Attenuators

12-525. The series type consists of either a single, uniform section of metalized glass for small attenuations or a series of three uniform sections, one at either end for matching and a main attenuating section in the center, to achieve large attenuations. The two types are illustrated in figure 12-175. As shown, the ends of the metalized glass tubing are soldered to bullets so that each unit can be inserted as the center conductor of a coaxial casing.

12-526. The electrical design of a T-type or disk-type coaxial attenuator (as shown in figure 12-176) is similar to the design of low-frequency T sections, discussed in paragraphs 12-487 through 12-510. It is most useful up to approximately 4000 mc. Since at the higher frequencies the size of the



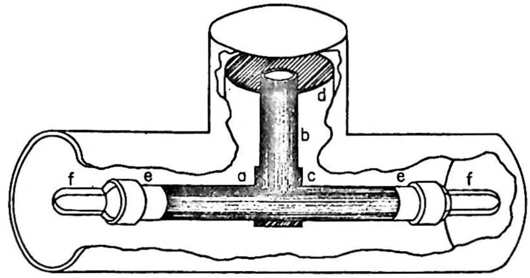
a: SERIES METALIZED SECTION
b: SHUNT METALIZED DISK
c: GLASS TUBING
d: PLUG-IN BULLET

Figure 12-176. T-Type Coaxial Attenuator

resistive elements become comparable to the wavelength, a point is reached at which they can no longer be considered as lumped components. These types are most suitable over the frequency band of 1000 to 4000 mc for attenuations greater than 10 db, and the swr is usually around 1.3.

12-527. The shunt element may consist of either mica or glass upon which is evaporated a metallic film. It provides a large surface area for realizing the resistance necessary for the design. Below 10 db, this resistance assumes large values and the corresponding film may often be so thin as to discourage fabrication.

12-528. Chimney-type coaxial attenuators use either one or two lossy stub lines as matching circuits for the design of broadband attenuators. A single-chimney unit is illustrated in figure 12-177. This type is similar to the T-type coaxial attenuator. A stub line appears in place of the shunt disk. It provides a way of realizing attenuations on the order of 10 db or less without being restricted by films which are too thin. However, in order for the attenuator to remain matched over a prescribed frequency range, the input impedance presented by the stub arm must remain real despite the fact that it is terminated in a short circuit. For small values of $1/\lambda$, the reactance of a



a: SERIES METALIZED SECTION
b: CHIMNEY METALIZED SECTION
c: CHIMNEY HOLDER
d: SHORTING PLATE
e: GLASS TUBING
f: PLUG-IN BULLET

Figure 12-177. Single-Chimney Coaxial Attenuator

short-circuited line is very nearly zero for the condition that:

$$\frac{R_p}{R_o} = \sqrt{3}$$

and

$$R_o = 60 \ln \frac{D}{d} \text{ ohms}$$

where:

R_p = total resistance of shunt metalized glass section

R_o = characteristic impedance of short-circuited lossless stub

D = inner diameter of outside line

d = outer diameter of inside line (metalized glass)

NOTE

Once R_p and d are known, you can compute D. Attenuators using a single chimney operate over the frequency range of 0 to 4000 with swr's of less than 1.3.

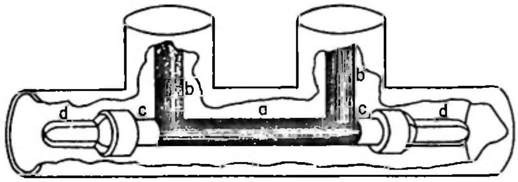


Figure 12-178. Double-Chimney Coaxial Attenuator

12-529. The double-chimney coaxial attenuator is shown in figure 12-178. You can use this type of line for large attenuations with broadband frequency responses. You can use the stub sections for attenuation equalization as well as for matching.

12-530. The design procedure consists of determining the input driving-point impedance function of the microwave device when properly terminated, approximating this function by an equivalent lumped circuit representation, then determining an ideal lumped matching circuit arrangement which (in combination with the above circuit) produces a constant resistance input impedance, and, finally, transforming the lumped matching circuit arrangement into the final microwave structure.

12-531. VARIABLE. Variable dissipative coaxial attenuators are devices which employ a means for shorting out parts of a metalized-glass element. The resulting attenuation is proportional to the film which remains exposed to the microwave power. The lossy element may be placed either in series or in shunt within a coaxial casing.

12-532. Three arrangements have been developed. The first employs a metal rod which is inserted into the metalized-glass tubing from the far end of the attenuator. The attenuation is dependent on the position of the rod. The coupling between the film and the rod is through the glass. The second type uses a close-fitting metal tube which slides over the metalized-glass element. In the third type, a metalized-glass plate is inserted in shunt through a slot in the outer conductor of a coaxial line; a variation of this is to fix the lossy plate between the outer and inner conductors and insert a movable short circuit from the outer conductor.

12-533. WAVEGUIDE ATTENUATORS.

12-534. Both fixed and variable attenuators for rectangular waveguides usually employ resistive plates inserted parallel to the electric field of the propagating te_{10} mode. In precision types, the resistive element is a metalized-glass plate which may be inserted either from one of the narrow side walls or through a slot milled in the center of the upper wall. The former is referred to as the vane type, or transverse insertion type. The latter is known as the flap type, or guillotine insertion type. Both types of design provide bilateral matching. A rotary type may also be designed by using a metalized-film element which rotates within a round guide flanked by rectangular-to-round transition.

12-535. VANE-TYPE. The vane-type attenuator makes use of an element which has matching sections near both ends; basic matching methods are illustrated in figure 12-179. The exact dimensions are determined experimentally. For a mean waveguide wavelength, λ_g, resistive transformer matching sections are approximately $\lambda_g/4$ in length. The taper match is approximately $\lambda_g/2$ long. Of the two methods, the resistive transformer match is preferred, since it results in broadband responses and is simpler to fabricate. The design of metalized-glass plates is based upon a systematic variation of plate dimensions, a plate dielectric, and the surface

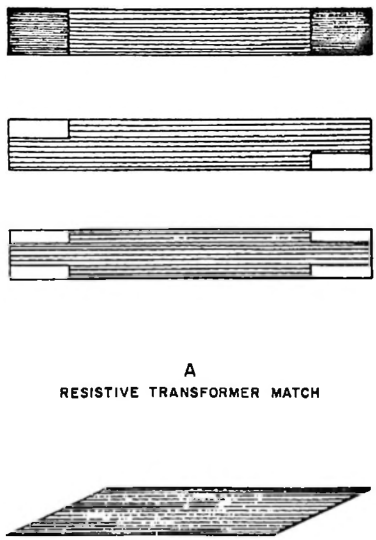


Figure 12-179. Methods for Measuring Vane-Type Attenuators

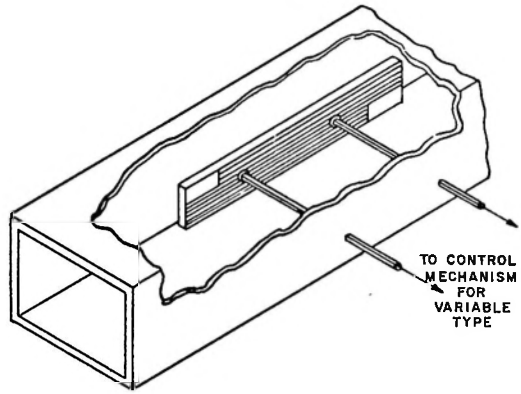


Figure 12-180. Vane-Type Waveguide Attenuator

resistivity of the metallic film until desired attenuation and swr characteristics are realized with minimum frequency sensitivity.

12-536. The casing design incorporates properly spaced metal struts which support the metalized-glass element. Usually, two struts are used, spaced an odd multiple of the mean waveguide wavelength apart, since they introduce shunting capacitances. For variable attenuators, a precision screw mechanism controls the insertion of the plate within the guide. A typical structure is shown in figure 12-180. These attenuators have been designed to provide an attenuation range of 0 to 40 db with swr's of less than 1.12 for signals as high as 18,000 mc.

12-537. FLAP-TYPE. The resistive element for flap-type attenuators is shaped as an arc of a circle to achieve bilateral matching. This is shown by the arrow in figure 12-181. In precision types, the casing design utilizes a narrow slot with polyiron chokes and a precision mechanism which lowers the glass plate into the guide. Since the attenuation of the plate increases exponentially, a nonlinear drive is used, as illustrated in figure 12-181.

12-538. Above 18,000 mc, the dimensions of waveguides approach small values. Therefore, the flap-type is simpler to design than the vane-type. Both glass and mica are used as a base for the metallic film. Flap-designed attenuators yield attenuation ranges of 0 to 40 db for the frequency range from 18,000 to 40,000 mc.

12-539. ROTARY-TYPE. The rotary waveguide attenuator consists of three sections of waveguides, two end sections made up of rectangular-to-round transitions and a center section of round waveguide. Choke joints permit you to rotate the center section axially. Each section has a resistive film, evaporated on thin mica, stretched across a diameter. The films at each end are in line with each other and fixed at right angles to the E-field of the te_{10} dominant mode. When all three films are in line, the attenuation is zero.

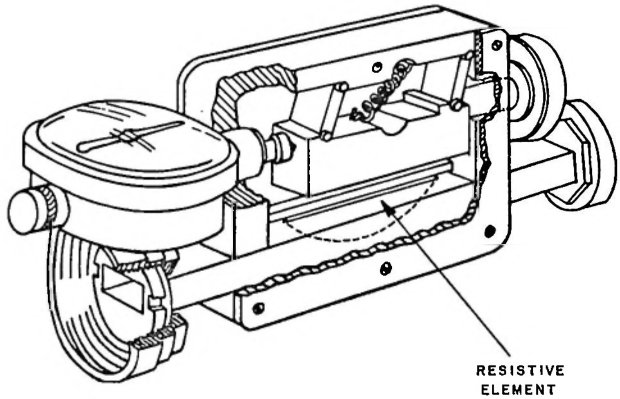


Figure 12-181. Precision Flap-Type Waveguide Attenuator

12-540. When you rotate the center section through an angle, θ , the electric field may be resolved into two components: E sin θ parallel to the film, and E cos θ perpendicular to the film. The function of the center film is to completely attenuate the sine component so that the output field is E cos θ. The purpose of the output-end film is to restore the original polarization. The E cos θ component splits into two parts: E cos θ sin θ is absorbed completely by the film, and E cos θ cos θ appears at the output terminals of the attenuator. Hence, the total attenuation is given by the following law:

$$A = 40 \log (\cos \theta)$$

12-541. The input-end film is included to make the unit bilateral. The above law is modified by the effect of reflections from the film ends and by the finite attenuation of the center film. This modification is substantial at attenuation settings above 60 db.

12-542. ATTENUATION MEASUREMENTS BELOW MICROWAVE FREQUENCIES.

12-543. GENERAL.

12-544. Direct methods involve the use of power meters, and at low power levels such methods are not preferred since they may introduce calibration errors. Since insertion loss assumes a constant load, it can, therefore, be expressed as the ratio of the magnitude of the output voltages with and without the unknown attenuating device in the circuit.

$$A = 10 \log \frac{P_0}{P_2} = 20 \log \frac{V_0}{V_2}$$

Consequently, vacuum-tube voltmeters can be employed to measure insertion loss, assuming that their loading effects are negligible and that their calibrations are accurate, even at very low voltages. Therefore, the most convenient and accurate methods are, essentially, comparison procedures.

12-545. LOW FREQUENCIES.

12-546. The circuit of figure 12-182 represents a substitution method for measuring attenuation in the low-frequency range. In this circuit there are two channels; one channel includes the unknown component under test; the second channel contains the reference standard, which is a precision variable attenuator composed of resistive elements. The output of each channel may

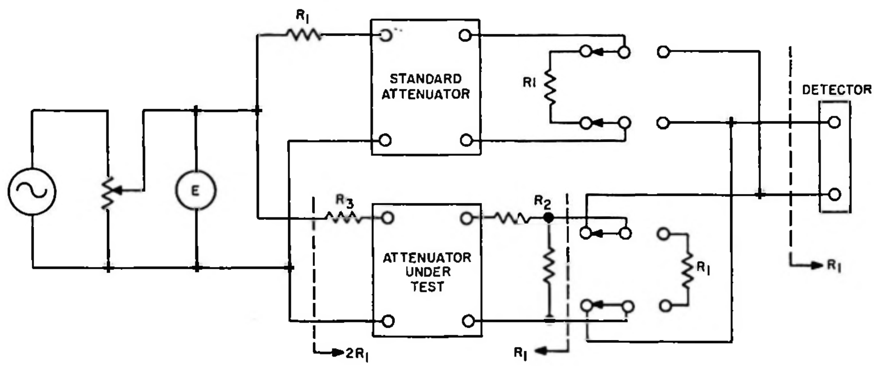


Figure 12-182. Arrangement for Measuring Attenuation at Low Frequencies

be connected to either a detector having a resistance equal to the matching resistance of the standard attenuator, or to a fixed resistance (R_1) of the same value.

12-547. If the unknown under test has the same matching resistance as a standard attenuator, you must make the input resistances, R_1 and R_3, equal and omit the matching element, R_2. For this condition, the two channels will be identical when the attenuations correspond. Therefore, the method of measurement is to switch the detector to one channel and then to the other channel, adjusting the standard attenuator until an equal output is obtained for either channel. The setting of the standard attenuator reads directly the attenuation of the unknown under test. The total input resistance of the circuit will remain unchanged during the switching sequence. When the unknown under test has a characteristic resistance which differs from that of the standard, the input resistance (R_3) and the matching elements (R_2) are adjusted so that the circuit still reads directly. The frequency range with this method extends to approximately 300 kilocycles per second.

12-548. INTERMEDIATE FREQUENCIES.

12-549. A method of measuring the attenuation of twisted pair transmission lines at frequencies ranging from 1 to 15 mc is illustrated in figure 12-183. The resistor, R, is used to match the characteristic impedance of the unknown line to that of the attenuator.

12-550. After you connect the equipment as shown in figure 12-183 and set the receiver sensitivity to correspond to normal receiving conditions, adjust the attenuator for a reasonable output level. Then insert the line sample between the receiver and the attenuator, as illustrated in figure 12-183. Adjust the attenuator to bring the receiver output back to the level noted above. Assuming that the receiver and oscillator remain stable, the attenuator change will correspond to the attenuation of the line sample. This technique may be extended to measure attenuations of other components besides transmission lines and cables.

12-551. VERY HIGH FREQUENCIES.

12-552. The vhf method to be described is particularly useful in measuring the attenu-

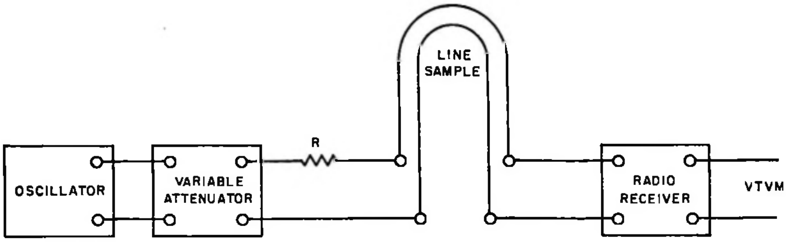


Figure 12-183. Measurement of Attenuation at Radio Frequencies, Equipment Arrangement

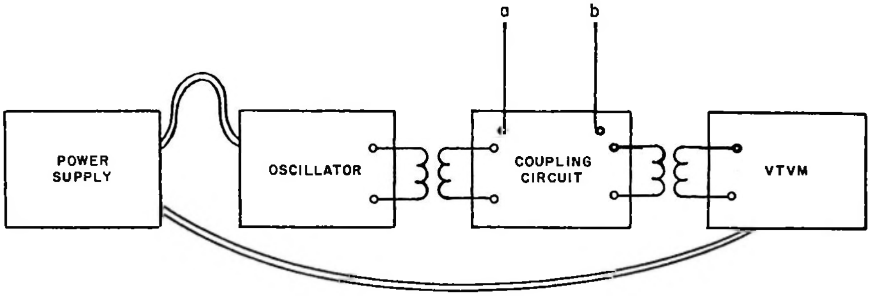


Figure 12-184. VHF Setup for Attenuation Measurements

ation of cables from 100 to 400 mc. It employs a coupling loop circuit which establishes and maintains matched conditions. The arrangement is shown in figure 12-184. In use, you plug the cable in, tune the oscillator to the desired frequency, and adjust the coupling loop circuit so that the equipment is matched. Figure 12-185 shows a simplified schematic representation of the oscillator and the coupling circuit. You must carefully shield the oscillator circuit to prevent signals other than those carried by the transmission line from reaching the vacuum-tube voltmeter, and loosely couple the coupling coil to the tank circuit. The coil should be almost completely external to the shield. The equivalent circuit of the oscillator and coupling is shown in figure 12-186. As indicated, the presence of the transmission line introduces a series resistance, R_R, into the generator coupling loop circuit. This reflected resistance has a magnitude which is controlled by varying the mutual inductance. Figure 12-187 shows L_3, C_3, and R_R as an equivalent impedance consisting of a capacitance, C, in series with a resistance, R_1. By adjusting C_3 so that the loop circuit is in resonance, the total reactance of the circuit becomes zero. By varying the coupling, an impedance match can be achieved. This occurs when $R = R_1$, and is indicated by a maximum reading on the vacuum-tube voltmeter at the far end of the line. You can vary the coupling manually by moving one component toward the other.

12-553. By a similar process, the transmission line can be terminated by a pure resistance equivalent to the characteristic impedance of the line. A simplified schematic of the transmission-line and voltmeter cir-

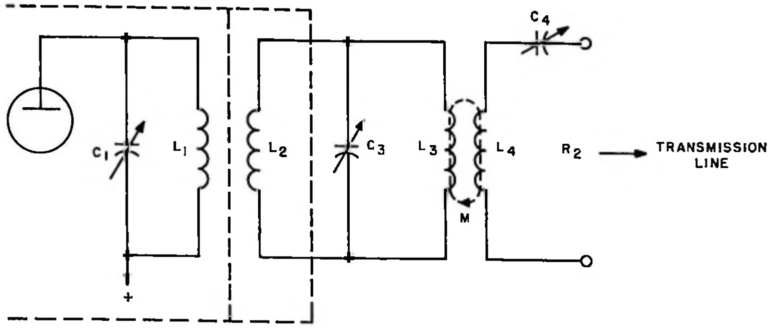


Figure 12-185. Generator Coupling Circuit

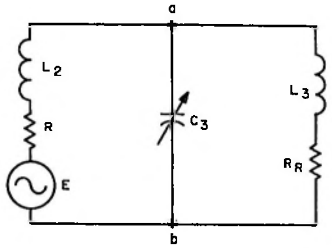


Figure 12-186. Equivalent of Generator Coupling Circuit

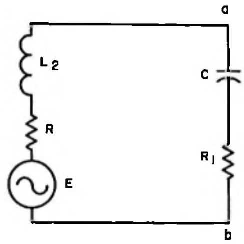


Figure 12-187. Equivalent of Figure 12-186

cuit is shown in figure 12-188. The impedance match is accomplished by varying the coupling between the two components until a maximum reading is indicated on the vacuum-tube voltmeter.

12-554. In setting up the equipment for a desired frequency, select an appropriate oscillator, vacuum-tube voltmeter, and coupling unit. Arrange these components as shown in figure 12-184, and plug in the cable. Tune the oscillator to the desired frequency and, in the following order, tune the oscillator coupling, the coupling unit, and the voltmeter circuit. The vacuum-tube voltmeter must be calibrated to read directly in db. First the attenuation of a cable sample 2 feet in length is measured, and then that of a cable sample 102 feet in length. The difference between these two readings represents the attenuation of 100 feet of cable.

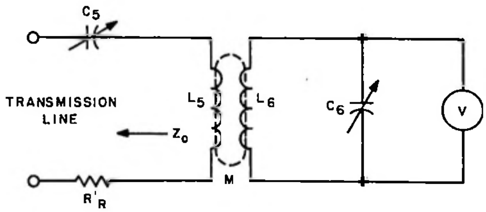


Figure 12-188. Vacuum-Tube Voltmeter Coupling Loop Circuit

Another procedure is to set the voltmeter reading to zero for the 2-foot sample so that the reading obtained for the 102-foot sample will be a direct measure of the attenuation of 100 feet of cable.

12-555. ATTENUATION MEASUREMENTS AT MICROWAVE FREQUENCIES.

12-556. A variety of transmission methods have been devised for measuring insertion loss at microwave frequencies. The method of detection and the type of reference standard serve to distinguish the various measuring schemes.

12-557. MICROWAVE DETECTORS AND AMPLIFIERS FOR RELATIVE POWER MEASUREMENTS.

12-558. A convenient and accurate method for measuring relative powers at microwave frequencies is based upon amplitude-modulating the microwave source with square waves and detecting by means of either crystals or bolometers.

12-559. CRYSTAL DETECTOR. The crystal detector is a semiconductor diode of silicon with a tungsten cat whisker. It acts as a square-law rectifier, and yields an output current proportional to the input power. When you use it with a modulated source, you must also provide a narrow-band amplifier set to the modulation frequency, to amplify the signal. The crystal should have a sensitivity of 5000 microvolts audio per microwatt of microwave power, and the minimum detectable signal power should be 1.8 x 10^{-6} microwatts for a single-cycle bandwidth. These factors will make the crystal ideal for measurements of small attenuations. For accurate measurements, this type detector has a limit of approximately 30 db. You can extend this limit by using special calibrating techniques.

12-560. BOLOMETER DETECTOR. A bolometer is a thermally sensitive device, usually consisting of a short length of Wollaston wire. This wire is produced by taking a short length of normal cross-section platinum wire and coating it with silver to about ten times its previous diameter. The composite wire is then stretched to reduce the cross-section of the platinum to the required value. The silver is finally etched off over the desired length.

12-561. As a detector, the bolometer resistance changes as it absorbs power. The change is a function of the power dissipated. The period corresponding to the modulating frequency is selected to be large as compared with the thermal time constant of the bolometer. As a consequence, the resistance change will follow the microwave power level at the modulation rate. Since the bolometer is biased from a constant-current source, the resistance variations will appear across the terminals as an audio voltage. The frequency of this voltage is that of the modulation signal and the amplitude of the voltage is proportional to the microwave power level. Typical time constants of Wollaston bolometers range from 80 to 250 microseconds. Therefore, modulating frequencies are limited to approximately 1000 cps.

12-562. Bolometers are normally biased to an operating resistance of 100 to 400 ohms, with 200 ohms being the most typical. For this case, the bias current is approximately 5 ma. For a milliwatt of microwave power superimposed on its dc bias power, the resistance change is about 10 ohms. For 100 percent square-wave modulation, the 10-ohm variation produces an audio square-wave output voltage of about 47 millivolts peak to peak. Its fundamental frequency component amplitude presented to a narrow-band amplifier is about 21 millivolts.

12-563. The resistance-power characteristic is assumed to be linear. However, for deviations of 2 percent from linearity, you should operate the bolometer at a power level below 200 microwatts. Its sensitivity is 21 microvolts audio per microwatt. The minimum detectable signal power for bolometers is 1 x 10^{-4} microwatts for a single-cycle bandwidth. The db range from noise level to 2 percent deviation from square law is about 50 db.

12-564. NARROW-BAND AMPLIFIER. The design of tuned or narrow-band amplifiers for crystal or bolometer input requires careful shielding and grounding, since the cable leading to the casings or mounts may cause pickup. Because of low input impedance and high gain, these small pickups may appear as disturbing influences. You can often achieve a narrow pass band by means of a twin-T feedback circuit applied to the second stage. Such an arrangement can be made tunable with a single control, thus permitting you to peak the gain at any desired frequency.

12-565. For bolometers normally biased to 200 ohms, a step-up matching transformer can be used at the input to the first tube. During operation, mismatches may result in significant errors. Therefore, the complete amplifier design is based upon considerations involving bandpass, center frequency, detector sensitivity, detector noise level, hum, and accuracy. The noise level of most amplifiers is approximately 0.02 microvolt when the input is loaded with a 200-ohm bolometer and the attenuation range is around 60 db. The meter scales should be calibrated square law, in units of both swr and db. Most amplifiers contain bolometer bias circuits. To reduce hum, you can battery-operate them.

12-566. SUPERHETERODYNE RECEIVER. Microwave superheterodyne receivers consist of a mixer and local oscillator coupled to an i-f amplifier, a linear second detector, a video amplifier, and an audio amplifier. Tuning is accomplished largely by the local oscillator. The main receiver gain of approximately 100 db appears in the i-f amplifier. The principal advantages are linear response and sensitivity to small signals on the order of 10^{-13} watt. The spectrum analyzer is basically just such a receiver, and is very useful as a power indicator for the measurement of large attenuations.

12-567. An ac vacuum-tube voltmeter such as the ME-6D is useful and accurate for measuring the af voltage which appears across the terminals of the bolometer detector. If the db scale is calibrated for linear detection in terms of 20 log $\frac{V_1}{V_2}$, for correct results involving square-law detection, you must divide the db scale reading by 2. This division is unnecessary for meters which are calibrated for use with square law detectors. A typical setup involving the bolometer-voltmeter method of measuring attenuation is illustrated in figure 12-189.

12-568. The modulator unit square wave modulates the klystron source. In order to prevent frequency modulation, it must be stable and the generated waveform must have a sharp rise and fall. 100 percent modulation is required. Tuner No. 1 matches the equipment to the oscillator so that maximum stable power is transferred. Buffer No. 1 sets the power level so that only the maximum permissible power enters the bolometer. It also maintains a constant load as seen by either the generator or the equipment. Its attenuation is usually around 10 db. When used, the frequency meter serves to set the oscillator frequency at the desired value. The auxiliary line, which you can insert at plane a in place of the output end of the equipment, is employed to match the non-oscillating generator after it has been adjusted for frequency and maximum power transfer. The generator acts as a load, and Tuner No. 2 is varied until the slotted-section indicates a swr of less than 1.02. Buffer No. 2 preserves the generator match when you vary the frequency meter after you adjust Tuner No. 2. Tuner No. 3 matches the load end of the equipment as seen by the slotted section at plane a. The measurement of attenuation proceeds in the following way after the equipment has been matched for a desired frequency.

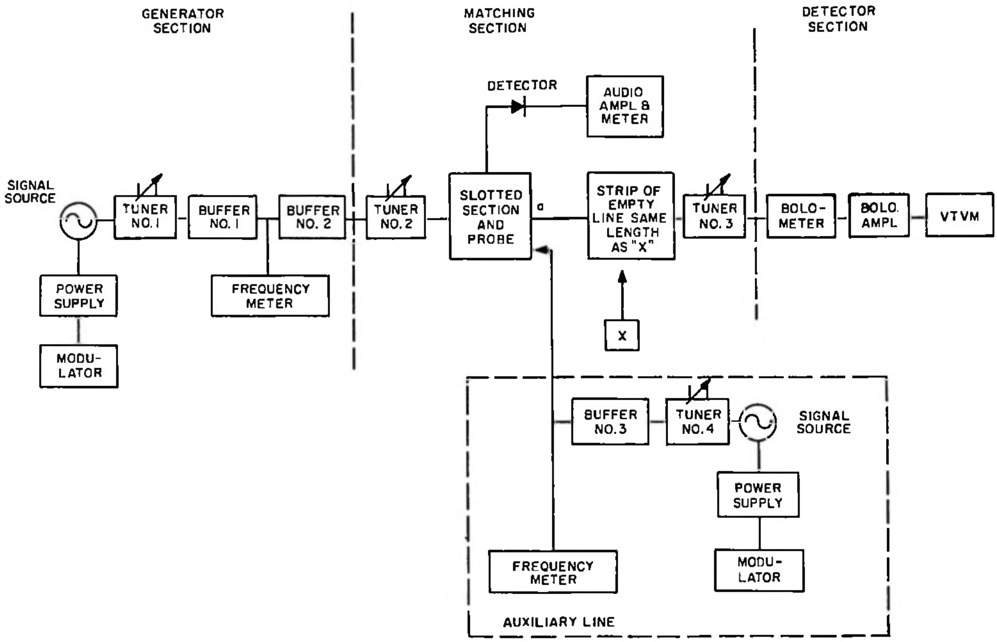


Figure 12-189. Line Setup for Measuring Attenuation by Bolometer-Voltmeter Method

12-569. Replace the unknown attenuator with an empty casing, or line length, and adjust the indication on the output meter at a convenient level (N). Without disturbing the complete equipment, insert the unknown in place of the empty section. Read this output as N_2. Repeat the first part to check level N_1. If it has remained stable, then the insertion loss corresponds to 1/2 $(N_1 - N_2)$ if you calibrated the meter for linear detection and the measurement is performed on one scale range.

12-570. In measuring the attenuation of a variable attenuator after you have determined the insertion loss corresponding to its minimum setting by the procedure described above, insert attenuation in steps and read the corresponding changes in output level until the whole unit is calibrated. Check level N_1 to make sure that it has not varied before accepting the results. Use the average of several stable runs.

12-571. For high-power measurement of attenuation, you must replace the bolometer detector with a water calorimeter. It measures the temperature rise in the liquid flow through a heating element. Calorimetric measurements are, by definition, the primary standard of power and usually serve as a reference standard in standardizing laboratories.

12-572. The measuring of insertion or attenuation by substitution methods relies upon comparing the unknown attenuation with a reference standard. The general procedure is to maintain the output power level constant as you introduce an unknown attenuation by removing a similar amount from the standard attenuator. Therefore,

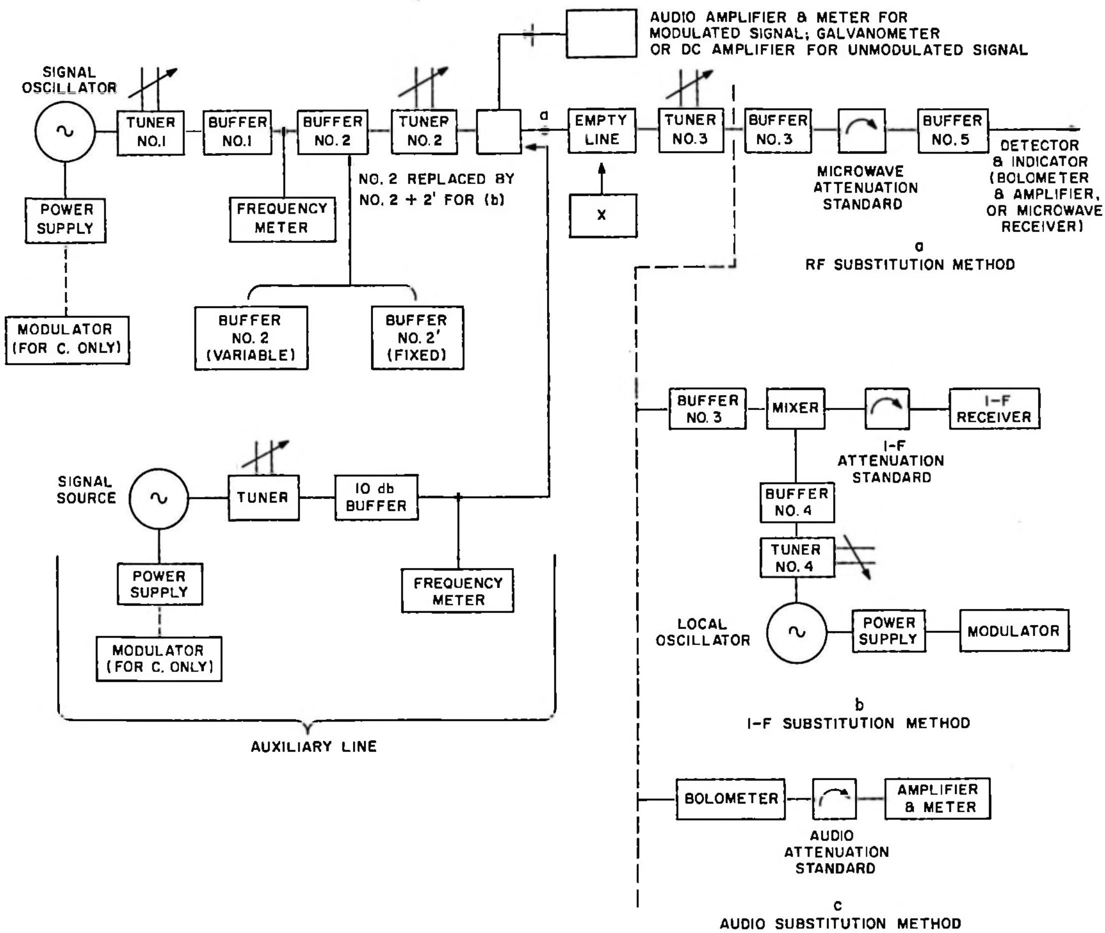


Figure 12-190. Substitution Methods for Measuring Attenuation

the decrease in the setting of the reference standard is equivalent to the unknown attenuation you inserted. Various arrangements are possible, depending upon the type of standard employed, as shown in figure 12-190.

12-573. The rf substitution method employs a calibrated microwave dissipative attenuator. The calibration is usually in terms of a cutoff attenuator at specific frequencies. You can use this method to measure attenuation up to about 70 db.

12-574. The i-f substitution or heterodyne method uses a standard which has been calibrated at the intermediate frequency. The intermediate frequency is derived by converting the radio frequency by means of a linear converter. The accuracy is limited by the linearity of the frequency converter. However, you can realize attenuation measurements of approximately 50 db.

12-575. The audio substitution method makes use of a calibrated audio attenuator and a bolometer detector. It is essentially limited to 40 db.

12-576. SOURCES OF ERRORS. Probable sources of errors in performing attenuation measurements by the methods described are: power level variation, bolometer nonlinearity, changing bolometer resistance, bolometer mismatch, amplifier nonlinearity, meter movement and scale inaccuracies, reading of the meter, range switch inaccuracies, effect of detector noise, calibration of standards, resettability of the standard, and mixer nonlinearity. The maximum over-all accuracy of each method is around $\pm$ 0.10 db. In addition, mismatches introduced by the generator and load may yield a phasing error.

12-577. MICROWAVE ATTENUATION MEASUREMENT METHODS.

12-578. REFLECTION METHODS. Alternative means for determining microwave attenuation rely upon measurements involving the swr and reflection coefficient. These methods are most useful for measuring attenuation below 20 db. Various modifications can be introduced for very small attenuation values.

12-579. The general expression for the insertion loss of a component is given by:

$$A = 10 \log \frac{P_O}{P_2} \text{ db}$$

where:

P_O = delivered output power with the component out of the circuit

P_2 = delivered output power with the component inserted

These conditions were illustrated in figures 12-156 and 12-157. You can separate the insertion loss into two components by the introduction of P_1, as shown in figure 12-157. Since:

$$\frac{P_O}{P_2} = \frac{P_O}{P_1} \cdot \frac{P_1}{P_2}$$

therefore:

$$A = 10 \log \frac{P_O}{P_1} + 10 \log \frac{P_1}{P_2}$$

The first component represents attenuation by reflection, or:

$$A_{ref} = 10 \log \frac{P_O}{P_1}$$

The second component is caused by energy loss, or:

$$A_{dis} = 10 \log \frac{P_1}{P_2}$$

In measuring microwave attenuators, you must evaluate these components individually. In microwave equipments you must match the generator and load impedances to the characteristic impedance of the low-loss transmission line; that is, $Z_G = Z_L = R_O$. When you introduce the attenuator, a mismatch may occur at the input terminals, which results in standing waves. The power reflected will yield:

$$A_{ref} = 10 \log \frac{(S_o + 1)^2}{4S_o}$$

where S_o is the swr at the input terminals of the attenuator when you connect its output terminals to the matched load. You can determine the attenuation component (A_{dis}) by measuring the power ratio $\left(\frac{P_2}{P_1}\right)$ or the transmission efficiency (η), which can be evaluated from reflection coefficient measurements. The method involves a sliding short circuit. When the attenuator under test is reversed and terminated by a sliding-short, the reflection coefficient of its input terminals follows a circular locus as the sliding-short varies in position. It can be shown that the radius of this circle (r) is equivalent to the efficiency (η). Therefore:

$$A_{dis} = 10 \log \frac{1}{\eta} = 10 \log \frac{1}{r}$$

The total insertion loss of the attenuator becomes:

$$A_r = A_{ref} + A_{dis} = 10 \log \frac{(S_o + 1)^2}{4rS_o}$$

The measurement procedure involves a set-up similar to that shown in figure 12-189 with a few modifications. Replace the bolometer termination by a load. It is not necessary to modulate the source. Consequently, the swr amplifier can be replaced with a sensitive galvanometer. As described in paragraphs 12-568 and 12-569, match the load by means of Tuner No. 3 to as nearly a perfect match as possible with the unknown attenuator out of the equipment. Insert the attenuator in place of the strip of empty line, and measure S_o. Now, reverse the attenuator and terminate it with a sliding-short. For a selected position of the short circuit, covering a distance of a half-wavelength, measure the corresponding swr, S_c, and the position of its first minimum (ℓ_m), from the terminals of the attenuator nearest to the slotted section. The magnitude of the reflection coefficient (ρ) is computed for each case:

$$\left|\rho_c\right| = \frac{S_c - 1}{S_c + 1}$$

Its phase angle is found from the following relationship:

$$\theta_c = 2\beta\ell_m \pm \pi$$

This type of measurement has been used with excellent results up to 20 db. However, it is limited by the errors in measuring swr, and the position of the minima, as well as the losses associated with the sliding-short. The radius, r, may be determined with less accuracy in terms of the minimum and maximum values of the swr as the sliding-short is varied. If the circular locus of the reflection coefficient encloses the origin, then:

$$r = \frac{\rho_{max} + \rho_{min}}{2}$$

$$= \frac{S_{max} S_{min} - 1}{(S_{max} + 1)(S_{min} + 1)}$$

If the circle lies outside the origin:

$$r = \frac{\rho_{max} - \rho_{min}}{2}$$

$$= \frac{S_{max} - S_{min}}{(S_{max} + 1)(S_{min} + 1)}$$

12-580. The two-point method represents a rapid way for you to measure the characteristic impedance and attenuation constant of cables. The attenuator under consideration in the preceding paragraph is replaced by a cable which you terminate with a sliding short. The input impedance of the cable is related to the reflection coefficient. Solving for the input impedance:

$$Z_i = R_o \left(\frac{1 + \rho_c}{1 - \rho_c}\right)$$

The reciprocal of Z_i gives the input admittance, or:

$$Y_i = \frac{1}{R_o} \left(\frac{1 - \rho_c}{1 + \rho_c}\right)$$

Normalizing Y_i with respect to the characteristic admittance of the line yields:

$$Y_i = \left(\frac{1 - \rho_c}{1 + \rho_c}\right)$$

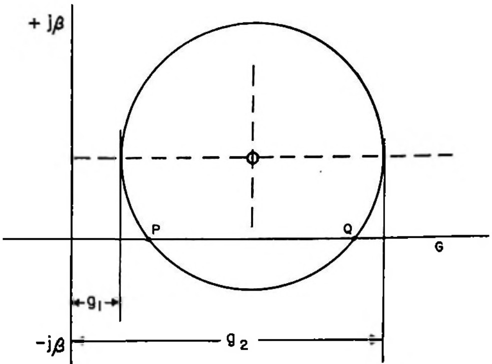


Figure 12-191. Admittance Circle in the Complex Plane

As you vary the position of the sliding-short, the normalized input admittance will describe a circle when plotted in the complex plane. This is shown in figure 12-191. The displacement of the center of the circle from the real axis is influenced by the non-uniformity of the cable under test and the mismatch between the measuring line and the cable. For small displacements, the minimum and maximum conductances (g_1 and g_2) correspond rather closely to the intercepts (P and Q) when the characteristic impedance of the cable is nearly equal to the characteristic impedance of the line.

12-581. For the condition that $g_1 - 1$ is smaller than g_2, the minimum pure conductance (intercept P) is proportional to the swr (S_1) when its minimum is $\lambda/4$ from the cable input. The maximum pure conductance (intercept Q) is proportional to the swr (S_2) when its maximum is $\lambda/4$ from the cable input. From these swr measurements, it can be demonstrated that the normalized conductance at the two extremes of the G axis are as follows: g_1 is approximately equal to $\frac{1}{S_1}$, and g_2 is approximately equal to S_2 for the assumption that g_1 is smaller than 1, which is smaller than g_2. However, this condition may not hold when the attenuation of the sample is high and its characteristic impedance does not approach that of the measuring line. When the characteristic impedance of the line is higher than that of the cable, g_1 larger than 1 may occur. In this case, g_1 is approximately equal to S_1 and g_2 is approximately equal to S_2. When the characteristic impedance of the line is lower than that of the cable, g_2 may be smaller than 1. In this case, g_1 is approximately equal to $\frac{1}{S_1}$, and g_2 is approximately equal to $\frac{1}{S_2}$.

As soon as you know the minimum and maximum conductances, the characteristic impedance of the cable (R_o') and the attenuation constant of the cable (α) can be computed as follows:

$$R_o' = \frac{R_o}{\sqrt{g_1 g_2}} \quad \text{ohms}$$

and

$$\alpha = \frac{1}{\ell} \tanh^{-1}\sqrt{\frac{g_1}{g_2}} \quad \text{nepers}$$

where:

ℓ = length of the cable sample

R_o = characteristic impedance of the line

Therefore, g_1 is approximately equal to S_1 or $\frac{1}{S_1}$ and g_2 approximately equals $\frac{1}{S_2}$ or S_2. The measuring procedure is as follows:

a. Vary the sliding-short so that the voltage minimum is $\frac{\lambda}{4}$ from the input terminals of the cable, and measure swr S_1.

b. Vary the sliding-short so that the voltage maximum is $\frac{\lambda}{4}$ from the input

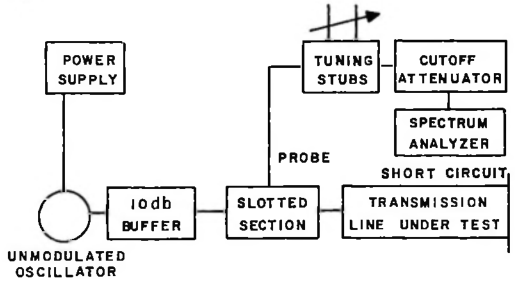


Figure 12-192. Reflection Method for Measuring Attenuation, Block Diagram

terminals of the cable, and measure swr S_2.

This rapid method is fairly accurate for detecting flaws in cables. However, its accuracy is not so great as the method discussed in paragraph 12-579.

12-582. RESONANCE METHODS. You can perform the measurement of very small attenuation constants of transmission lines by means of resonance methods based upon the Q-factor defined by:

$$\frac{1}{Q} = \frac{\Delta f}{f}$$

where f is the resonance frequency for maximum power absorption, and Δf is the frequency increment below and above the frequency (f) at which the admittance magnitude reduces to one-half its peak value.

12-583. The equipment arrangement of a practical microwave setup is illustrated in figure 12-192. To avoid reflections, the slotted section and the transmission line under test are of the same dimensions. You can adjust the maximum power in the probe by the cutoff attenuator to a convenient magnitude. Keeping the frequency constant, vary the distance through which the probe moves from resonance. Move the probe in either direction from resonance in order to obtain the total distance between half-power points. This total distance corresponds to the total attenuation. The applicable formula of this distance is $\delta = \pm \alpha \ell$, where α is the attenuation constant, and ℓ is the line length to the sliding-short. This method may also be extended to include waveguide lines. Then the attenuation constant (α) can be written as

$$\alpha = \frac{2\pi}{\lambda g} \cdot \frac{\delta}{\ell} \text{ nepers per meter}$$

where:

λ_g = waveguide wavelength (meters)

ℓ = resonant length of cavity (same units as δ)

δ = change in cavity length to decrease amplitude of oscillation to 0.707 of the maximum amplitude (same units as ℓ)

12-584. One resonance method employed to measure the attenuation of waveguides is accurate to within 3 percent. A short length of the waveguide sample having an axial length of $0.5m\lambda_g$ (where m is an integer and λ_g is the waveguide wavelength) is joined to two adapter waveguide sections, as illustrated in figure 12-193. The terminating adapter is short-circuited by an end plate. The input adapter is provided with an iris. This combined assembly forms a resonator. When the waveguide sample is removed, the adapter sections are connected together, forming a second resonator whose resonant frequency is that of the foregoing unit. The contact losses of the connecting interfaces are negligibly small. Actually, you can show that:

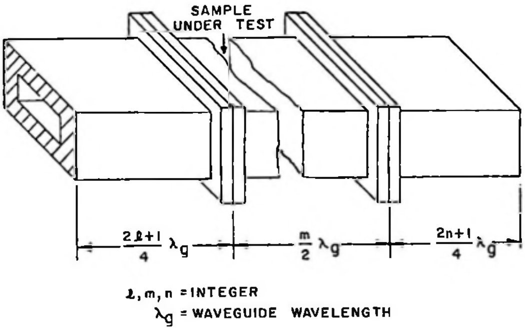


Figure 12-193. Resonator Assembly Under Test

$$(m + \ell + n + 1)\Delta f_2 \cdot (\ell + n + 1)\Delta f_1 =$$

$$1.92\sqrt{\lambda_o}\ mAf\ \frac{4}{a}\left[\left(\frac{\lambda_o}{\lambda_c}\right)^2 + \frac{a}{2b}\right] 10^{-5}$$

where:

a and b = long and short internal dimensions of the waveguide, respectively

f = resonance frequency

Δf_2 and Δf_1 = frequency half-power bandwidths of the resonance curves with and without the waveguide sample, respectively

λ_o = free-space wavelength

λ_c = cutoff wavelength of the te_{10} mode

A = relative attenuation normalized to the theoretical value of pure copper

By substituting the corresponding numerical values, a = 2.85 cm, b = 1.26 cm, λ_o = 3.32 cm, λ_c = 5.70, f_o = 9.030 mc, $\ell + n + 1 = 10$, and m = 15, this last equation reduces to:

$$2.5\Delta f_2 - \Delta f_1 = 0.975A$$

Thus, if you measure the frequency half-power bandwidths, you will obtain the relative attenuation (A) of the waveguide sample. A block diagram of the measurement setup is shown in figure 12-194. The output of the resonator is displayed as a typical resonance curve on the oscilloscope. Its frequency half-power bandwidth can be precisely measured by the frequency marker which is produced from the differentiated output of the calibrated frequency cavity.

12-585. You can determine the attenuation of a sample of waveguide line if you measure the swr. The method outlined below is particularly useful when the swr is high, that is, not less than 20. It consists of examining the standing-wave pattern in the vicinity of a voltage minimum. Insert the unknown and connect the sliding-short to its output terminals. Accurately set the probe of the standing-wave indicator to give minimum deflection on the detector. This reading is noted as P_m. Then move the probe along the guide a distance (x) in each direction to give two further readings corresponding to $2P_m$. The standing-wave pattern in the vicinity of its minimum in a lossless transmission line may be expressed, in terms of the voltage amplitude squared, as:

$$V^2 = V^2_{min}(\cos^2\beta x + S^2 + \sin^2\beta x)$$

where:

V^2_{min} is approximately equal to P_m

S = swr

$$\beta = \frac{2\pi x}{\lambda g}$$

λ_g = waveguide wavelength

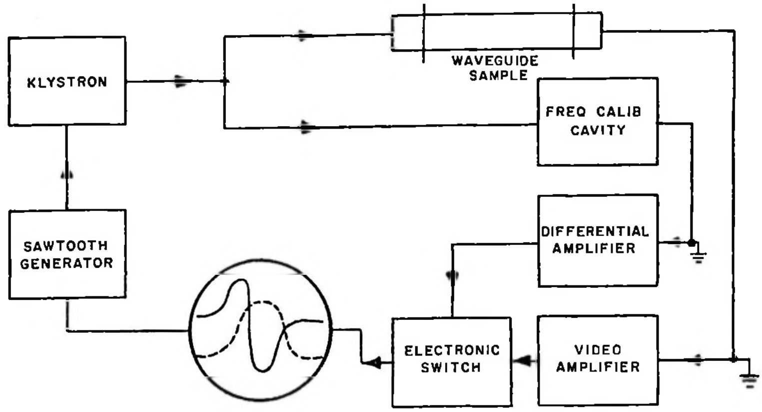


Figure 12-194. Resonance Method for Measuring Attenuation, Block Diagram

From the measurements performed:

$$\frac{V^2}{V^2_{min}} = \frac{P}{P_m} = 2$$

Thus:

$$S^2 = 1 + \frac{1}{\sin^2\left(\frac{2\pi x}{\lambda g}\right)}$$

If S is large, as it is when fairly short lengths of waveguide are being measured, then S is approximately equal to $\frac{\lambda g}{2\pi x}$. In the guide near the probe, you may express the voltage amplitude in terms of V_i, the voltage amplitude of the incident wave, and V_r, the voltage amplitude of the reflected wave. From these voltage components, the swr may be written as

$$S = \frac{(V_i + V_r)}{(V_i - V_r)}$$

The attenuation due to one-way transmission through the test sample is:

$$A = 10 \log\left(\frac{V_i}{V_r}\right) \text{ db}$$

Hence:

$$A = 10 \log\left(\frac{S+1}{S-1}\right) \text{ db}$$

The loss in the sliding-short can be evaluated separately by placing it directly on the end of the slotted section. This value can then be subtracted from A to determine the loss of the sample.

SECTION X

TIME INTERVAL MEASUREMENT

12-586. TECHNIQUES.

12-587. GENERAL.

12-588. The techniques of time measurement can be broadly categorized into two groups. The first group comprises those measurement methods which depend upon the comparison of the unknown time interval with a standard of frequency or time. For electronic applications, sinusoidal generators of a precisely known frequency are generally used as standards. The generator signals may be used to drive electromechanical clocks, as marker signals on cathode-ray-tube traces, for counting purposes, etc. The second group of measurement techniques depends upon the measurement of some physical quantity (for example, distance or voltage) whose time related value can be determined from a fundamental physical law. Thus, for example, the time it takes a freely falling body to move through a given distance can be directly computed from the laws of mechanics. Or the time it takes the voltage across a discharging capacitor to reach a given fraction of its original value can be determined from electrodynamic relations.

12-589. In general, the first group is more precise, primarily because in electronic measurements frequency can be determined with greater accuracy than any other physical quantity. In general, measurement techniques of the second type do not allow the measurement of the time varying quantity with high accuracy; also, the critical equipment parameters are unstable. Nevertheless, as will be seen below, these methods have considerable usefulness, especially when simplicity is desired and a lesser degree of accuracy can be tolerated, or when measuring extremely short time intervals for which comparison techniques are not suitable.

12-590. There are many subsidiary groupings in which time measurement techniques may be categorized, including long or short interval, digital or analog, and others. In general, these groups will overlap; thus, a frequency standard may be used in a digital device such as a counter, or in an analog instrument such as a synchronous motor-driven clock. Actually, many time measuring instruments are hybrids, employing a combination of techniques, deriving maximum benefit from each. For the sake of continuity, in the discussion that follows the methods are divided in accordance with the general concept of the instruments. Their relations to the several groupings will be evident.

12-591. CATHODE RAY TUBE METHODS.

12-592. As in many fields of electrical measurement, the oscilloscope plays a prominent role in time measuring devices, its application to radar facilities being one of the more outstanding examples. To illustrate, figure 12-195 represents a Type A scope presentation used in radar facilities.

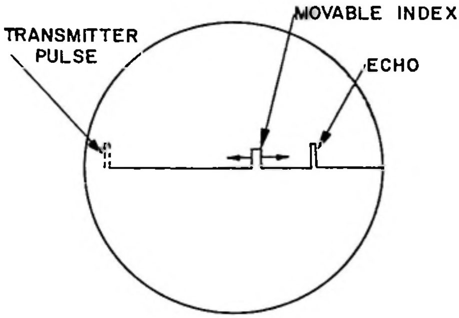


Figure 12-195. Type A Scope Presentation with Movable Pedestal Index

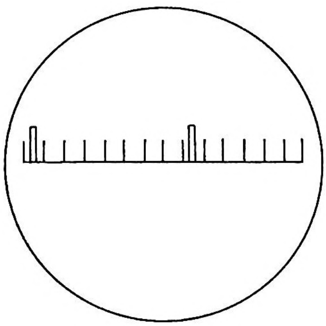


Figure 12-196. Cathode-Ray Tube Display with Standard Time Markers

The transmitter pulse, shown dotted at the left, starts the sweep and time delay generator. You adjust the moving pedestal index to coincide with the echo signal, and you determine the time delay, or range, from the calibrated setting of the delay generator control.

12-593. An inherent requirement of most radar facilities is the measurement of the time between a transmitted pulse and a delayed reply pulse, which may be an echo from a target or an answering signal from a ground beacon. From a knowledge of the velocity of propagation of the signal, you can convert the time difference directly into distance. Many radar and navigational facilities depend upon precision sweep generators or continuously variable time delay generators for their operation. Usually, some form of RC timing circuit is used. Among the more familiar timing circuits are multivibrators, "bootstrap" circuits, phantastrons, sanatrons, and sanaphants. All of these depend, in some way, upon the time behavior of the voltage across a capacitor which is being charged by a battery through a resistor. The precision and stability obtained in these circuits depends primarily upon the extent to which the design makes this relation independent of the tube and circuit parameters.

12-594. STANDARD TIME MARKERS. Oscilloscopes can also be used for time measurements with time standards in a variety of arrangements. To illustrate, you can employ a standard to either simply calibrate or actually control the sweep duration. In either case, precise sweep linearity is required if accurate interpolation is desired. In one of the better arrangements, the standard time signals may be impressed upon the sweep in the form of amplitude or intensity markers, thereby providing a permanent scale calibration and eliminating the linearity problem. In all of these schemes, the signals corresponding to the beginning and end of the unknown interval are also applied as amplitude or intensity markers, and the time interval is determined by the distance between them or by counting the number of standard markers falling between the signals. This latter case is illustrated in figure 12-196, where a linear sweep is shown. You can synchronize the sweep, markers, and signal so that repetitive displays can be generated and direct visual measurements effected.

12-595. One of the design objectives of all measuring instruments is a readability consistent with the inherent accuracy of the device. When a time standard is used with an oscilloscope, it is often desirable to pro-

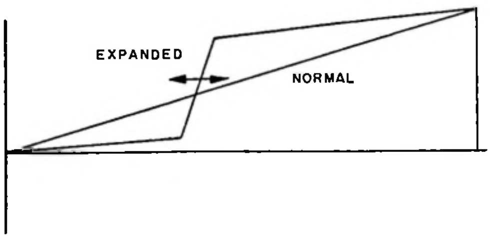


Figure 12-197. Comparison of Waveforms of Normal and Moving Expanded Sweeps

vide some form of expanded scale beyond that available from a simple linear sweep. In some instruments this is achieved with the combination of a linear sweep and a moving sweep magnifier. The waveforms which provide this type of operation are shown in figure 12-197.

12-596. SPIRAL SWEEP. Another method that has been found favorable for oscilloscope time measurements involves the use of a spiral sweep. This has the advantage of providing a linear scale whose useful length, for a typical 5-inch cathode-ray tube, is on the order of 50 times the tube diameter. A block diagram of a typical arrangement is shown in figure 12-198. The start and stop signals are fed to a gate multivibrator to initiate and terminate, respectively, a negative-going gate. This is fed to a sawtooth generator, which starts and ends in synchronism with the gate. The sawtooth is fed to a modulator, along with the output of a crystal oscillator. The resultant signal is a linearly decreasing sine wave which, after passing through the proper phase-shifting network, is fed to the deflecting plates of the cathode-ray tube, thereby producing a spiral sweep. To provide calibrated subdivisions of the sweep, the output frequency of the oscillator can be multiplied (for example, by a factor of 10), and applied to the grid to produce intensity markers. The unknown time interval is then measured by computing the total number of revolutions of the sweep.

12-597. In practice, the accuracy obtainable in this sweep, as in all cathode-ray-tube displays, is limited primarily by the size of the tube, that is, the scale length, and not by the standard. In one case, an absolute accuracy of $\pm$ 1 microsecond was achieved with a maximum duration of 2000 microseconds, using a 5-inch cathode-ray tube and an angular velocity of 100 microseconds per revolution. In another case, employing the same size cathode-ray tube but an angular velocity of only 2 microseconds per revolution, an absolute accuracy of $\pm$ 0.05 microsecond was obtained. A maximum duration

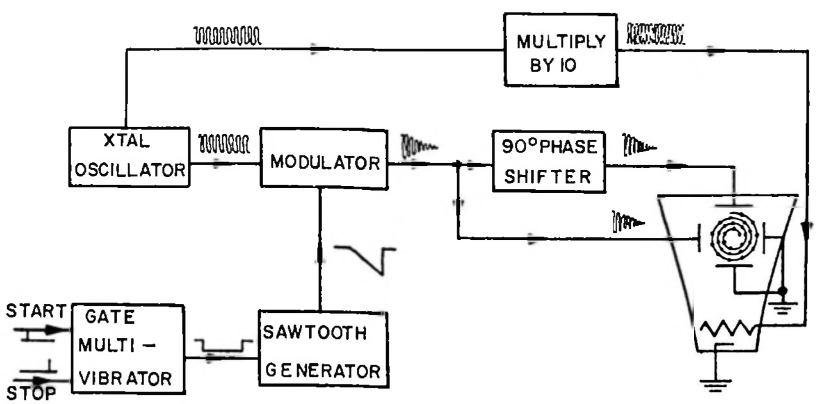


Figure 12-198. Time-Measuring Equipment Employing a Cathode-Ray Tube Display with Spiral Sweep, Block Diagram

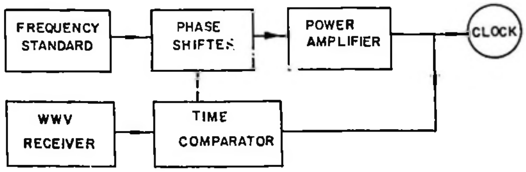


Figure 12-199. An Accurate Time Service Setup for Synchronous Clocks, Block Diagram

of 100 microseconds was available in this case.

12-598. RASTER DISPLAY. Another method for obtaining increased scale length for a cathode-ray tube uses a raster display similar to that employed in television equipments. The horizontal sweeps can be synchronized with a crystal oscillator, and calibration markers applied, as in the spiral sweep described above. It is important that the line flyback time be sufficiently short so that no significant time is "lost" as compared with the desired accuracy. An instrument of this type displaying a raster of 10 lines, each of 10,000 microseconds duration, has an accuracy of ± 15 microseconds.

12-599. MOTOR DRIVEN CLOCK APPLICATION.

12-600. CONTINUOUS TIME INDICATOR. A clock employing a synchronous motor to drive a mechanical gear train is limited in accuracy only by the precision and stability of the driving source frequency, and the backlash and accuracy of the gears themselves. As a consequence of their inherent precision, clocks are generally used in conjunction with frequency and time standards whenever a continuous indication of elapsed time is desired.

12-601. One of the various arrangements is shown in block diagram form in figure 12-199. A frequency standard is employed whose output is at the operating frequency of the clock motors. You apply this output, through a continuous phase shifter, to a power amplifier and then to the clocks. The output is also compared with the standard second intervals transmitted by station WWV. One method for doing this employs a clock with one revolution per second. The position of the clock hand is compared, either visually or aurally, with the arrival of the standard ticks. If there is an error, the phase shifter is rotated in the proper direction to correct this, one revolution of the phase shifter producing a 360-degree change in phase in the amplifier output and a correction of 1/f in time. With this method, it has been possible to achieve a stability on the order of ± 0.1 second in a 24-hour period, or about 1 part in 1,000,000.

12-602. SINGLE TIME INTERVALS. In addition to their primary use as continuous time indicators, synchronous clocks have also been applied to the measurement of single time intervals. The basic requirement of such an arrangement is to prevent the clock hands from rotating except during the time interval to be measured. You can accomplish this by using a special clock in which the hands are coupled to a continuously energized synchronous motor through a friction clutch. An electromagnetic brake prevents the hand drive shift from rotating, forcing the clutch to clip, until a signal corresponding to the start of the unknown interval releases it. The stop signal subsequently re-energizes the brake. Because of the inherent time difference (on the order of 10 milliseconds) between the release and reapplication time of the brake, the clock is not directly usable, even with correction, for intervals much less than about 30 milliseconds.

12-603. By employing a time multiplying circuit, however, this lower limit can be reduced considerably. The arrangement described below provides an improvement of 1000 to 1. Figure 12-200 is a simplified

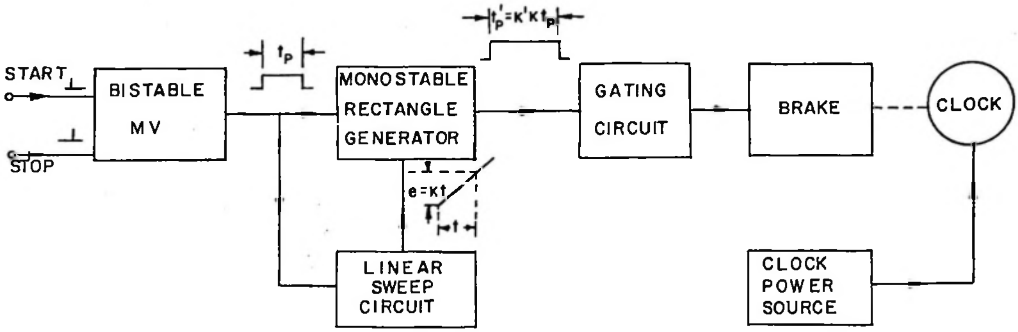


Figure 12-200. Single Time Interval Measuring Arrangement Using a Synchronous Clock, Block Diagram

block diagram of the complete arrangement. You must apply the start and stop signals to a bistable multivibrator which generates a pulse of duration (t_p) equal to the unknown interval. You apply this pulse to a linear sweep generator and to a monostable rectangle generator. The sweep generator starts with the beginning of the pulse, while the monostable generator is triggered by the trailing edge. The sweep circuit output is fed to the monostable circuit, which produces a gating signal whose duration (t_p') is a direct function of the sweep generator output voltage at the time t_p. Now apply the gating signal to a gating circuit to release the clock brake during the interval t_p'. The duration of the gating signal is determined by the slope of the sweep signal and the constants of the monostable circuit. By varying either of these, you can obtain different ranges for the instrument.

12-604. COUNTER APPLICATION.

12-605. The recent introduction of digital counting techniques has led to the development of time and frequency measuring apparatus of great flexibility and accuracy. These instruments offer numerous advantages including operation over a wide range, direct decimal presentation of the result, and a display time which can be varied from zero to infinity.

12-606. Numerous circuit arrangements have been described for such devices, differing principally in such things as scaling circuitry, read-out arrangements, and operating range. However, all use the same basic method of counting the number of cycles of a precision oscillator which occur during the unknown time interval. This is illustrated in figure 12-201. The output of the standard oscillator is fed to a gate which is opened by the application of a start signal and closed by the application of a stop signal. During the time the gate is opened, the oscillator signals are fed to the scalers, where they are counted. At the end of the interval, the number of cycles is read off the scalers. If the oscillator frequency is

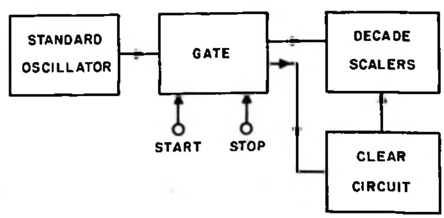


Figure 12-201. Basic Counter Equipment Arrangement for Time Measurements, Block Diagram

some power of 10 and decade scalers are used, then the number of counts gives the time interval directly in decimal values; for example, if 1 megacycle is used, the result is given directly in microseconds. The gate output is also fed to the clear circuit, which, operating on the trailing edge of the gate waveform, returns the scalers to zero after some predetermined delay, which is frequently made adjustable to suit the operator's needs.

12-607. SCALER SPEED. The operating range of counter instruments is limited at the low end primarily by the counting speed of the scalers. At the present time, approximately 10 mc appears to be the upper limit for reliable results, limiting the smallest resolvable time element to 0.1 microsecond. The upper limit is determined by the product of the period of the standard oscillator and the number of scaling units employed.

12-608. COUNTING PERIOD. Accurate results have been achieved with counter instruments. However, they are limited by the accuracy of the standard oscillator and by the inherent uncertainty of a $\pm$ 1 counting period, which is present in all unsynchronized counters as a result of the random phasing of the standard oscillator and the time interval to be measured. For example, if a standard frequency of 1 megacycle is used, and an unknown interval of 100.1 microseconds is to be measured, then either 100 or 101 peaks of the 1-megacycle signal might pass through the gate to the scalers, producing a reading of either 100 or 101 microseconds, respectively. For maximum accuracy, it is apparent that the highest possible standard frequency should be used consistent with the total number of counts that can be displayed. Over-all accuracies of $\pm$ 0.0002 percent $\pm$ 1 counting period have been realized in commercial equipment.

12-609. There is an interesting instrument which avoids the ambiguity of the $\pm$ 1 count by combining the counting technique with an interpolating cathode-ray-tube display. The maximum time interval it measures is 1000 microseconds, and its absolute accuracy is $\pm$ 0.01 microsecond, considerably better than can be obtained with straight-forward scaling methods. A block diagram of the device is shown in figure 12-202. The start and stop signals are fed to the gate to permit the scalers to register the number of standard oscillator pulses which appear during the unknown time interval. The total number of complete periods of the standard oscillator which occur during the interval is always one less than the number of counts, regardless of the relative phase of the interval with respect to the standard oscillator signal. The remainder of the time interval is measured by the cathode-ray-tube display. You can accomplish this by feeding the standard oscillator signal directly through a 90-degree phase shifter to the deflecting plates of the oscilloscope to produce a circular sweep. The start and stop signals are mixed and fed to the grid of the tube, producing intensity markers. You should make the start signal larger for the purpose of identification. The sum of the time intervals between the start signal and the first count, and between the last count and the stop pulse, is indicated by the angular distance between the two markers, reading from start to stop in the direction of sweep rotation.

12-610. In the instrument discussed, if a standard oscillator frequency of 100 kc were used, it would produce a 10-microsecond sweep. Angular resolution of 1 part in 1000 would be required to obtain an accuracy of $\pm$ 0.01 microsecond. The upper limit of 1000 microseconds is due to the use of only two scalers. This can be extended by a factor of 10 for each additional scaler used. The use of a standard frequency of 1 megacycle would increase the accuracy by the same factor, provided that the operating

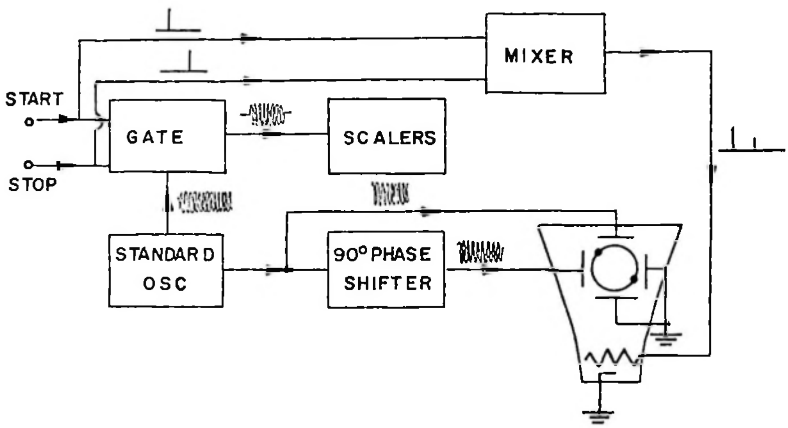


Figure 12-202. Time Measuring Equipment Arrangement Employing Counters with an Interpolating Cathode-Ray Tube Display, Block Diagram

time of the gate and the rise time of the start and stop signal were made correspondingly smaller. As in all counting equipment arrangements, it is necessary that these be, at most, on the same order of magnitude as the smallest desired resolvable time.

12-611. VOLTMETER APPLICATION.

12-612. In paragraphs 12-592 through 12-598, it was explained that time measurements can be performed by applying a time varying voltage to a cathode-ray tube as a sweep signal. These signals may take a variety of forms, the most common in time measurements being linear and spiral. Where you do not use markers, the cathode-ray tube can be considered as a special type of voltmeter, converting the varying voltage into a visual time base against which the unknown interval is compared. It is necessary, of course, for you to know the time-voltage function to make the measurement.

12-613. DIRECT VOLTMETER INDICATION. In a somewhat similar manner, you can measure the time intervals with conventional voltmeters. Consider the arrangement

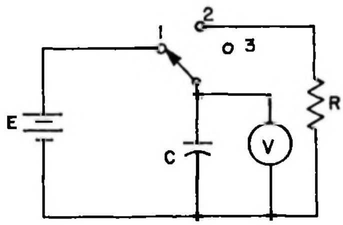


Figure 12-203. Simple RC Time Measuring Method Employing Direct Voltmeter Indication

shown in figure 12-203. Before starting the measurement, throw the switch to position 1 and allow capacitor C to charge to the battery potential, E. At the start of the unknown interval, throw the switch to position 2 and permit the capacitor to discharge through resistor R. Finally, at the end of the interval, throw the switch to position 3, disconnecting the capacitor. The voltage across the capacitor, after being switched to position 2, is given by:

$$e = E\,\epsilon^{-\frac{t}{RC}}$$

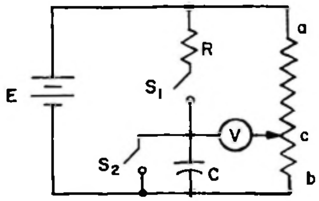


Figure 12-204. RC Measuring Method with Potentiometer Comparison

If you measured the voltages across the capacitor in positions 1 and 3, then the time interval the switch remained in position 2 would be given by:

$$T = -RC \ln \frac{e_3}{E}$$

It is assumed, of course, that no charge is lost by the capacitor through the voltmeter. In actual practice, it would be necessary to employ high-impedance devices, such as vacuum-tube or electrostatic voltmeters, to insure against such loss.

12-614. POTENTIOMETER COMPARISON. This concept has been applied in somewhat modified form to several time-measuring instruments. In one of these a basic difference, as shown in figure 12-204, is that instead of measuring the actual voltage across the capacitor, you compare it with the setting of the arm of a potentiometer which you connect across the same battery as the RC circuit. Before the start of the measurement, close switches S1 and S2. Start the measurement by opening S1, and end it by opening S2. Next, adjust the potentiometer arm until the voltmeter reads zero. Since the capacitor is charged, rather than discharged, during the interval, a slightly different expression from that given in paragraph 12-613 is required to determine the unknown time. Thus,

$$T = -RC \ln \left(1 - \frac{e}{E}\right).$$

If a linear potentiometer is used, this can be rewritten directly in terms of the potentiometer setting as:

$$T = -RC \ln \left(1 - \frac{cb}{ab}\right) = -RC \ln \frac{ac}{ab}$$

The major sources of error in this measuring method arise from the errors and instabilities in the values of R and C, the leakage of a charge from the capacitor, and the closing time of the switches. For time intervals between 200 and 1750 milliseconds, accuracies of better than ± 0.6 percent can be achieved.

12-615. To avoid the logarithmic scale which results in the methods just described, you can employ constant-current sources to charge the capacitor, thereby deriving linear scales. The voltage across a capacitor is related to its charge by the expression:

$$V = \frac{Q}{C}$$

If the charge has been accumulated at a uniform rate, that is, I is constant, then:

$$V = \frac{IT}{C}$$

where:

I = constant current

T = time

12-616. ULTRA-SHORT TIME INTERVAL MEASUREMENTS.

12-617. In the study of a number of physical phenomena, such as spark gap breakdown, nuclear particle velocity, and phosphorescent decay time, the need arises for the

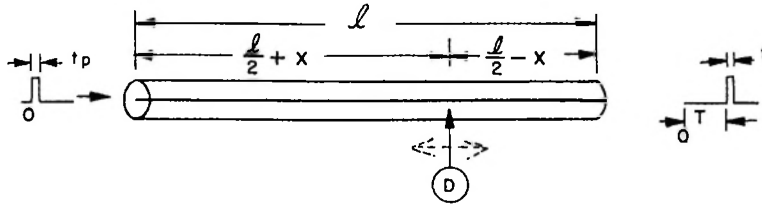


Figure 12-205. Transmission Line Method of Measuring Short Time Intervals with a Traveling Probe

measurement of extremely short time intervals, on the order of 10^{-8} to 10^{-10} seconds. The time-measuring procedures discussed thus far are not particularly suitable for intervals much less than 10^{-7} seconds in duration. In some cases, this limit could be and actually has been extended, but only at the expense of a considerable increase in equipment complexity. To make measurements of these small time intervals, you may find it convenient to use the propagation velocity of electromagnetic waves as the standard. In a typical arrangement, the two pulses whose time difference is to be determined can be applied to the opposite ends of a transmission line, as shown in figure 12-205. If the pulses differ in their starting times by T, the pulses will superpose at some distance (x) from the center of the line. The first pulse will travel a length:

$$\frac{\ell}{2} + x = ct$$

where c is the velocity of propagation, ℓ is the length of the line, and t is the time. Similarly, the second pulse will travel a distance:

$$\frac{\ell}{2} - x = c(t - T)$$

Combining these two equations and solving for x, you will obtain:

$$x = \frac{cT}{2}$$

If you know the velocity, you can determine the time interval directly by locating the point of superposition. The equation for x specifies the point at which the two signals superpose for the maximum time. Progressively shorter intervals of superposition exist on each side of this point until, assuming equal pulse widths (t_p), at a distance greater than $\pm ct_p/2$ from the maximum, the two signals will be seen separately. For maximum accuracy, the detector must be capable of distinguishing the maximum. In the figure, a single traveling probe and detector are indicated. Alternatively, a group of fixed detectors can be used. With such an arrangement, measurements as low as 10^{-10} seconds can be realized with accuracies on the order of 3×10^{-11} seconds.

12-618. In another arrangement based on similar principles, you connect the ends of a charged transmission line to terminating impedances by fast, low-impedance switches. From the waveforms developed across the loads, as observed by a high-speed oscilloscope, you can determine the time difference of closure. Intervals of up to 3 microseconds with an accuracy of $\pm$ 0.1 microsecond can be obtained.

12-619. In those fields where the delay of a high-voltage-producing phenomenon, such as spark gap breakdown, is to be measured, the Kerr cell has been quite useful. The Kerr cell is essentially an electro-optical shutter. It consists of a transparent container with metallic end walls containing a special liquid. Under normal conditions, when you place it between crossed Nicol prisms, no light is transmitted. However, when you apply voltage to the end plates, the liquid becomes double refracting, converting the incident-plane polarized light into elliptically polarized light. As a result, some light is transmitted. This change in behavior occurs almost instantaneously with the application of voltage.

12-620. In spark gap delay measurements, you observe the gap through the Kerr cell. The pulse applied to the gap is also applied to the cell through a transmission line, and the length of the line adjusted until the breakdown can be seen through the cell. You can then determine the delay from the line length and propagation velocity.

SECTION XI

EXTERNAL PERFORMANCE FACTORS

12-621. GENERAL.

12-622. DEFINITION.

12-623. External performance factors consist of all external natural and man-made disturbances which interrupt or interfere with the electrical or electronic properties of operation, maintenance, or testing, and which cause improper operation or indication, or diminished equipment performance. These external factors are normally categorized as natural, mutual, and man-made.

12-624. PERFORMANCE FACTORS.

12-625. NATURAL. This type of performance factor can be the result of any one or a combination of the following phenomena:

a. Natural electromagnetic disturbances, such as lightning, aurora, or solar activity.

b. Atmospheric interference from electrical storms.

c. Seasonal interference represented by electrical changes associated with the different seasons (spring, summer, autumn, and winter).

d. Climate interference, which is the difference in electrical disturbance between areas that have different climates, such as Africa versus the area surrounding the North or South Pole.

e. Celestial interference, which results in a continuous noise or electrical disturbance and apparently emanates from other galaxies such as the Milky Way or from remote cosmic stars.

f. Precipitation interference, which is the electrical disturbance created by rain, sleet, sea spray, smoke, dust, or sand.

g. Fading disturbances, which originate in the medium through which electrical waves are propagated. These disturbances are associated with a signal passing through the troposphere, atmosphere, stratosphere, ionosphere, etc.

h. Chemical disturbances, caused by such transmission mediums as the salt water of an ocean.

i. Geographical interference, which is the result of locating a transmission or reception point in the vicinity of a mountain containing an excess of metallic content that creates an interfering magnetic field.

12-626. MUTUAL. This type of performance factor exists when one or more electrical or electronic services causes abnormal operation or diminished performance in one or more other equipments, facilities, sites, or systems. This interference may be in the form of cross-talk, improper indication, diminished performance, or spurious response. The following factors can cause mutual interference:

a. Spacing interference where two similar types of transmitters, such as radio, sonar, radar, etc, are physically located close together, permitting the signal from one to override the signal of the other.

b. Frequency interference, which is caused by two transmitters operating on adjacent frequency channels; the fundamental and/or its harmonics may be close enough in frequency to cause a combination or mixture with the adjacent transmitted frequency; the fundamentals and/or their harmonics combine to produce an interfering harmonic.

c. Reflections, refractions, or bounced waves can mutually interfere or combine with their own origin at the point of reception.

d. Spacing interference where two dissimilar transmitters such as radio and sonar, radar and loran, radio and radar, etc, are physically located close together, permitting the signal of one to combine or mix electrically with the other.

12-627. MAN-MADE. This type of performance factor is produced by man, and consists of many forms of miscellaneous electrical, electronic, and mechanical devices that generate and transmit electrical interference. These interference-creating devices are in general use and produce radiations incidental to their normal functioning. The following list illustrates some of the types of man-made interference:

a. Internal or equipment noise.

b. Radiation interference intercepted by the antenna.

c. Conducted interference resulting from the direct physical connection between equipments.

d. Mutual coupling resulting from the placement of cables from unsuppressed electrical equipment near cables carrying currents for the prime equipment.

12-628. The following are some of the possible sources of man-made interference.

a. Arc lights

b. Battery-charging regulators

c. Calculating machines

d. Dial telephones

e. Electric bulbs

f. Electric buzzers

g. Electric generator flashing signs

h. Electric railway signals

i. Electric razors

j. Electric traffic signals

k. Electric typewriters

l. High-frequency industrial and medical equipment

m. High-tension transmission lines

n. Ignition equipment

o. Machine-driving belts

p. Magnetos

q. Neon lamps

r. Power line leakage

s. Rectifiers (dry-type or mercury-arc)

t. Relays

u. Switches

v. Teletypewriter equipment

w. Thermostats

x. Vibrators, etc

12-629. NATURAL INTERFERENCE.

12-630. ATMOSPHERIC.

12-631. THUNDERSTORMS. Atmospheric interference is caused by the numerous thunderstorms that occur over the surface of the earth. In ordinary communications equipment, this interference appears as noise, a constant background rumble with loud crashes occurring at irregular intervals. The noise may not be heard at all times, but it is always present in radio receivers and may be the source of an unidentifiable interference problem.

12-632. LIGHTNING. Thunderstorms can cause this type of interference because their rapidly moving air masses cause clouds to become electrostatically charged. Lightning, the resultant discharge of these clouds, produces electromagnetic energy waves which are scattered in all directions. These waves are received locally as overriding volume crashes. In addition, the waves may be transmitted to reception antennas located at distant points because the waves may be reflected or refracted from the ionospheric layer (a layer of the earth's upper atmosphere containing charged atoms and atomic particles) at such an angle as to be directed toward the distant receiving antennas.

12-633. CELESTIAL.

12-634. GENERAL. All types of celestial noise are generally encountered at frequencies above several hundred megacycles. These noises are described under several different classifications in the following paragraphs.

12-635. COSMIC. The continuous noise received from other galaxies is normally a minor source of irritation, except in those instances where highly sensitive test equipment it used in calibration procedures or where accurate measurements of minute voltages are attempted. This noise has, at times, resembled or acted like the transmission of intelligence signals. However, no proof or disproof has been offered as a satisfactory explanation of the peculiar sequence or arrangement of these signals. The common explanation for these noises is that they are due to magnetic storms resulting from the thermonuclear reactions continually occurring on the suns of these distant galaxies. These noises are not particularly directional because the transmitting galaxies completely surround our own galaxy.

12-636. GALAXIAL. The noises received from within our own galaxy are normally directional because they originate from definite traceable sources. Again, this is a type of minor interfering noise of a comparatively constant nature which affects only highly sensitive equipment.

12-637. STELLAR. Noise received from the stars within our own galaxy is highly directional and normally possesses a greater amplitude than cosmic or galactic interfering signals. Again, the normal amplitude of these signals is relatively low and affects only very sensitive equipments.

12-638. SOLAR. The noise received as a result of the thermonuclear reactions occurring on the sun is the greatest source of noise existing outside the sphere of our own planet. These noises do have sufficient amplitude to interfere with, or distort, some of our sensitive equipment. During periods of sun-spot activity, this highly directional source will vary as the earth rotates, and maximum interference will result when the receiving antenna is directed toward the sun.

12-639. In the arctic regions, severe magnetic storms resulting from sun spots have crippled electrical communications, including submarine cable. These interfering noises may last a short time or may continue for several days.

12-640. PRECIPITATION STATIC.

12-641. Noise caused by the gradual accumulation of an electrostatic charge on an antenna or other exposed conductive parts of electronic equipment is referred to as precipitation static. This charge builds up as the result of bombardment by charged precipitation particles. Since this noise occurs during bad weather, it can be distinguished from vacuum tube shot effect noise, which sounds approximately the same. Either wet or dry particles may result in the introduction of static noise.

12-642. WET PRECIPITATION. The classification of wet precipitation can be applied to rain, sleet, snow, sea spray, etc. Interference from precipitation static raises the average noise level of the receiving equipment, and the desired signal must override the amplitude of the static noise in order to be discernible. A greater amount of precipitation static will normally occur in a decreasing humidity period, such as during a rainstorm, than in an increasing humidity period, such as the period just before a rainstorm. In low-frequency radio receivers, the result of precipitation static is a loud rushing noise in the audio output.

12-643. DRY PRECIPITATION. Dry particles, such as smoke, dust, or sand, will produce a greater electrostatic charge than wet precipitation. Consequently, more static will occur during a sandstorm than during a rainstorm. The effects of precipitation, except for a greater noise voltage amplitude, is the same for dry particles as for wet particles.

12-644. CLIMATIC.

12-645. TORRID ZONE. The hot, high-humidity torrid zone is more prone to wet precipitation noises than any other area. The high humidity also creates corrosion conditions which result in additional noise and interference factors. However, the magnetic storms occurring on the sun do not represent a major source of interference in this zone.

12-646. ARCTIC ZONE. In the dry cold of the arctic with its sub-zero temperatures, communications is not primarily affected by high humidity or heat conditions. However, the disturbances created on the sun are more effective in this zone than in other zones. Severe magnetic storms sometimes completely disrupt communications in this zone.

12-647. CHEMICAL.

12-648. The salt water in an ocean will affect transmission or reception on board a submarine and will also affect the operation of transoceanic or other underwater cables. This saline solution is highly conductive for noise currents.

12-649. SEASONAL.

12-650. GENERAL. In any given area, the change from spring to summer or from any season to another will result in changes in both atmospheric noises and interference caused by solar radiation.

12-651. ATMOSPHERIC. The atmospheric seasonal changes are primarily due to changes in temperature and humidity. As the temperature or humidity gradually increases, the interfering noises will increase in direct proportion.

12-652. SOLAR RADIATION. A seasonal change is the result of changing the earth's face presented to the sun. This permits more or less of the directed solar radiation to reach a particular earth sector.

12-653. SPHERIC LAYERS. The ionospheric, stratospheric, and other layers in the upper atmosphere of the earth are composed of bands of electrons and ions. These bands change their relative distance from the earth for each seasonal change, that is, increase their height above the earth in winter and decrease their height in summer. More electrical waves will be refracted back to earth at greater distances from the transmitting medium during the winter than during the summer. As a consequence, desirable signals which have sufficient power will be transmitted greater distances during winter by bouncing from a layer of greater height. However, noise signals will normally not have sufficient power to reach this height and bounce back to earth. Therefore, during the winter season, the transmission and reception of desirable electrical signals will be possible over greater distances with less interference.

12-654. DAY VERSUS NIGHT. During the night, when the portion of the earth in which you are located is not facing the sun, you will not receive the same amplitude of solar noises. The electron bands in the upper atmosphere will lift and become a greater distance from earth. Therefore, again, transmission and reception will be vastly improved at greater distances because these layers have increased their height above the earth in the same fashion as for the winter months. Considering that solar radiation from the sun is not primarily directed toward the night side of earth, interfering noise from this source will diminish.

12-655. GEOGRAPHICAL.

12-656. GENERAL. Interference caused by geographical conditions is associated with the metallic or chemical content of the earth surrounding the location of transmission and reception of electrical signals.

12-657. METALLIC. In specific areas or points, the earth may contain a high metallic content which will effectively introduce a magnetic field. This field may mutually couple to the transmitting or receiving equipment via the desired signal, it may couple the desired signal to ground, or it may reduce the power of the signal and thereby shorten the transmission or reception distance. This interference will normally remain constant.

12-658. CHEMICAL. The interfering noise signals which accompany volcanic eruptions are normally effective only in a local region. The noise is caused by electrostatically charged particles resulting from the movement of gas and lava up through the earth's surface, by the thermal heat of the lava, and by the precipitation of dust or smoke particles.

12-659. FADING.

12-660. GENERAL. Signal fading is not a noise-producing type of interference. It is classed as interference only because it makes reception of a desired signal difficult and thus interferes with the efficiency and accuracy of electrical or electronic equipment.

12-661. MEDIUM. Fading or fluctuation of a desired signal may be due to disturbances in the medium through which the signal is propagated. The tropospheric, stratospheric, and ionospheric layers above the earth's surface constitute the medium. An incoming signal may lose or gain strength, both conditions are termed <u>fading</u>. However, it is only when the normal signal strength becomes less that reception becomes difficult. The signal strength may drop so low that the signal fades or disappears in the background noise. While the background noise may remain constant, the desired signal may raise above or fall below that level. The frequency of the fading cycle may be slow or rapid and may result in an instantaneous complete loss of the signal.

12-662. SIGNAL PLUS NOISE-TO-NOISE RATIO. For all practical purposes, the effect of fading is a function of the signal plus noise-to-noise ratio of the particular equipment involved. If the background noise level remains constant and the signal diminishes, the reception of a weak signal becomes difficult. The application of automatic gain control (agc) is useless, because increasing the gain to amplify the fading signal to the normal level will cause the noise level to be amplified an equal amount, and thus the result is the same poor signal plus noise-to-noise ratio. The peculiar atmospheric conditions that cause fading may last for hours or only a few minutes. Generally, fading is more prevalent during the summer months and during daylight hours. Fading is more serious at medium and high frequencies.

12-663. MAN-MADE INTERFERENCE.

12-664. INTERNAL NOISE.

12-665. Internal interference, known as set noise, is present to some extent in every electrical or electronic receiver. This noise arises from the natural action of electrons in transit within vacuum tubes and in other circuit components. If the receiving equipment is perfectly aligned and all of the internal components are in the best condition, the internal interference will still exist. However, if you perform proper maintenance with emphasis on accurate circuit alignment, internal interference will be minimized.

12-666. EXTERNAL NOISE.

12-667. GENERAL. For years electrical and electronic equipment has been subjected to the buzzes, roars, crackles, and crashes of man-made static under the impression that this type of noise was inevitable. However, in recent times, the methods of noise reduction or complete elimination have been advanced and refined to the point where man-made noise need not be tolerated. In fact, numerous municipalities have passed ordinances making it a misdemeanor, punishable by fine and/or imprisonment, to operate equipment and electrical applicances which interfere with electrical or electronic reception. Many cities and towns are employing specialists to locate and suppress man-made external interference.

12-668. Many kinds and types of equipment will produce undesirable radio frequency impulses which are transmitted and travel out through the air exactly the same as if they were deliberately prepared for broadcast. In addition, some equipments radiate back through the power line to other equipments unless the radiation is stopped by an impedance or absorbed by a reactance.

12-669. SOURCES. The majority of radio noises are produced by one of the following: (1) By disturbances in the power supply and transmission lines, caused by leaks to ground or to other conductors, such as a tree limb or any object that the lines may be touching. The object through which the leak occurs may be a leaky lightning arrestor, a cracked insulator, insulated tie wires, loose pieces of wire hanging on a line, or even defects in the generator. (2) By commutating devices, such as the commutators on motors and other apparatus. (3) By equipments which make and break the circuit, such as thermostatic contacts on flashing lights.

12-670. Radiated interference usually enters the affected equipment through the antenna lead-in cable, through nonsignal leads such as power or headphone leads, or through defective or poor shielding. This type of interference will pass through the equipment in the same manner as the desired signal. Conducted interference enters all stages of an equipment through the input power leads. The result is loss of equipment sensitivity, distortion of the desired signal, the presence of noise at the receiver output, or possibly a combination of two or more of these effects.

12-671. TRACING THE INTERFERENCE. The general location of the noise can be determined by a simple observation. If the noise is absent or greatly attenuated in the surrounding area outside the building or location containing the affected equipment, you may assume that the noise is originating in the building or the lines leading to it. On the other hand, if the noise generally pervades the entire surrounding area, you can assume that the noise is originating from a source external to the affected equipment building or location.

12-672. You can use a portable radio, powered by a set of self-contained batteries, to determine whether a noise is reaching the radio through the air or through the power line. In fact, there are few better noise-finding instruments than a sensitive portable radio. If the portable radio contains a loop antenna, the loop can be rotated to the direction of maximum noise. This places the loop broadside to the noise, but does not tell you whether the noise is in front of or behind the antenna. However, if you now walk in either direction that the broadside of the antenna points, the noise will increase as you approach the noise source or decrease as you move away from it. If this method leads you to a group of transformers, do not assume that the transformers are the noise source; the noise is probably coming into the transformer area through one or more lines. Occasionally, transformers do leak and create noise, but they are often blamed when some other equipment is the actual noise source. You should walk directly under, and follow, each of the lines away from the transformer area to see whether the noise decreases or increases. If the noise increases along one of the lines, follow this line to the noise source.

12-673. The noise may emanate from a group of equipments contained within a given building. Pulling (opening) the main switch for each equipment until the noise suddenly ceases will show you which major circuit contains the noise. If the noise is still present after all of the switches have been opened, the noise source is not in that building.

12-674. On every base there are numerous places where wires touch tree limbs or other objects and create noise. A single leaf can produce enough noise to destroy effective radio communication in a given area. If the leak is on the secondary side of the line, the noise will be apparent only to the equipment that wire serves. If the leak is on the primary side, it may affect all of the equipments in the area. Cracked insulators are difficult to locate, but a good portable radio will indicate this noise source when you pass under the insulator. Hardware noise, such as that produced by two metal brackets touching, can be removed by separating the parts that touch or by bonding them tightly together to form a common ground. You must always be aware of the fact that wires loosen during the day and during warm weather and that they tighten during the night and during cold weather. Therefore, a wire that touches nothing at one time may touch many things at another time.

12-675. An ordinary exposed wire leadin will act as a part of the antenna, and may contribute more noise than the antenna itself, because comparatively strong coupling may exist between the leadin and the equipment wiring which carries the noise. A shielded leadin plus a series filter should remove this type of noise.

12-676. Commutating devices, from those in the largest motor down to those in electric razors, are responsible for various degrees of interference. Large three-phase motors seldom produce enough interference to worry about. In fact, few large motors produce interference, especially those motors without brushes or commutators, unless the motor is defective. Single-phase motors of the shaded-pole and capacitor types are also noise-free when not defective. Brush-type motors, either shunt, compound, or series wound, produce noise, the violence of which varies inversely with their size and directly with their speed. Repulsion-start induction-run motors normally produce interference only during the starting cycle, although if the motor parts are badly worn they can create a terrific continuous racket. Always adjust this type of motor before attempting to apply noise filters. In general, brush type, series, shunt, and compound-wound ac and dc motors and generators are the principal noise makers.

12-677. Generator field excitation control leads will conduct and radiate interference. However, this type of noise can be eliminated with a bypass filter at the point where the leaks leave the generator housing. Noise from the ac leads of the alternator will frequently have its origin in the small dc generator used for filter excitation. Consequently, when working on ac generating equipment, you should clear up the interference of the exciter unit before checking the alternator.

12-678. Telephone equipment produces several kinds of noise, such as a low-pitched buzz or hum from the 30-cycle ringer, relay clicks, and dial tone noise.