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CHAPTER 13 TRANSMISSION LINE AND WAVE-GUIDE PRINCIPLES

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CHAPTER 13

TRANSMISSION LINE AND WAVEGUIDE PRINCIPLES AND MEASUREMENTS

13-1. INTRODUCTION. Any transmission arrangement consists of three essential parts: a source of energy, a transmission medium to transmit energy to a receiving device, and the receiving device, which usually converts the electric energy into some other form. Attached to a power transmission line, an electric generator can be used as the source of energy; high-voltage lines with a transformer at each end may be the transmitting medium; and a motor, lamp, or heater may be the receiving device for converting the electric energy into a useful form. In a long-distance telephone connection, a transmitter may be considered as the source of energy; the line from the speaking party to the listening party, with all of its associated conductors, coils, and connections, may be thought of as the transmission medium; and the telephone receiver may be considered as the third part of the transmission arrangement, or the device which converts small electric currents into audible vibrations of air, called sound waves.

13-2. In this chapter you are concerned with the transmission line or other medium through which energy is carried from the transmitting device to the receiving device. The chapter provides an analysis of the characteristics of ordinary wire conductors at relatively low frequencies that is, voice frequencies and carrier frequencies ranging up to approximately 150 kc) and of special transmission media such as waveguides at the higher frequencies, including radio frequencies extending up to several thousand megacycles.

SECTION I

TYPES OF TRANSMISSION LINES

13-3. GENERAL CLASSIFICATION.

13-4. Transmission lines may be broadly classified according to their use as either power or communication lines. Within each of these classifications, lines may be further described in terms of a particular characteristic (for example, high-tension lines in power circuits, long lines in telephone circuits, and tuned lines in radio circuits). In accordance with certain structural details, and regardless of its particular application, a given line may be accurately described as a single-wire, multiwire, or coaxial line.

13-5. SINGLE-WIRE OR GROUNDED LINES.

13-6. Historically, the invention of the telegraph developed the first need for extensive electrical conducting methods. Because it had already been discovered that the earth could be used as a return conductor of electricity, and because of the considerably greater economy, the first telegraph lines were single-wire, grounded lines, as illustrated in figure 13-1. Although single-wire lines were adequate for the early telegraph methods, harsh, unexplainable noises made conversation difficult or impossible when these lines were used to interconnect telephone instruments. When a number of single-wire telephone lines were run close together, the conversation on one line could be heard on all the other lines (cross-talk). With the expansion of power circuits and the building of electric railways, both of which were usually close to the telephone lines, single-wire telephone lines became practically useless because of the noise currents entering them from the power and railway circuits. Because of these troubles, single-wire lines are used as telephone lines today only in emergencies or where the added expense of more efficient lines is not justified. However, single-wire lines are still used in certain telegraph and radio circuits.

13-7. TWO-WIRE LINES.

13-8. Years of experimentation, together with the development of excellent mathematical analyses of line operation, proved that a metallic return (see figure 13-2) is the only proven cure for noise and cross-talk troubles. The need for a metallic return led to the construction of two-wire lines composed of two metallic conductors of the same material and spaced the same average distance throughout their length. When bare wires are used in such a line, the spacing must be great enough to prevent contact between the wires. As will be explained throughout the text, the conductor size, type of dielectric material, and spacing all affect line characteristics and must be chosen carefully; figure 13-3 shows various types of lines. The principal advantages of two-wire lines are their greater freedom from interference from external circuits and their considerably higher efficiency due to lower losses. Because of these and other advantages to be discussed later, two-wire lines are used extensively in radio circuits as well as in telegraph and telephone circuits. It might be well to mention here that a detailed study of two-wire lines provides the

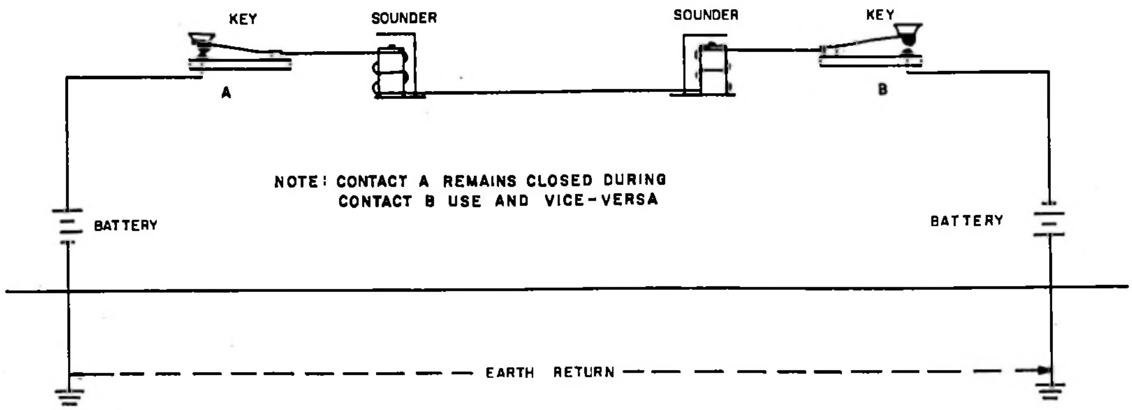


Figure 13-1. Earth Used as Return Conductor

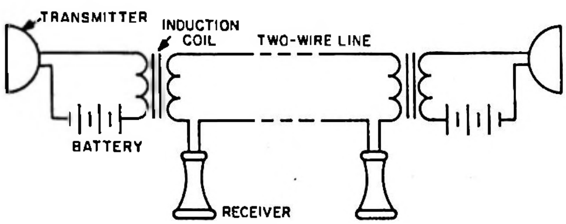


Figure 13-2. Metallic Return of Two-Wire Line

basic information needed for the study of coaxial lines, waveguides, and antennas.

13-9. THREE- AND FOUR-WIRE LINES.

13-10. As their names imply, these lines are similar in construction to the two-wire line, but have additional conductors. Multi-wire lines are used in polyphase power circuits, and may also be used in telephone and radio circuits to provide improved shielding and other advantages at a cost lower than that of coaxial lines.

13-11. COAXIAL LINES.

13-12. A coaxial line, usually called coaxial cable, consists of a cylindrical conducting rod or wire within an outer hollow cylindrical conductor (see figure 13-3). Because of the shielding of the inner conductor by the outer conductor, the transmitted energy is largely confined to the space between the inner conductor and the inner wall of the outer conductor; that is, none of the transmitted energy should appear on the outer surface of the cable. However, under certain conditions of operation, undesirable currents, called surface currents, may flow on the outer surface of the cable, and steps must be taken to prevent or reduce these currents to a minimum. Coaxial cables are particularly useful as rf (radio-frequency) transmission lines, partly because the shielding prevents radiation from the line and partly because of the broad band of frequencies that can be transmitted over these lines with little or no frequency distortion. Ordinary telephone lines transmit a relatively narrow frequency band extending from about 150 cycles to 2500 cycles, and special equalized telephone lines handle a somewhat broader frequency band of 60 cycles to approximately 10,000 cycles; however, a properly designed coaxial cable transmits with negligible frequency distortion a frequency band approximately 3 mc

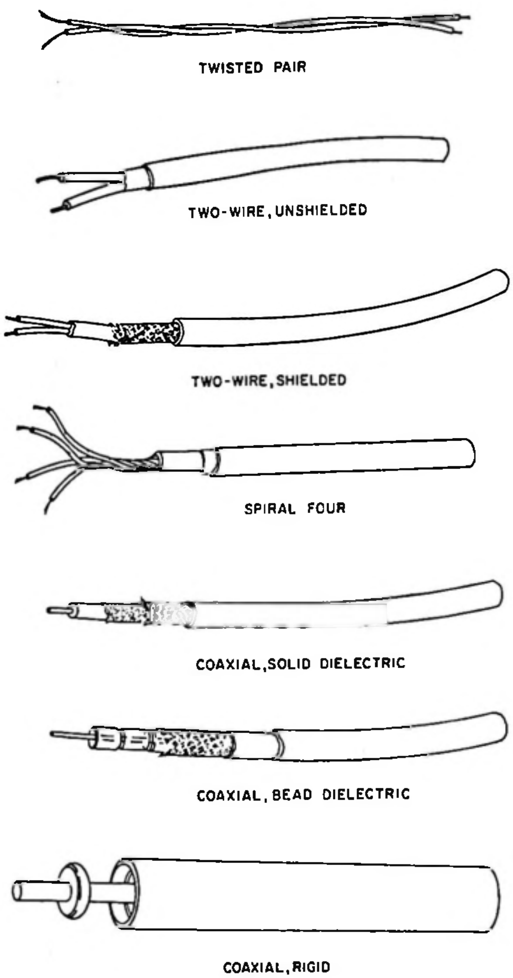


Figure 13-3. Types of Transmission Lines

(megacycles) in width. Thus, special coaxial cables are used to interconnect television stations in networks similar to those used for broadcasting speech and music.

13-13. LINE APPLICATIONS.

13-14. Although overhead power and communications lines spread across the nation like a gigantic spiderweb, the average person is scarcely aware of their existence, accepting them as part of the scenery. These lines are far more than ugly blots on the landscape; they make possible the telephone, the radio, the television, the washing machine, and thousands of other conveniences in the home. To power these devices, electrical and communications engineers have spent years of research and labor that have led not only to lines that carry electric energy from point to point, but also to lines that act as circuit elements, such as capacitors, inductances, and resistances. Thus (apparently in violation of certain fundamental electrical laws), transmission lines, which are constructed of very nearly perfect conducting materials, act as insulators, and hollow pipes (see figure 13-4) guide electromagnetic waves.

13-15. LINE THEORY.

13-16. Every transmission line operates in accordance with the fundamental laws of electricity and magnetism. Some of the more advanced concepts, particularly field concepts, are presented in the following sections of the chapter. Before beginning the study of transmission line theory, you should review the elementary concepts of magnetism discussed in Chapter 3. However, a brief repetition of the important points of Chapter 3 is provided in Section II of this Chapter 13.

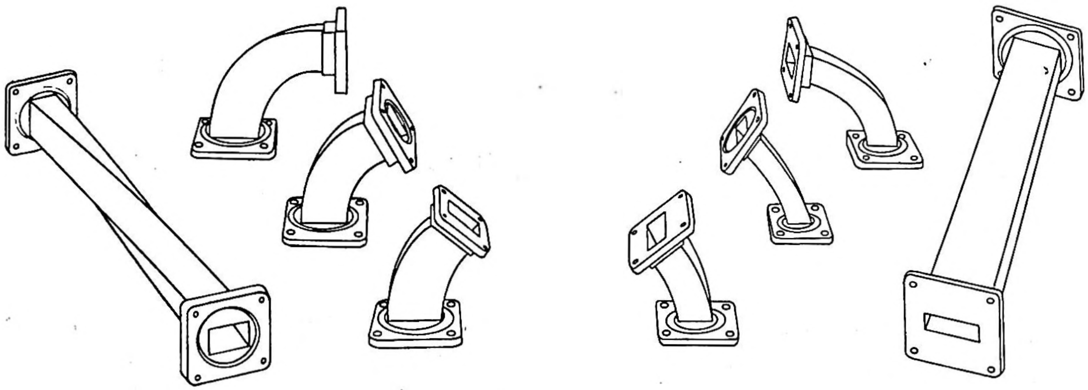


Figure 13-4. Typical Waveguide Sections

SECTION II

TRANSMISSION LINE PROPERTIES AND ELECTRICAL FUNDAMENTALS

13-17. CIRCUIT PROPERTIES.

13-18. Transmission lines possess the three inherent characteristics or properties of all electrical circuits: resistance, R; inductance, L; and capacitance, C. It is a natural law that all three electrical properties exist simultaneously together; therefore, it is impossible to design any circuit element to possess only one or two of these three characteristics.

13-19. LUMPED AND DISTRIBUTED PROPERTIES.

13-20. A circuit element can be designed to act as a relatively large quantity, or lump, of any one electrical property. This is done by reducing the other two properties to minimum values, so that their effects are negligible at the operating frequency of the circuit element. In part A of figure 13-5, the inductance is lumped as a relatively large quantity in coil L, and the other characteristics, represented by R and C, are distributed throughout the coil as minimum values. Capacitance C, in part B figure 13-5, is a lumped capacitance which necessarily possesses small distributed quantities of the other two electrical properties; resistance is present as shunt resistance R1 and as series resistance R2; the inductance, L, is due chiefly to the lead wires. Resistor R, in part C of figure 13-5, also has small distributed quantities of inductance L and capacitance C, as well as its large lumped quantity of resistance. At frequencies below which their distributed properties become important, coil L, capacitor C, and resistor R act as an inductance, a capacitance, and a resistance, respectively. At higher frequencies, the coil may act primarily as a capacitance or resistance, the capacitor primarily as an inductance or resistance, and the resistor primarily as an inductance or capacitance.

13-21. TRANSMISSION LINES AS CIRCUIT COMPONENTS.

13-22. Capacitors, inductance coils, and resistors are thought of as separate components which are inserted in an electrical circuit to place or localize definite quantities or lumps of capacitance, inductance, or resistance, according to the circuit requirements at the point of insertion. Similarly, transmission lines may be regarded in certain applications as components which are inserted in an electrical circuit for the purpose of producing the desired operation of the circuit as a whole. Transmission lines differ from other circuit components chiefly in the distribution of the electrical properties, L, C, and R; that is, L, C, and R are not present as localized or lumped quantities, but each is distributed evenly throughout the length of the line.

13-23. SCHEMATIC REPRESENTATION.

13-24. The simple schematic shown in part A of figure 13-6 represents a generator connected to a series circuit comprising a coil, L, a capacitor, C, and a resistance, R. The inductance, the capacitance, and the resist-

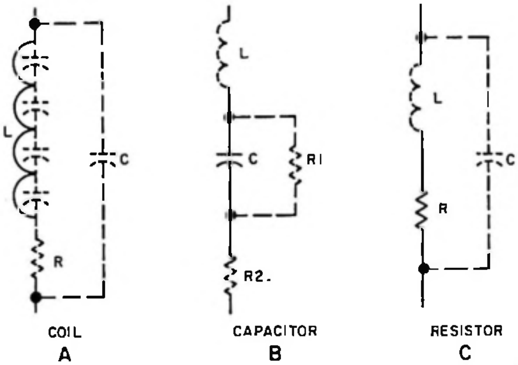


Figure 13-5. Basic Properties of Every Circuit

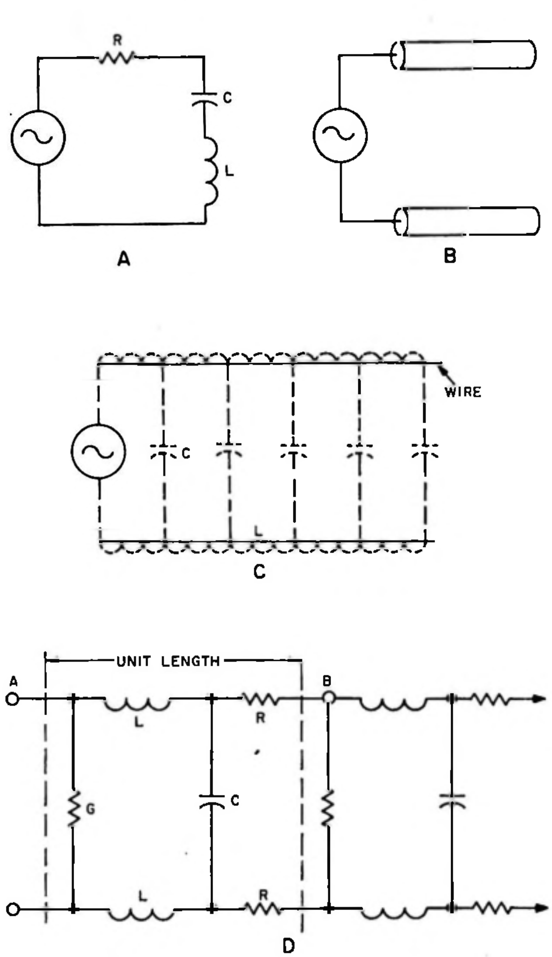


Figure 13-6. Distribution of R, L, C, and G Along a Transmission Line

ance are lumped quantities. In part B of figure 13-6 a two-wire transmission line, composed of two parallel wires, is connected to the generator terminals. In order to include the inductance and capacitance of such a line, the schematic shown in part B is redrawn in part C of the figure. This schematic shows that the inductance, represented by the dotted turns about each wire, and the capacitance, represented by the series of dotted capacitors between the wires, are distributed throughout the entire length of the line. A transmission line not only contains distributed inductance and capacitance, as illustrated in part C of figure 13-6, but also series resistance R and shunt conductance G. The series resistance is due to the inherent opposition of the wire to a flow of current, and the shunt conductance is due to the imperfect insulating quality of the medium between the wires. To include the series resistance and the shunt conductance in the schematic, the diagram shown in part C may be redrawn as shown in part D.

13-25. UNIT LENGTH.

13-26. In dealing with transmission line theory, it is convenient to consider each line in terms of unit lengths; that is, the line is divided into small lengths, each of which is known as a unit length. In part D of figure 13-6, the portion of the line between points A and B is one complete unit length; G is the shunt conductance, C is the distributed capacitance, L is the distributed inductance, and R is the series resistance. Each unit length is identical to every other unit length throughout a line.

13-27. GENERAL CONSIDERATIONS.

13-28. Resistance and shunt conductance cause a waste of power within a transmission line. In general, therefore, transmission lines are designed so that their resistance and shunt conductance are reduced to the minimum possible values. In well-designed rf lines the resistance and shunt conductance are so low that, for most purposes, they may be neglected. Of course, there are some rf line applications in which the resistance, minute as it may be, forms a basis for the action of the line.

13-29. It might be well to note at this time that not all transmission lines have the even distribution shown in part D of figure 13-6. The submarine cable and certain telephone cables, such as the spiral four (see figure 13-3), are excellent examples of this; they generally have large capacitance and relatively little inductance. In order to overcome this, coils (lumped inductances) may be inserted at intervals in the cable, adding or loading the line with the necessary inductance but producing an uneven distribution.

13-30. Whether composed of lumped quantities of L, C, and R, as in part A of figure 13-6, or of distributed quantities of L, C, and R, as in part D of the figure, an electrical circuit always operates in accordance with certain basic electrical phenomena. In dealing with low-frequency circuits made up of lumped values of L, C, and R, it is possible to explain their action in terms of voltage and current, without reference to the more basic concepts of electric and magnetic fields. In order to deal intelligently with circuits of the type illustrated in part D of figure 13-6, however, the student should have a more complete knowledge of basic electrical theories. For this reason, parts of this chapter are devoted to a discussion of electrical concepts that lead to a better understanding of transmission lines and coaxial cables.

13-31. KINDS OF ELECTRICITY (ELECTRIC CHARGES).

13-32. Electricity consists of two elementary particles, arbitrarily called positive

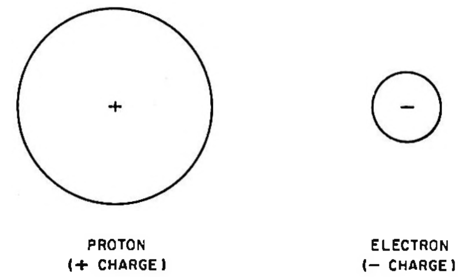


Figure 13-7. Equal Quantities of Opposite Kinds of Electricity

and negative. The positively charged particle is called a proton; the negatively charged particle is called an electron (see figure 13-7). The proton and the electron are elemental electric charges; although the mass of the proton is about 1800 times greater than that of the electron, the proton and the electron are equal quantities of opposite kinds of electricity.

13-33. ELECTRIC FORCE.

13-34. The behavior of every electrical circuit, from the simplest to the most complex, is due to the electric force that exists in the space about every elemental charge. This electric force produces an action on other charges at a distance; like charges repel one another, and unlike charges attract one another. As indicated in part A of figure 13-8, the force between two positive charges pushes them apart, and as indicated in part B, the force between two negative charges also pushes them apart. However, as indicated in part C, the force between two unlike charges pulls them together.

13-35. CHARGED BODIES.

13-36. NEUTRAL CONDITION. The atoms of all normal material bodies are thought to contain equal numbers of individual negative charges (electrons) and positive charges (protons). Because the electric force of an

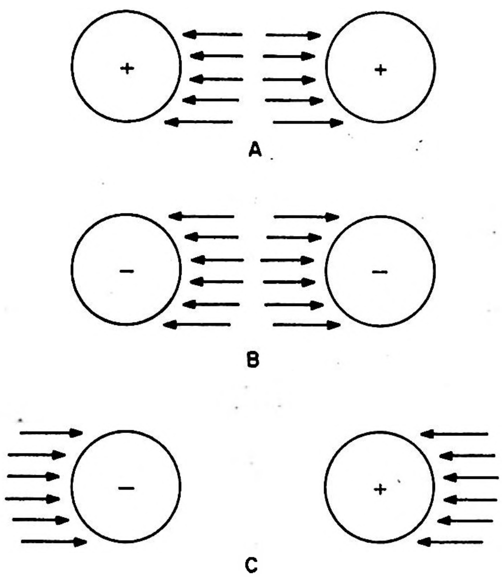


Figure 13-8. Repulsion of Like Charges and Attraction of Unlike Charges

electron is equal and opposite to that of a proton, the equal numbers of electrons and protons in a normal body produce no action on other electric charges at a distance. In other words, the electric force due to one kind of charge is neutralized by the electric force due to the opposite kind of charge, and the body is said to be neutral, or uncharged.

13-37. CHARGED CONDITION. A body is said to be electrically charged whenever the number of its electrons is not equal to the number of its protons; the excess of one or the other constitutes an electric charge. Hence, a body is negatively charged whenever electrons have been added to it, so that the number of its electrons exceeds its normal number of protons. The body is positively charged whenever its normal number of electrons has been reduced, so that the number of its protons exceeds that of its electrons.

13-38. ELECTRIC FIELD (LINES OF FORCE).

13-39. It has been stated that the forces acting between electric charges are called electric forces. The space surrounding an electric charge is called an electric field; that is, an electric field is the space in which the electric force about a charge is present to act on other charges that may be within the space. However, for simplicity, the space is given the property of the force within it; that is, an electric field is thought of as the force existing at every point in the space about an electric charge.

13-40. DIRECTION OF FIELD.

13-41. The direction of an electric field is arbitrarily defined as the direction in which the field tends to move a positive charge. Lines drawn to represent these directions are called electric lines of force. In part A of figure 13-9, the lines of force represent the directions of the field about a positively charged sphere; the lines of force about a negatively charged sphere are shown in part B of the figure. Although drawn only in the plane of the paper, the lines of force actually extend in every direction. If a posi-

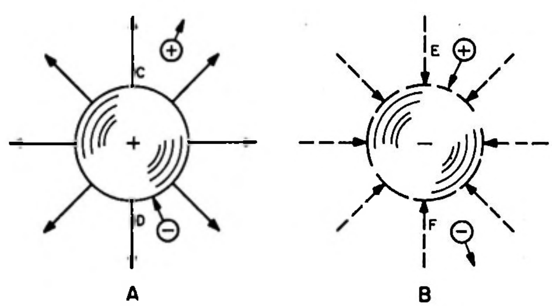


Figure 13-9. Electric Lines of Force Used To Indicate Direction in Which a Plus (+) Charge Is Urged by the Field

tively charged body, designated as C, is placed in the field about the positively charged sphere, it will be urged away from the sphere; the same field will attract a negatively charged body, designated as D, to the sphere. The short arrow on each of the inserted charged bodies indicates the direction in which it is urged. Because the sphere in part B of figure 13-9 is negatively charged, it attracts the positively charged body, designated as E, toward it, and repels the negatively charged body, designated as F.

13-42. FUNDAMENTAL CHARACTERISTICS.

13-43. An electric charge, whether positive or negative, cannot exist without an associated electric field; the charge and its associated field cannot be separated. The electric field of every charge extends to infinity, it exerts forces on all electric charges within it, and it possesses inertia. Because a charge and its field are inseparable, whenever a charge is in motion, its electric field must also be in motion.

13-44. RESULTANT FIELDS.

13-45. ZERO FIELD. If equal and opposite charges are placed together (see figure 13-10), the resultant individual fields will be coexistent and equal; neither field is made nonexistent by the other. However, the electric field of the positive charge (solid arrows) is in a direction opposite to that of the negative charge (broken arrows). Consequently, the electric field of the positive charge will urge an inserted positive charge, designated as A, in the direction of force f_1 (solid arrow), and the oppositely directed field of the negative charge will simultaneously urge the inserted positive charge in the direction f_2 (broken arrow). An inserted negative charge, designated as B, is also acted on by equal and opposite forces, f_1

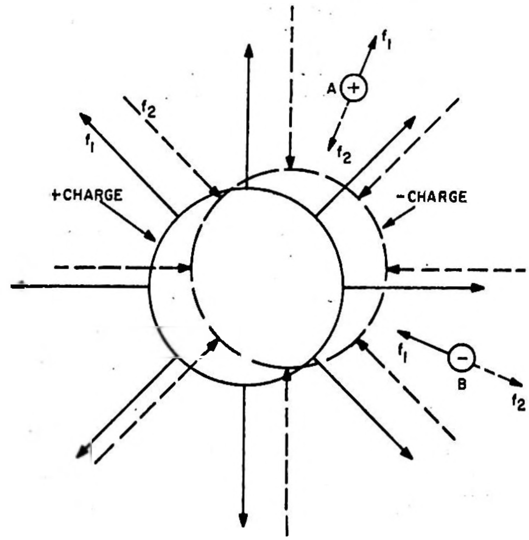


Figure 13-10. Fields of Individual Charges Combining To Produce a Zero Resultant Field

and f_2. Because these oppositely directed forces are exactly equal, the effect is the same as if neither field existed. In other words, the coexistent equal and opposite fields produce no action on inserted charges, and the resultant electric field is said to be zero. (In practice, the effects of the fields of inserted charges cannot be neglected. However, it is assumed here that the inserted charges are so small that their fields are negligible.)

13-46. CHARGES IN A LINE OF INFINITE LENGTH. If a number of positive charges are placed in a line of infinite length, as shown in part A of figure 13-11, the electric fields of the individual charges will be coexistent and combine to produce a resultant field, the direction of which is everywhere perpendicular to the line of charges. A similar line of negative charges is surrounded by a similar resultant field, except that the direction of the force is the opposite, as shown in part B of the figure. An un-

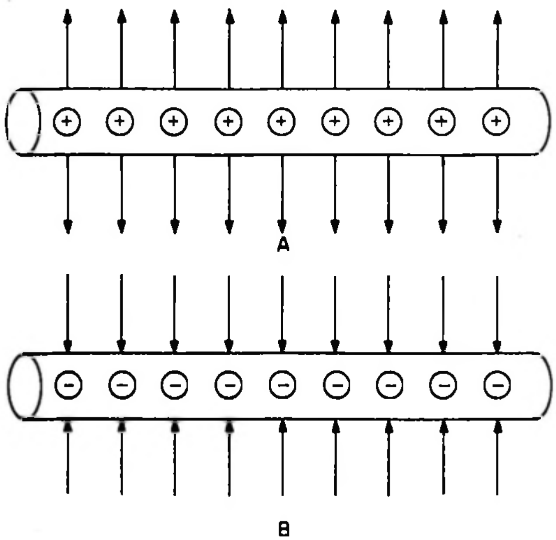


Figure 13-11. Direction of Electric Field Produced by a Long Line of Positive or Negative Charges

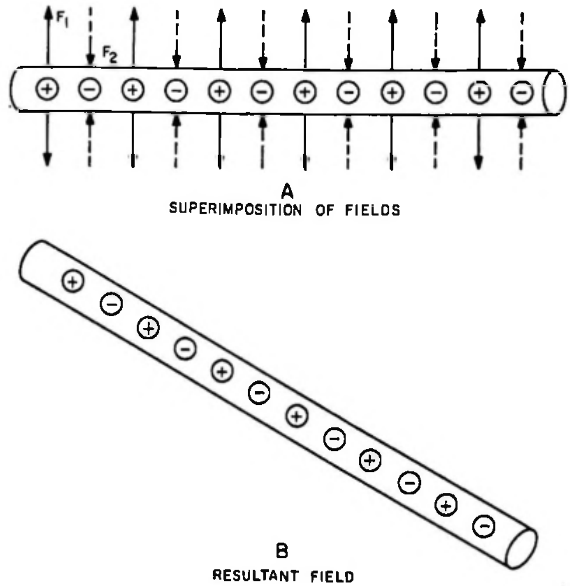


Figure 13-12. Effect of Combining the Equal but Opposite Fields Shown in Figure 13-11

charged conductor may be considered in terms of equal numbers of equal positive and negative charges arranged in a line. Part A of figure 13-12 shows crudely, and in only one plane, that the fields of the positive charges, f_1, and the negative charges, f_2, are superimposed (occupy the same space); therefore, the resultant field in the space is as though neither of the oppositely directed field exist. For simplicity, an uncharged body is usually pictured without any lines of force, as shown in part B of the figure; it is rarely necessary to show the elemental fields that combine to produce a resultant field.

13-47. REPRESENTATION OF FIELDS. In this manual, unless otherwise stated, lines of force are always drawn to represent the resultant field. Thus, in part A of figure 13-11, the lines of force represent the resultant field due to the excess number of positive charges in the conductor; the lines of force in part B of figure 13-11 represent the field resulting from the excess number of negative charges. Compare part A of figure 13-11 with part A of figure 13-12, and note the simplicity obtained in the former by considering only the excess charges in the conductor. It should be noted here that, although spoken of as though they were real, lines of force are imaginary and represent only a condition of the space through which they are drawn.

13-48. FIELD ABOUT SEPARATED CHARGES. The field strength, S, at a point in the field about a charge is proportional to the quantity of charge, Q, and is inversely proportional to the square of the distance, d, of the point from the charge; this is expressed by the simple formula: $S = Q/d^2$. In figure 13-13 equal charges, designated as A and B, are not superimposed as in figure 13-10, but are rather separated by a distance from A to B. The arrow on line A to X represents the direction that the electric field of the positive charge will urge a posi-

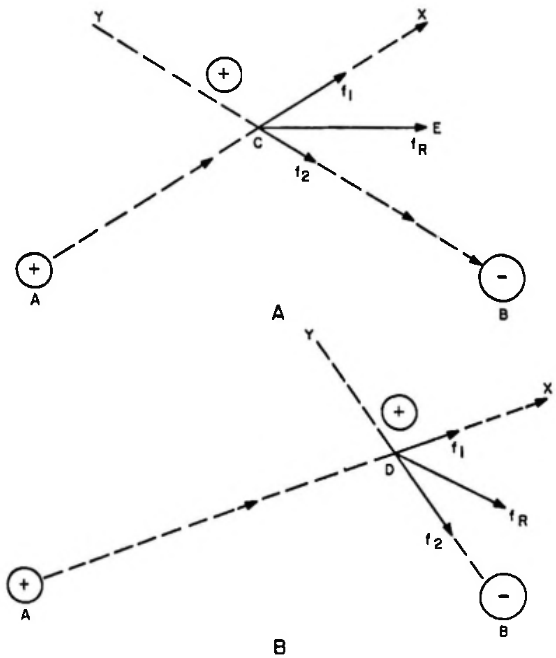


Figure 13-13. Direction of Resultant Field as a Function of the Individual Fields at Each Point

tive charged placed at any point along line A to X. The arrow on line Y to B shows the direction that the electric field of the negative charge will urge a positive charge placed at any point along line Y to B. In part A of figure 13-13 the quantity Q_A of the positive charge, designated as A, equals the quantity Q_B of the negative charge, designated as B, and the distance from A to C equals the distance from C to B. From the field strength formula, the field strengths produced at point C by the two charges have the following relationship:

$$S_1 = Q_A/d^2 = Q_B/d^2 = S_2$$

In words: At point C field strength S_1, due to the positive charge at point A, equals field strength S_2, due to the negative charge at point B. Hence, a positive charge placed at point C is simultaneously urged by a force, f_1, from C toward X and by an equal force, f_2, from C toward B. Obviously, the charge cannot move in two directions at the same time, but forces f_1 and f_2 acting together produce at C a resultant force, f_R, in the direction of the line from C to E. Consider now the condition of the field at point D in part B of figure 13-13. Distance A to D is greater than distance D to B; therefore, force f_1 is less than force f_2. Under this condition the resultant force, f_R, at point D is downward and slightly to the right. If a free positive charge is placed at numerous points in the combined fields of charges A and B, it will be found that the superimposition of the individual fields will produce a different resultant condition at every point. Thus, a free positive charge, placed at point C in figure 13-14, will come under the influence of a different resultant force as it leaves C and follows the curved path from point C to B. Similarly, a positive charge placed at point D will follow the curved path from point D to B. Note that the curved resultant line of force from point A to C to B is perpendicular at its ends to the surfaces of point A and point B. In figures 13-14 and 13-15 a few of the elemental lines of force are indicated by broken lines. The solid lines in figure 13-14 represent the resultant lines of force in the electric field about two

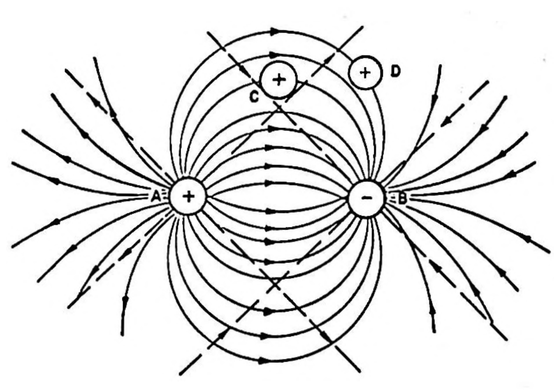


Figure 13-14. Field about Equal Unlike Point Charges

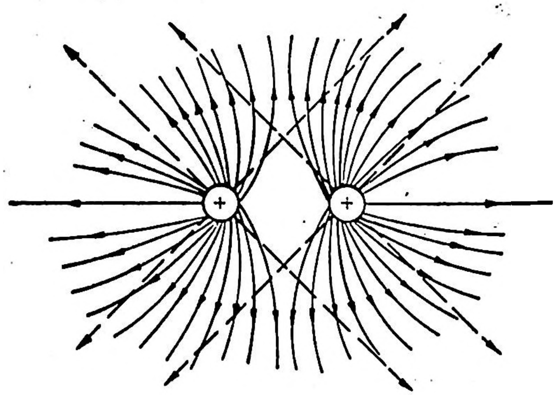


Figure 13-15. Field about Equal Like Point Charges

equal and unlike charges; the solid lines in figure 13-15 represent the resultant lines of force about two equal and like charges. As used in these figures, each line of force merely indicates that a force acts on a positive charge placed at any point on the line, in the direction of that line, and that the force acts in the opposite direction on a negative charge placed at the same point on the line.

13-49. ELECTRICAL UNITS.

13-50. Up to this point only elemental charges have been mentioned as a unit quantity of electricity, but in order to cope intelligently with practical work, it is necessary to establish other electrical units. You are undoubtedly familiar with the practical units, such as the volt, ampere, ohm, farad, etc. These practical units were developed from more basic units. When first used in this manual, units other than practical units will be defined.

13-51. UNIT OF FORCE.

13-52. The commonly used unit of force is the dyne; this unit is defined as the force which, acting on a mass of 1 gram (.035 ounce) for 1 second, increases its velocity by 1 centimeter per second.

13-53. ELECTROSTATIC UNIT—QUANTITY OF ELECTRICITY.

13-54. This unit is defined as that quantity of electricity which, when concentrated at a point in a vacuum, repels an equal quantity of like charge at a distance of 1 centimeter (.3937 inch) with a force of 1 dyne, or attracts an equal quantity of unlike charge with the same force.

13-55. THE COULOMB.

13-56. The cgs (centimeter-gram-second) electrostatic unit of charge is the statcoulomb. The number of elemental charges in a statcoulomb has been calculated to be greater than 2,000,000,000. Even this huge number of elemental charges provides such a very small unit that it is not used in practical electricity; the practical unit is the coulomb, 1 coulomb being equal to 3×10^9 statcoulombs, or 3,000,000,000 statcoulombs. In terms of elemental charges, the coulomb is equivalent to 6.28×10^{18} electrons (or protons). Although not used as frequently as the volt, ampere, and ohm, the coulomb is a practical unit and is a measure of quantity, like gallons and bushels. The relationship of the coulomb to other units may be clearer when it is realized that if 1 coulomb of electricity is passing a point in a circuit in 1 second, the current is 1 ampere. Thus, when the current in amperes is known, the number of electrons passing any point in a circuit can be calculated.

13-57. FORCE BETWEEN CHARGES.

13-58. Experiment has shown that the force between two charges (in a near vacuum) varies inversely as the square of the distance between them (see figure 13-16), and directly as the product of the two quantities

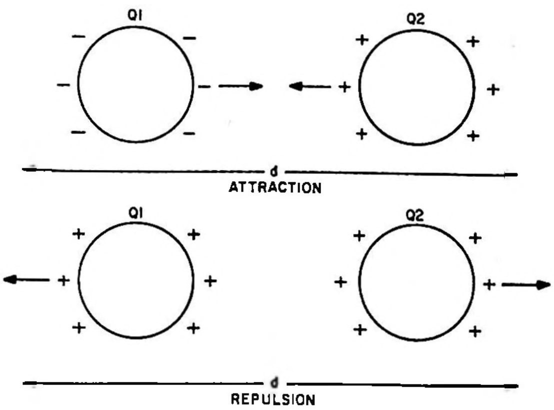


Figure 13-16. Force Between Charges

of charge. Thus, regardless of the kind of charge, the combined effect of quantity and distance is expressed by the simple equation:

$$F = Q_1Q_2/d^2$$

where:

F = Force between the two charges

Q_1 = Quantity of one charge

Q_2 = Quantity of the other charge

d = Distance between Q_1 and Q_2

For most practical purposes, the force existing between charges in air is considered to be equal to the force existing between charges in a vacuum. If Q_1 and Q_2 are like charges, the force between them is one of mutual repulsion; if they are unlike charges, the force between them is one of mutual attraction. Use of the force formula is illustrated in figure 13-17, where the figure F represents the force between charges.

13-59. ELECTRICAL FIELD STRENGTH.

13-60. DEFINITION.

13-61. The electric field strength (intensity) at any point in the field about a charge is defined as the force which would act on a unit positive charge (placed as a test charge) at that point in the field. This definition comes from the basic force equation: $F = Q_1Q_2/d^2$. With field strength defined as the force acting on a unit positive charge placed at a point in the field, the field strength at various points in the space about one charge, Q_1, is determined by maintaining a second

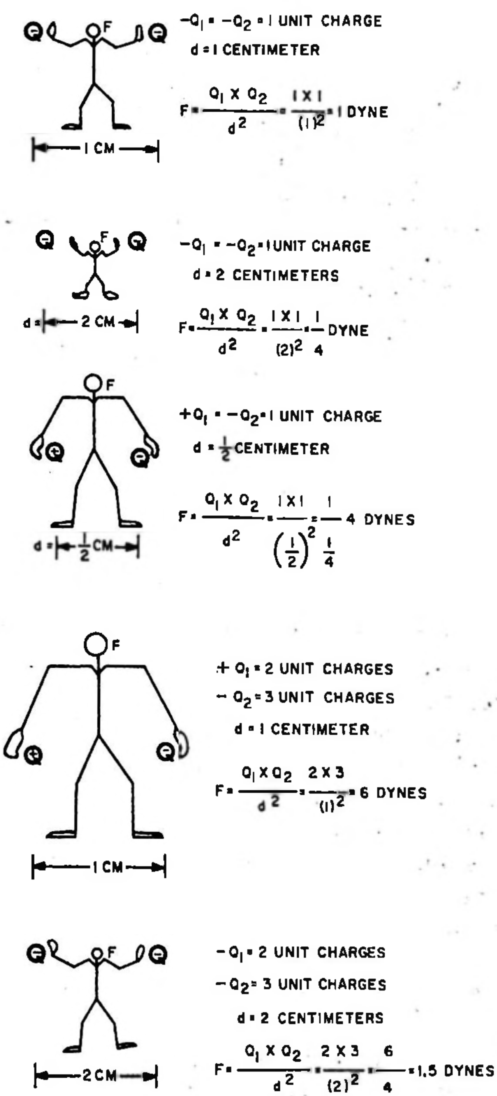


Figure 13-17. Calculation of Force Between Charges

charge, Q_2, at unit value and moving it as a test charge at various distances from the first charge. Thus, in calculating field strength, Q_2 is always numerically equal to 1; consequently, it can be dropped from the equation. The equation for field strength, S, at various points in the field about any charge then becomes $S = Q_1/d^2$. The illustration in figure 13-18 show the application of the field strength formula in determining the field strength at points in the field about different quantities of charge. Note that the figure S is shown as acting only on the inserted test charge Q_2. This is because field strength is the force at a point—not the force between charges.

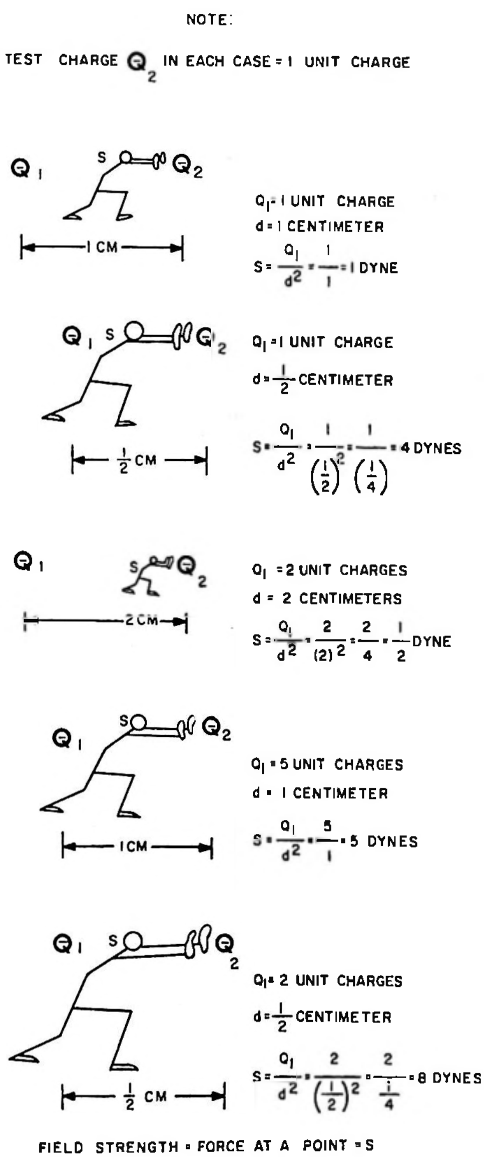


Figure 13-18. Field Strength, S, at a Point

13-62. TOTAL FORCE AT A POINT.

13-63. If the field strength at a point is known, the total force at that point, when the inserted charge is greater or less than unity, can be found by multiplying the field strength by the quantity of inserted charge:

$$F = SQ_2$$

where:

F = Force

S = Field strength at point of insertion

Q_2 = Quantity of inserted charge

The various parts of figure 13-19 show how the total force is calculated when only the field strength and the quantity of the inserted charge, Q_2, are known.

13-64. FIELD POTENTIAL.

13-65. The difference between field potential and field strength must be thoroughly understood. Field potential is a measure of work, and field strength is a measure of force. Perhaps this difference can best be understood from the simple equations for electric force and electric potential. The force equation is $F = Q_1Q_2/d^2$, and the potential equation is $P = F \times d$. Bearing in mind that potential P equals work W, the potential equation can be written as $P = W = F \times d$. As previously explained, the force at a point (field strength S) is determined by maintaining Q_2 of the force equation at unit

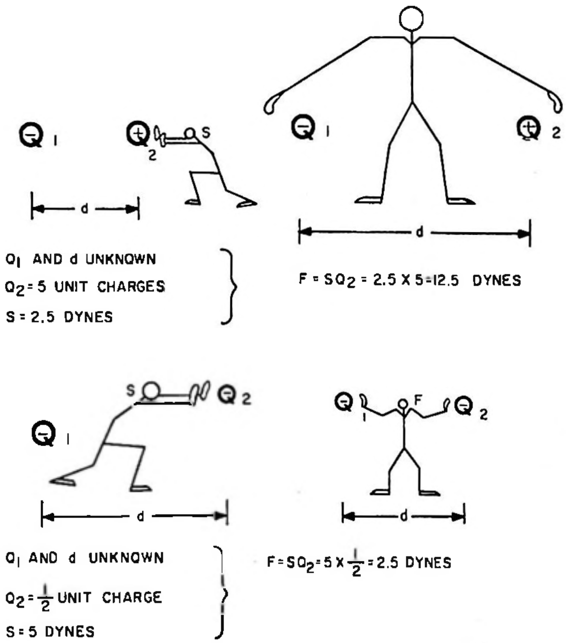


Figure 13-19. Field Strength, S, Versus Quantity of Inserted Charge, Q_2

value (equal to 1); therefore, $S = Q_1/d^2$. Substituting Q_1/d^2 (from the field strength equation) for F in the potential equation, the potential equation becomes:

$$P = W = Q_1/d^2 \times d$$

$$= Q_1/d$$

The field strength and field potential equations are shown below for comparison.

$$S = F = Q_1/d^2$$

$$P = W = Q_1/d$$

You can see that (1) field strength is a measure of force, but field potential is a measure of work; (2) both field strength and field potential vary directly with the quantity of charge, Q_1; and (3) field strength varies inversely as the square of the distance, d, of the point from the charge, but field potential varies inversely as the distance, d.

13-66. EFFECT OF NEARBY CHARGES ON POTENTIAL.

13-67. An electric charge and its field are inseparable. For this reason, the potential at a point is not only a function of the quantity of charge (if any) at that point; it is also a function of the kind and quantity of other charges at nearby points.

13-68. EFFECT OF INDIVIDUAL CHARGE.

13-69. Assume, for example, that with a quantity of plus charge, Q_1, at point P in part A of figure 13-20, the potential of P is 4 volts. Suppose further that, instead of the quantity of plus charge Q_1, an equal quantity of negative charge Q_2 is at point P in part B of figure 13-20, and that the potential of P is —4 volts. Considering only the field lines between points D and G, values representing the potentials at a few points are shown in parts A and B of figure 13-20. These individual potentials are a representation of the field when it is considered that all other charges are too far away to produce any noticeable effect.

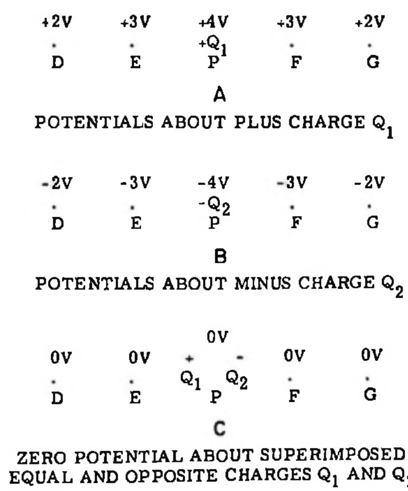


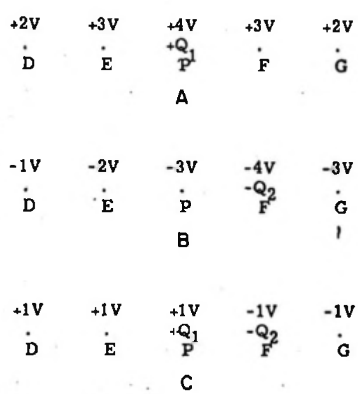
Figure 13-20. Potential at a Point Affected by all Nearby Charges

13-70. EFFECT OF SUPERIMPOSITION.

13-71. In part C of figure 13-20, the negative charge Q_2 is superimposed on the positive charge Q_1, and the potential at P and at all other points is zero. This is true because the superimposed equal and unlike charges produce everywhere a neutral condition. Note that part C of figure 13-20 is the equivalent of superimposing part A on part B; the potentials at the points in part C are obtained by algebraic addition of the potentials at corresponding points in both parts A and B.

13-72. EFFECT OF SEPARATED CHARGES.

13-73. In part A of figure 13-21, the conditions are identical to those in part A of figure 13-20, but in part B of figure 13-21, negative charge Q_2 is at point F instead of at point P. When the two charges, Q_1 and Q_2, are placed as shown in part C of figure 13-21, the potential at P is reduced from plus 4 volts to plus 1 volt, because of the superimposition at point P of a field, due to Q_2, which has a negative potential of minus 3 volts at point P. Similarly, the potential at point F is raised from minus 4 volts to minus 1 volt; that is, because of the nearby positive charge, Q_1, the effect of the negative charge, Q_2, is reduced. Part C of figure 13-21 represents the equivalent of superimposing part A on part B, and the potentials in part C are obtained by algebraic addition of the potentials at corresponding points in both parts A and B.



AT P:
POTENTIAL DUE TO Q_1 = +4
POTENTIAL DUE TO Q_2 = -3
POTENTIAL AT P = +1V

AT F:
POTENTIAL DUE TO Q_1 = +3V
POTENTIAL DUE TO Q_2 = -4V
POTENTIAL AT F = -1V

POTENTIAL DIFFERENCE PD_{P-F} = +1-(-1) = 2V

Figure 13-21. Nearby Charges Versus Potentials and Potential Differences

13-74. POTENTIAL DIFFERENCE AND POTENTIAL.

13-75. The potential difference between points P and F in part C of figure 13-21 is:

$$PD = P - F = +1 - (-1) = 2 \text{ volts}$$

Thus, the potential at a point is the algebraic sum of the individual potentials at that point, but the potential difference between two points is equal to the algebraic difference between the potentials at the two points. It should be noted that the potential at a point is not clearly defined unless the polarity is indicated, but potential difference is never given polarity. Potential is a vector quantity (it has both magnitude and direction), and potential difference is a scalar quantity (it has magnitude only). You have probably noticed that points P, E, and D in part C of figure 13-21 are at the same positive potential, and points F and G are at the same negative potential. Points at the same potential are called equipotential points, and are discussed in the following paragraph.

13-76. EQUIPOTENTIAL LINES.

13-77. As previously explained, lines of force are drawn to describe the field of force about an electric charge. Lines may also be drawn to describe the field of potential about electric charges. Such lines are called equipotential lines, and they usually represent surfaces along any one of which the potential is constant.

13-78. SPHERICAL SURFACES.

13-79. Consider that a charge Q of 12 units is at the center of a sphere 2 units in radius, as shown in part A of figure 13-22. The potential of the sphere is 6 units or, assuming practical units, 6 volts. At point A, 3 units of distance from the central charge, the potential is Q/d = 12/3 = 4 volts. Now, is this equation is used to determine the potential at every point 3 units of distance from the central charge, it will be found that the potential at every such point is 4 volts. Together these equipotential points form a spherical surface about the charge Q. Thus, although drawn as circles in the plane of the paper, the equipotential lines in part A of figure 13-22 actually represent concentric spheres about charge Q.

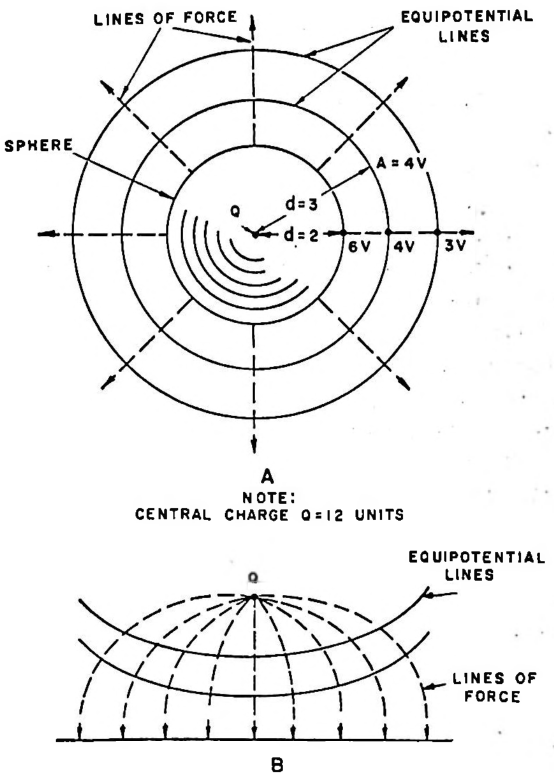


Figure 13-22. Equipotential Lines

13-80. FORCE AND EQUIPOTENTIAL LINES.

13-81. Some of the equipotential lines about a point charge close to a flat plate are represented by solid lines in part B of figure 13-22, and a portion of the field of force is represented by dashed lines. Notice that at every point of intersection, lines of force and equipotential lines are mutually perpendicular. This means that there is no component of force along an equipotential surface.

13-82. CONDUCTORS AND INSULATORS.

13-83. The atomic structure of certain materials is such that their electrons move freely, but in a haphazard manner, from atom to atom. Other materials do not permit such a haphazard movement of electrons within them, and hence do not have free electrons. When placed in an electric field, the materials containing large numbers of free electrons permit these electrons to flow readily through them, and are called conductors. Because of the closeness with which their electrons are bound to the atoms, the other materials do not permit such a flow of electrons, and are called nonconductors, insulators, and dielectrics. The ability to conduct electrons varies in different materials; consequently, there are good and poor conductors, and good and poor insulators.

13-84. SIMPLE CONDUCTION.

13-85. When a conductor is connected directly across two points between which there is a potential difference, the electric field between the two points necessarily extends through the conductor, and the electrons in the conductor are urged from points of lower potentials to points of higher potentials. A simple example of this concept is shown in figure 13-23. Assume that chemical action, represented by the dashed-line arrows within the battery, removes electrons from plate 1 so that it becomes positively charged and, at the same time, adds electrons to plate 2, so that it becomes negatively charged. This is a separation of electric charges, and an electric field now exists between the two plates; the excess charges on each plate possess potential energy which tries to force the charges through the battery toward the opposite plate. The positive charges on plate 1 cannot move because they are bound to the atoms in the metallic conductor making up that plate. The excess electrons on plate 2 possess a potential energy which tries to force the electrons through the battery, as indicated by the solid arrows, but the chemical action of the battery prevents this movement of electrons within the battery.

13-86. SIMPLIFIED EXPLANATION.

13-87. Conductor R contains electrons that are free to move under the influence of an electric field. When conductor R is connected between the two plates, point X is connected directly to plate 1 and assumes the potential of plate 1; point Y is similarly connected to plate 2 and assumes the potential of plate 2. Thus, an electric field extends through R as indicated by the solid arrows through R. (For convenience, these lines of force are drawn to show the direction in which the field will urge electrons.) The positive charges in conductor R are bound to the atoms, and, therefore, cannot move under the influence of this electric field. Many negative charges (electrons) in the conductor are free to move, and, under the influence of the electric field, do move in the direction indicated by the solid arrows. At the instant conductor R is connected, every electron in the conductor comes under the influence of the field, and each electron is urged from points of lower potential (from all points below X) toward points of higher potential (at X). As free electrons move upward through R, the potential of X is lowered by the flow of electrons to it; at the same time, the potential of Y is raised by the flow of electrons from it. However, electrons at X are attracted to points of higher potential on plate 1, and other electrons are drawn from the lower potential at plate 2 toward the increased potential at Y. This drift of electrons, between the point of lowest potential (at plate 2) and the point of highest potential (at plate 1), tends to reduce the potential difference between plates 1 and 2. However, the chemical action of the battery maintains the potential difference between plates 1 and 2. Thus, there is an electric current through the conductor and the battery as long as this potential difference is maintained. When a generator of electric current and its load are directly connected, as shown in figure 13-23, the simple explanation previously discussed can be used to explain the transmission of energy from the generator to the load. However, when the generator and the load are separated by appreciable distances, it is necessary to include an explanation of how the electric field

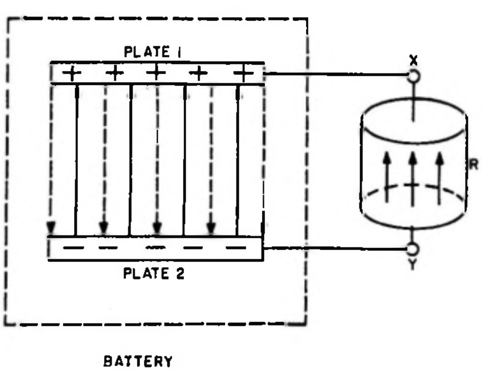


Figure 13-23. Simplified Concept of Conduction

within the generator is transmitted over these distances. Such explanations must be made in terms of electric and magnetic fields (electromagnetic field).

13-88. ELECTROMAGNETIC FIELD.

13-89. It can be shown that whenever an electric field is in motion, there comes into being a force which is not measureably present while the electric field is at rest. This additional force is called a magnetic force, and the space in which it is present is called a magnetic field. In this text a magnetic field is considered as an attribute or property of an electric field in motion. A moving electric field and its associated magnetic field always exist together, and the space in which the two fields are co-existent is called an electromagnetic field.

13-90. FIELD ABOUT ELECTRON IN MOTION.

13-91. ELECTRIC FIELD. In figure 13-24, assume that the electron is moving through space at a constant speed and that its direction of motion is away from the observer.

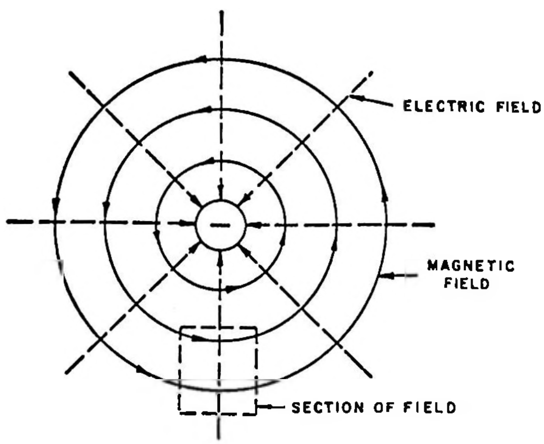


Figure 13-24. Electromagnetic Field about an Electron Moving Away from Observer

Since the electron and its field are inseparable, the electric field is necessarily moving in the same direction as the electron, that is, away from the observer. The broken-arrow lines of force do not represent the direction in which the electric field is in motion; these lines of force represent only the direction of the field force. Considering that the electron motion is uniform, the electric lines of force can be represented as straight lines.

13-92. MAGNETIC FIELD. The magnetic field is represented by concentric circular lines of force, with arrowheads indicating the direction of the magnetic force about the electron. The magnetic field is due to the motion of the electric field; consequently, the magnetic field must also be moving away from the observer. Whenever an electron is moving away from an observer, its electric and magnetic fields also move away from the observer, and the direction of the resultant circular magnetic field is counterclockwise.

13-93. CONSIDERATION OF FIELD ONLY.

13-94. When you are dealing with electromagnetic waves, you must usually consider the electromagnetic field without reference to the electron (or other charge) displacement that produced it. For example, if the section of the electromagnetic field in figure 13-24 is redrawn as shown in figure 13-25, so that it can be considered without reference to the electron, the following rule is applicable: A moving electric field and its associated magnetic field are coexistent and so related that, if they are moving from an observer with the electric lines pointed upward, the magnetic lines will point to the right. At every point in an electromagnetic field the electric field is at right angles to the magnetic field. Also, the direction of the force in each field is at right angles to the direction in which the fields are in motion. This relationship is usually expressed

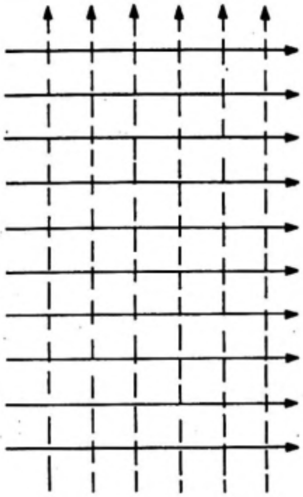


Figure 13-25. Electromagnetic Field Moving Away from Observer

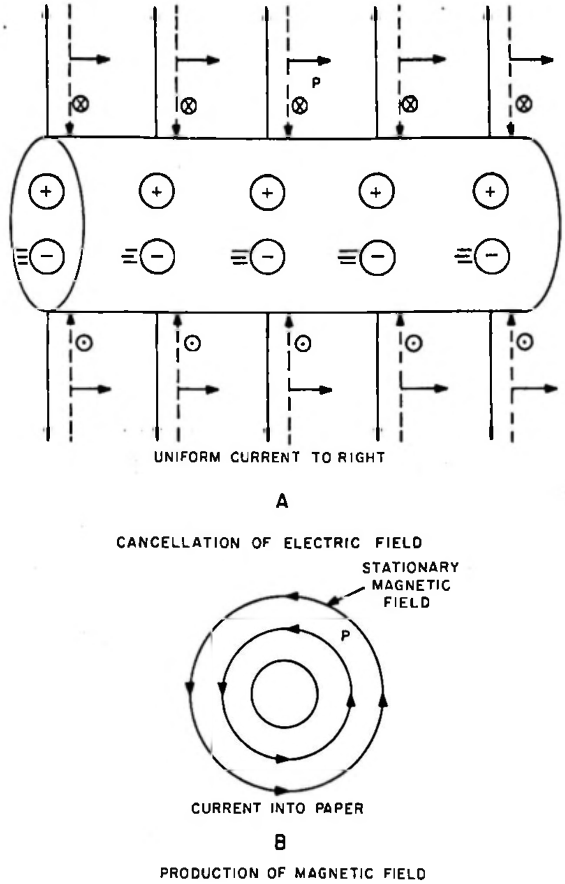


Figure 13-26. Electric and Magnetic Field Effects of a Uniform Current

by saying that the electric and magnetic fields are perpendicular to each other and to the direction of motion.

13-95. STATIONARY MAGNETIC FIELD.

13-96. Consider the flow of current through a conductor. First assume that a uniform current is flowing from left to right in the conductor shown in figure 13-26. The solid arrows in part A of this figure than represent the electric field due to the stationary protons (positive charges) within the wire, and the broken arrows represent the electric field due to the electrons. The short arrows pointing to the right indicate that the electron field is moving to the right. Despite this movement of the electron field to the right, there is no resultant electric field about the wire, because the two electric fields cancel each other at all points. However, the electron field is an electric field in motion and produces about the wire a circular magnetic field. Immediately above the wire, the direction of this magnetic field is into the paper, and immediately below the wire its direction is out of the paper, as indicated by the conventional circled crosses and dots. Since the current is constant, an observer looking along the wire in the direction of the electron flow would see only a stationary circular magnetic field about the wire, as shown in part B of the figure. If an electron were at rest at point P in the stationary magnetic field, no force would be exerted on it because (1) there is no resultant electric field about the wire, and (2) a magnetic field has no effect on an electric charge unless there is relative motion between the charge and the magnetic field.

13-97. ELECTROMAGNETIC PULSE.

13-98. For simplicity, part A of figure 13-26, is redrawn in part A of figure 13-27

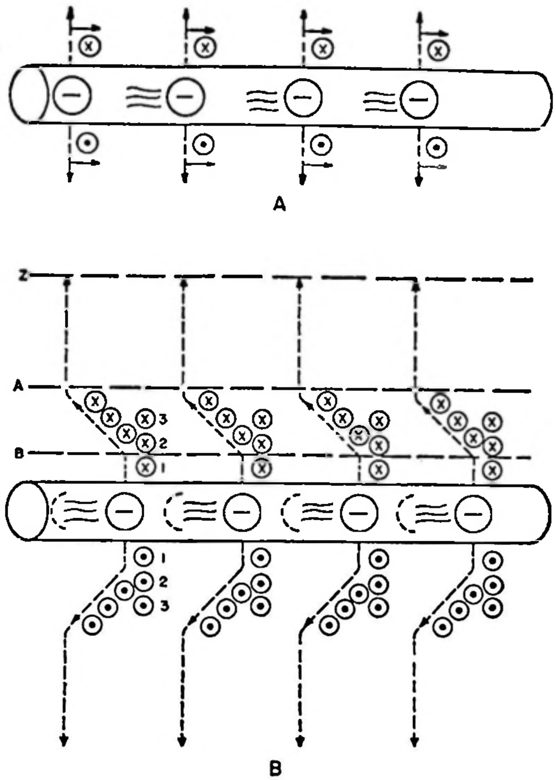


Figure 13-27. Production of Electromagnetic Pulse and Growth of Magnetic Field

to show a smaller number of moving electrons; the positive field is omitted, and the electric lines of force indicate the direction in which an electron is urged. Assume that the electron motion to the right is uniform, so that the moving electric field can be considered to be made up of straight lines of force. As in the preceding figure, the stationary magnetic field is represented by the circled crosses and dots. The short arrows to the right indicate the direction in which the electrons and their electric fields are in motion. Suppose, now, that the current is increasing to a new and higher uniform value. While the current is increasing, the velocity of the electrons is also increasing. Consider that the electrons shown in part A of figure 13-27 have reached the positions shown in part B of the figure at the instant that the current and, hence, the electron velocity again become constant. Because the electric field possesses inertia, not all parts of this field moved simultaneously to the right as the velocity of each electron increased. The portion of each line of force represented as being immediately above each electron has very nearly kept up with the electron motion, and is, therefore, adjacent to its electron. However, the portion of each field line between horizontal reference lines A and B has not kept up with its electron and is now distorted. The electric force in this portion is now to the left, that is, opposite the direction of electron acceleration. The portions of the electric field lines between horizontal lines A and Z are still in their original positions; that is, because of their inertia, the electric field lines between A and infinity (Z) have not yet moved.

13-99. EFFECT OF DISTORTED FIELD ON ELECTRONS.

13-100. An electron in the distorted portion of the electric field would be urged to the left; that is, it would be urged in a direction opposite that of the electron flow in the conductor. (This phenomenon forms the basis for self-inductance and mutual inductance, which will be explained later.)

13-101. PRODUCTION OF PULSE.

13-102. The electric lines of force shown in part B of figure 13-27 must straighten out and become perpendicular to the surface of their associated individual electrons. In order to bring about this straightening of the electric lines, the distorted portions must move outward from the wire. These distorted portions also constitute a moving electric field, which produces an accompanying magnetic field. This electromagnetic field moves outward from the conductor with the velocity of light. It is called an electromagnetic pulse.

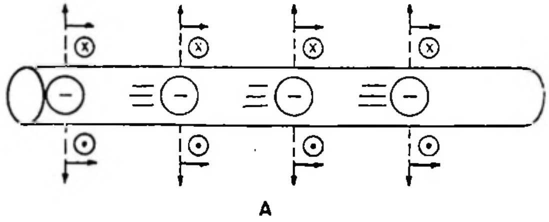
13-103. MOVING MAGNETIC FIELD.

13-104. GROWTH OF MAGNETIC FIELD.

13-105. As the electromagnetic pulse moves outward (see part B of figure 13-27), it adds energy to the space through which it moves; that is, in moving outward the pulse carries and leaves behind it the energy which will constitute the stronger stationary magnetic field about the higher uniform current through the conductor. This concept is shown within the limitations of a simple drawing in part B of figure 13-27. Consider the instant that the distorted portion of the electric field is in the position shown. A stationary circular magnetic line, represented by 1, has been established close to the conductor; a second magnetic line, 2, is being established, and line 3 will be established as the distorted portion moves out from the conductor. At the same time that the magnetic field is being built up about the conductor. the distorted electric field is diminishing and the resultant electric field is becoming zero. This is indicated by the straight electric lines between the conductor and horizontal line B. While the electron velocity (that is, the current) is changing, the field about the conductor is quite complex. For simplicity, it has become customary to neglect the electromagnetic pulse and consider only the changing magnetic field, except when dealing with circuits designed to produce a maximum of radiation.

13-106. DECREASE OF MAGNETIC FIELD.

13-107. Now consider the electromagnetic pulse produced when the current and, hence, the velocity of the moving electrons are reduced. For this study, assume that the electron motion in the conductor is uniform and to the right, as shown in part A of figure 13-28. Under this condition, the lines of the electric field (drawn to indicate the direction in which an electron would be urged) are straight, and there is a stationary magnetic field about the conductor.



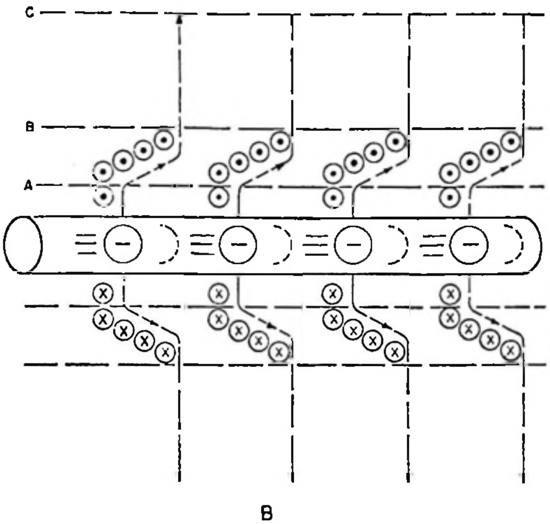


Figure 13-28. Production of Electromagnetic Pulse and Reduction of Magnetic Field

13-108. EFFECT OF ACCELERATION. Suppose that the current is suddenly reduced to a lower uniform value. At the moment the current is reduced, the electric lines in A are moving at a constant speed to the right and their inertia tends to keep them moving at the same velocity; therefore, these electric lines do not instantly change their speed of travel to the lower speed of the electron. In consequence of this, at the instant that the electron velocity has again become uniform (see part B of figure 13-28), the portions of the field lines between reference lines B and C have moved farther to the right than the portions of the lines between the conductor and reference line B. In other words, lines between B and C have moved approximately to the positions that they would have occupied if the current had not been reduced. Note that the direction of the distorted lines and, hence, the electric force between lines A and B are now to the right; that is, the distorted portion will urge electrons in the same direction that the conductor current is flowing.

13-109. REDUCTION OF MAGNETIC FIELD. As in the case of the increasing current previously discussed, the distorted portions of the electric lines move outward from the conductor. In this case, however, the moving electric lines are pointed to the right and their associated magnetic field lines point out of the paper. In moving outward, this pulse cuts through the magnetic field already existing about the wire. Since the magnetic component of the pulse itself is opposite that of the existing magnetic field, the existing magnetic field is weakened. Note that the direction of the existing magnetic force is not changed, but the resultant stationary magnetic field becomes weaker than that which existed before the current was reduced.

13-110. CHANGING MAGNETIC FIELD.

13-111. As previously mentioned, this complex action may be considered in terms of the magnetic field about the wire. In the case of an increasing current, the magnetic field is said to be expanding; therefore, its lines are moving out from the wire. In the case of a decreasing current, the magnetic field is considered to be collapsing; therefore, its lines are moving toward the wire. It has been stated that a moving electric field and its associated magnetic field are always coexistent. The converse is also considered as a basic phenomenon; that is, a moving magnetic field and its associated electric field are always coexistent.

13-112. SELF-INDUCTANCE.

13-113. Although usually explained in terms of the changing magnetic field (Chapter 3), self-inductance can be explained simply in terms of the moving electric field about the charges in motion. First consider the electron fields when the electron velocity is increased in conductor C, shown in part A of figure 13-29. Before the switch, SW, is closed, the current is zero. When the switch is closed, the current would change instantly from zero to its maximum value of I = E/R if the change were not opposed by the self-inductance of the conductor.

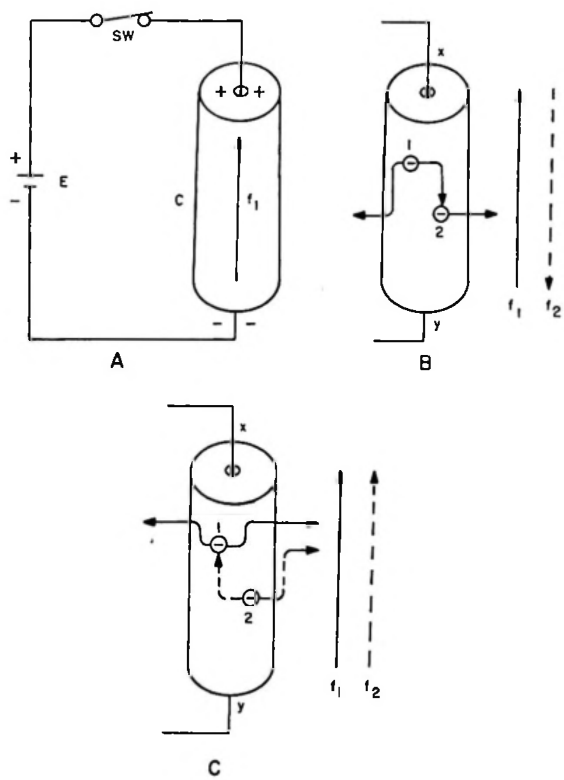


Figure 13-29. Self-induction in Terms of Electric Fields

13-114. OPPOSITION TO CURRENT CHANGE.

13-115. At the moment the switch, SW, in figure 13-29 is closed, the applied voltage, E, produces an electric field through the conductor, C; this field is an applied force, f_1, which urges the free electrons toward points of higher potential (upward in figure 13-29). The current tries to rise instantly to its maximum value of E/R. However, as the electrons start their net movement upward, their velocity is changing while increasing from zero. The elemental field of each electron, therefore, becomes distorted, as explained in connection with figure 13-27. Assume that two individual electrons, 1 and 2, are in the positions shown in part B of figure 13-29. When these electrons start upward under the influence of the applied voltage, their individual electric fields become distorted, so that the force in the distorted portion of the field of 1 is in a direction opposite that of the applied field. It is apparent that this distorted portion of the field about electron 1 is opposing the upward movement of electron 2. The field of every moving electron in the wire is similarly distorted; consequently, the upward movement of every electron is opposed. Thus, an opposing force is exerted, not only on electron 2, but at every point in the conductor. This opposing force is called the induced force, f_2, and it prevents the current from instantly reaching its maximum value (equal to E/R). After the current reaches its E/R value, the electron velocity is uniform and the elemental electron fields are no longer distorted. Under this condition, there is no induced force to oppose the electron flow, and the current is limited only by the resistance of the circuit. The opposing effect of inductance is present only when the current is changing, and the induced force is always in such a direction that it opposes the change. If the applied voltage, E, is reduced, the current must fall to a lower value. However, as explained in connection with figure 13-28, the distorted portions of the electron field are now reversed (see part C of figure 13-29). Therefore, the induced force and the applied force are in the same direction. Note that by now aiding the applied force the induced force is opposing the current change.

13-116. INDUCTIVE FORCES.

13-117. If only the induced force were present in the conductor shown in part B of figure 13-29, it would urge electrons from point X toward point Y, thus generating a potential difference between the conductor ends, with end X becoming positive and end Y becoming negative. The induced force shown in part C, however, would urge electrons from point Y toward point X, causing X to become negative and Y to become positive. These generated voltages, due to an induced force, are called induced voltages.

13-118. CAPACITANCE.

13-119. The amount of charge on a conductor is not only a function of physical characteristics, but it is also proportional to the applied potential. In electrostatics, the quantity of charge, Q, that can be stored on a set of parallel plates is expressed by the following equation:

$$Q = CV$$

where Q is the total charge on the plates, C is the capacitance, and V is the potential difference between the plates.

13-120. DEFINITION AND EXPLANATION.

13-121. Capacitance is the ratio of the quantity of electricity to the potential difference, or stated as a formula:

$$C = \frac{Q}{V}$$

13-122. Capacitance is a useful term because it indicates how much charge is stored on a conductor with a given potential. The farad is the unit of capacitance. A conductor has a capacitance of 1 farad when its potential is changed 1 volt by 1 coulomb of electricity; a capacitor has a capacitance of 1 farad when the potential difference between its plates is changed 1 volt by a transfer of 1 coulomb of electricity from one set of plates to the other. The capacitance of capacitors is generally much less than 1 farad; it is usually measured in microfarads (μf) or in micromicrofarads ($\mu\mu f$).

13-123. Since $C = \frac{Q}{V}$, a capacitor or conductor has a large capacitance if its potential is raised only a small amount by a large quantity of electricity. For example, if the transfer of one microcoulomb of electricity from one set of plates to the other in a capacitor changes its potential by 1 volt, then:

$$C = \frac{Q}{V} = \frac{1 \cdot 10^{-6}}{1} = 1 \cdot 10^{-6} \text{ farad} = 1 \ \mu f$$

If the similar transfer of four times as much electricity (that is, 4 microcoulombs) changes the potential of a second capacitor by 1 volt, the capacitance of the second capacitor is four times greater than that of the first, thus:

$$C = \frac{Q}{V} = \frac{4 \cdot 10^{-6}}{1} = 4 \cdot 10^{-6} \text{ farad} = 4 \ \mu f$$

13-124. EFFECT OF PLATE SEPARATION.

13-125. The formula for the quantity of electricity, Q, was previously discussed as:

$$Q = CV$$

Q may also be found by use of the following formula:

$$Q = \frac{KAV}{4\pi d}$$

where Q is the total charge on the plates, A is the area of the plates, K is the dielectric constant of the medium between the plates, d is the separation between the plates, 4π is a constant, and V is the potential difference between the plates. If the height, S, above ground is substituted in the formula for the distance, d, between capacitor plates, the revised formula still applies. The charge on each plate is proportional to the potential difference, V. Since 4π is a constant, if K and A are also held constant, the value of $Q = \frac{KAV}{4\pi S}$ can be determined for various values of the height above ground, S:

$$V = QS$$

Assume that a conductor has 1 unit of charge and that its separation (that is, distance) from the earth is 1 unit, then:

$$V = QS = 1 \cdot 1 = 1 \text{ volt}$$

The potential of the conductor will increase as S is increased. With S increased from 1 to 2 units, then:

$$V = QS = 1 \cdot 2 = 2 \text{ volts}$$

13-126. CHANGE OF C AS S IS VARIED.

13-127. Since V changes as S is changed, it follows that C must also vary with changes in S, because:

$$C = Q/V = \frac{Q}{QS} = \frac{1}{S}$$

Consider that a wire 2 inches long and 0.05 inch in diameter is parallel with, and 0.2 inch above, the ground. Under this condition, assume that the wire capacitance to ground is 1 $\mu\mu f$. A charge, Q, of 1×10^{-12} coulomb (1 micromicrocoulomb) changes the potential difference between the wire and ground by 1 volt, since:

$$C = Q/V$$

and

$$V = Q/C$$

$$= \frac{1 \times 10^{-12}}{1 \times 10^{-12}}$$

$$= 1 \text{ volt}$$

Assuming that the wire is at a positive potential with respect to ground and that the charge Q remains constant, the potential increases to 2 volts when S is doubled by raising the wire to a height of 0.4 inch. Although the wire retains the same quantity of electricity at the greater distance from ground, its capacitance is halved, since:

$$C = Q/V = \frac{1}{S}$$

$$= \frac{1 \times 10^{-12}}{2}$$

$$= 5 \times 10^{-13} \text{ farad}$$

$$= 0.5\ \mu\mu f$$

The graph in figure 13-30 shows that the variation of the capacitance, C, and the potential, V, as S is increased by raising the wire from 0.1 inch above the ground to a height of 20 inches. Note that the graph values representing C are not accurate values, since they are based on an assumed

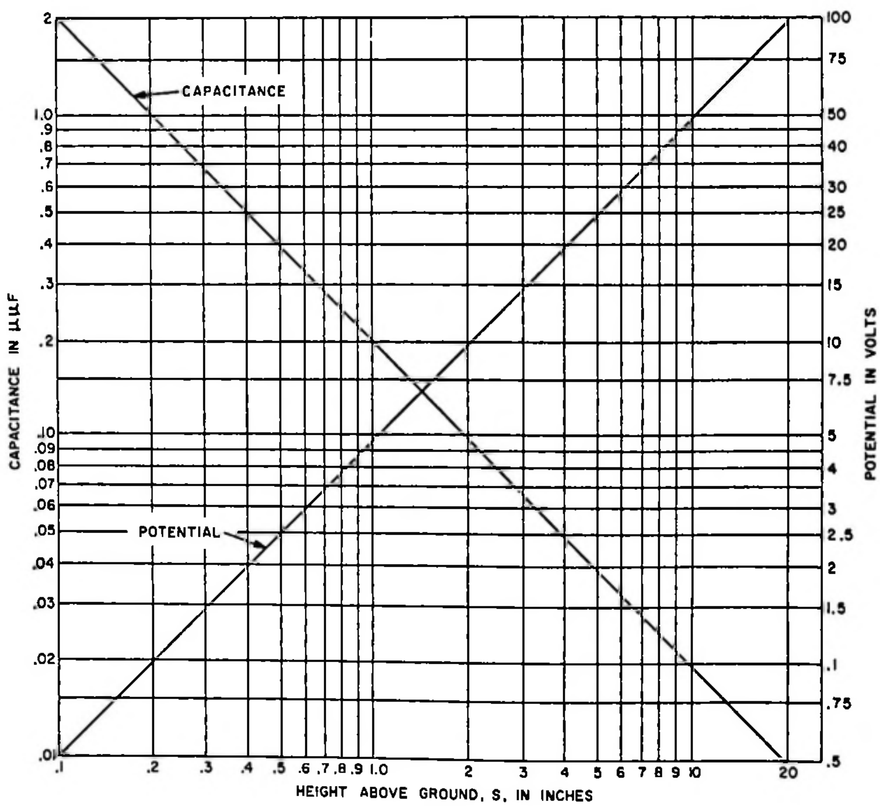


Figure 13-30. Graph Showing Variation of Capacitance and Voltage as Wire Is Raised Above Ground

value of 1 $\mu\mu$f at a separation S of 0.2 inch. The only purpose of the graph is to show in a general way the decrease in capacitance C as spacing S is increased.

13-128. INFINITY VERSUS ZERO.

13-129. It is reasonable to assume that as the separation, S, is increased to a distance greater than can be represented by any number, no matter how great the number, the potential, V, will also increase to a value which is too great to be represented by a number. A value which is so great that it is always greater than any number that can be written is said to be infinite; an infinite value is represented by the symbol ∞.

13-130. ZERO CAPACITANCE. Since the capacitance is inversely proportional to the potential difference, if V is made infinite by increasing S to an infinite value, the capacitance under this condition is said to be zero; that is:

$$C = Q/V = 1/\infty = 0$$

The concept that 1 (or any number) divided by an infinite value, that is, by ∞, is always equal to zero can be understood from an examination of figure 13-30. With increasing numerical values of S, C becomes smaller and smaller and approaches a value of zero. Consequently, if S is increased to a value greater than the largest number that can be written, it is reasonable to assume that C becomes zero in accordance with the mathematical concept that $1/\infty = 0$.

13-131. INFINITE CAPACITANCE. Of course, it is impossible to move a wire to an infinite distance. However, scientists have concluded that the velocity at which electromagnetic waves travel through free space is equal to the velocity of light; if we assume that a wire is in free space, its capacitance to ground is zero and the velocity at which an electrical disturbance travels along the wire is thought to be equal to the velocity of light in free space. The relationships of wire length, wavelength, and frequency are functions of the wire inductance, capacitance, and resistance.

SECTION III

ELECTRIC TRANSMISSION

13-132. APPLICATION OF BASIC CONCEPTS.

13-133. ANALYSIS OF LINE ACTION.

13-134. The action of any transmission line can be analyzed by means of mathematical equations. The advantage of mathematical analysis is, of course, that a single equation may explain a complicated circuit action precisely and quantitatively. Because this manual is intended primarily for those who have not had the opportunity to study mathematics, only a few of the most useful equations are included in the text.

13-135. SIMPLIFICATION.

13-136. In most transmission-line circuits, a number of electrical effects occur simultaneously to produce the resultant line action. Whenever possible, simplicity is obtained by first considering the line in terms of each electrical effect as though only that effect were present. For example, in explaining the movement of voltage from one end of a line to another, the voltage (that is, the potential) at a point is considered as being due to the excess charge existing at that point. The apparent movement of this excess charge from point to point is then explained by the effect of its field on other charges within the wire. In this manner, reflection and oscillation are first described without reference to capacitance, inductance, and resistance. Also, unless otherwise indicated, it is assumed in this section that the wires or lines are in free space; that is, they are so far removed from earth that their operation is not affected by anything external to them.

13-137. MOVEMENT OF POINT OF CHARGE.

13-138. The apparent movement of an excess of one kind of charge from point to point along a wire (hence, the movement of a point of charge or point of potential from end of the wire to the other) can be explained in simple terms.

13-139. THEORETICAL CONDITIONS.

13-140. Consider that point 0 in part A of figure 13-31 is the near end of a straight wire which extends to infinity; that is, the wire is so long that the far end can never be reached. Also consider that the wire is isolated in free space, so far from the earth that capacitance to ground can be neglected. Before an excess of one kind of charge is produced on its near end at 0, the wire is electrically neutral and every point on the wire is at the same potential. It is impossible to illustrate the distribution of the millions of opposite kinds of electric charge within the wire.

13-141. EQUIVALENT POSITIVE CHARGE.

13-142. Although the positive charges within the wire cannot be moved along the line, the equivalent of a movement of positive charge in one direction is brought about by the actual movement of negative charge in

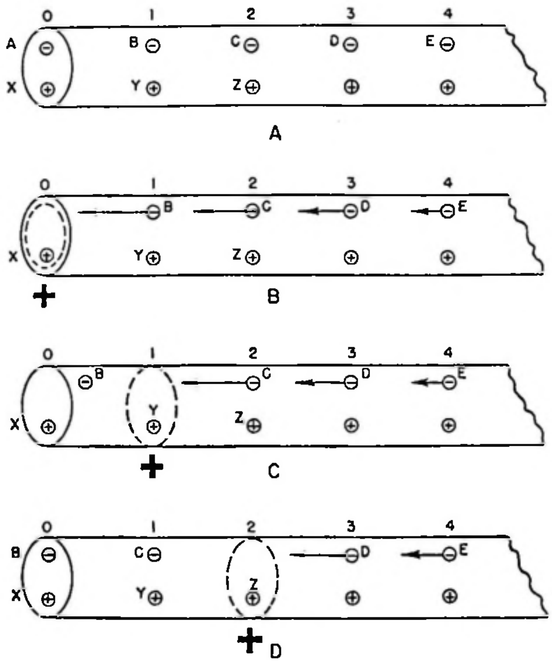


Figure 13-31. Simplified Concept of Movement of Point of Plus Charge

the opposite direction. Suppose that a portion of the normal negative charge (represented by electron A) is in some way moved from point 0 (see part A of figure 13-31). When this is done, as shown in part B of the figure, electrical balance no longer exists at point 0, because positive charge X (inclosed in dashed-line oval) forms an excess positive charge at point 0. In other words, the movement of electron A from point 0, and the consequent establishment of the positive charge, raise the potential of point 0. A potential difference now exists between point 0 and every other point in the wire.

13-143. FIELD DUE TO CHARGE.

13-144. Theoretically, the electric field of the positive charge established at point 0 extends in every direction from zero to infinity. At the moment the positive charge is established, the resultant electric field within the wire is strongest in the immediate vicinity of 0, and much weaker at points farther along the wire. For example, if the mutual attraction between the positive charge at point 0 and the electron designated as B at point 1 is 1 unit, the mutual attraction between an electron designated as C at point 2 (twice the distance 0-1 from zero) and the positive charge at point 0 may be less than 1/4 unit. Similarly, point 3 is three times the distance 0-1; therefore, the attractive force at point 3 may be less than 1/9 unit. This difference in force is roughly indicated by the relative lengths of the arrows on the electrons.

13-145. CURRENT IN VICINITY OF CHARGE.

13-146. Consider that when the positive charge is produced at point 0, the resulting field urges every electron in the wire toward that point (see part B of figure 13-31). However, because the greatest force is acting on the electrons which are nearest to the positive charge, the free electrons in this vicinity start moving toward point 0 at a greater velocity than the more distant electrons. This maximum current to the point of positive potential reduces the potential at point 0. Electrons between points 0 and 1 move away from point 1 much more rapidly than the more distant electrons move into point 1; therefore, point 1 (see part C of figure 13-31) now becomes positive because of its consequent momentary lack of electrons. As shown in part C of the figure, no positive charge has moved, but this current or movement of negative charge, designated as B, out of point 1 and into point 0 is the equivalent of moving the excess of positive charge (dashed-line oval) in the reverse direction (that is, from point 0 to point 1). Consequently, the potential of point 1 is now greater than the potential at any other point in the line.

13-147. TERMINOLOGY.

13-148. For simplicity, it has become customary to say that the charge or voltage moves down the line. It should be noted, however, that individual positive charges within the wire have not changed their positions; the movement of a group of electrons to the left constitutes a current that has produced an equivalent movement of the point of positive charge or potential in the reverse direction (to the right). The potential at point 1 thus becomes essentially the same as it would be if the original positive charge X had actually moved along the wire.

13-149. CONTINUED MOVEMENT OF CHARGE.

13-150. With the equivalent positive charge moved to point 1 (see part C of figure 13-31), the greatest electric force is now acting on all the electrons between points 1 and 2; therefore, the free electrons in this vicinity now produce the maximum current and move very rapidly toward the excess positive charge designated as Y at point 1. In moving away from the positive charge designated as Z at point 2, the current flows into point 1 to neutralize the positive charge there. At the same time, however, this current flows away from point 2 and the positive charge Z effectively becomes the equivalent positive charge. As indicated in part D of figure 13-31, the equivalent positive charge and, therefore, the point of positive voltage has already moved to point 2, and will continue to move toward the far end of the line. In this assumed case, the length of the line is infinite and the positive charge could move down the line forever without reaching the far end.

13-151. SPEED OF TRAVEL.

13-152. A further examination of figure 13-31 will show that no individual electron has traveled any great distance from its original position on the line. However, the equivalent positive charge—the point at which there is an excess of protons, and thus the point of electrical unbalance—travels down the line at the speed of light (186,000 miles per second). Note that the current in the wire is always maximum in the immediate vicinity of the equivalent positive charge as it moves along the line; that is, the voltage and current sweep down the line together and in phase. In an actual wire, the velocity is somewhat less than the speed of light in free space, because of the resistance and shunt conductance present in the wire. (The effects of these electrical constants will be considered later).

13-153. MOVEMENT OF POINT OF NEGATIVE POTENTIAL.

13-154. FIELD DUE TO CHARGE. The electrical unbalance caused by an excess of negative charge at point 0 will also travel down a straight wire in much the same manner as the equivalent positive charge. In part A of figure 13-32, the near end of the infinite line is again shown in its neutral condition. In part B of the figure, the near end of the line contains an excess negative charge (represented by electron A^1). Since point 0 now has an excess number of electrons, its potential is lower than that of every other point along the wire, and there is a potential difference or voltage between point 0 and every other point along the line. The field of the negative charge thus established at point 0 extends in every direction to infinity. The direction of the electric force of this negative charge is the reverse of that of the positive charge previously established at this end of the line (see part B of figure 13-31). This means that the electric field due to the excess negative charge at point 0 is forcing every free electron away from the near end of the line. This force is greatest on the electrons in the immediate vicinity of the negative charge therefore, maximum current occurs at this

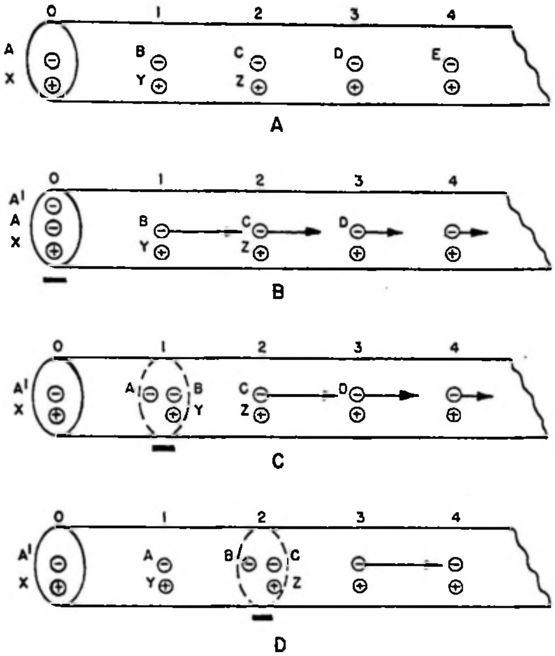


Figure 13-32. Simplified Concept of Movement of Point of Negative Charge

point, since the nearest electrons move very rapidly away from point 0 toward point 1.

13-155. MOVEMENT OF CHARGE. In this case, a negative charge can and does move along the wire from the positive charge, designated as X, toward point 1, to produce a balanced electric condition at the near end of the line. At the same time, the flow of current into point 1 produces an excess of negative charge at this point (see part C of figure 13-31). In turn, this increased negative charge at point 1 causes the free electrons between points 1 and 2 to start moving rapidly toward point 2, and the point of electrical unbalance (inclosed in dashed lines in figure 13-32) continues to move down the line in the manner described (see parts C and D of figure 13-32). It should again be noted that, although it is said that the negative charges move down the line with the speed of light, the actual displacement (that is, distance of travel) of any individual electron is very slight.

13-156. SUMMARY.

13-157. Whenever an excess of one kind of charge exists at a point, there is an electric potential at that point. Consequently, when the equivalent positive charge is first established on the wire (see part B of figure 13-31), there is a potential difference or voltage between point 0, which contains an excess number of positive charges, and the remainder of the line (which, for the moment, contains equal numbers of positive and negative charges). As the negative charges in the line shift successively to the left to neutralize the positive charge, the point of electrical unbalance or positive potential moves down the line with the speed of light. As the point of potential moves along the line, the current is always greatest in its immediate vicinity; thus, the voltage and current are said to travel in phase. Comparison of figures 13-31 and 13-32 will show that the movement of a point of negative charge down the line is similar to that of the point of positive charge, except that the current due to the charge is in the reverse direction.

13-158. REFLECTION OF CHARGE.

13-159. Before studying the effects of the inductance, capacitance, and resistance distributed along the length of an actual wire, you should first consider the movement of a charge along a wire of finite length. If an electric charge (whether positive or negative) is established at either end of a relatively short uncharged line, the point of potential will move toward the opposite end of the wire.

13-160. POSITIVE CHARGE.

13-161. As shown in part A of figure 13-33, if a positive charge is first established at the left end of the wire, it will move to the

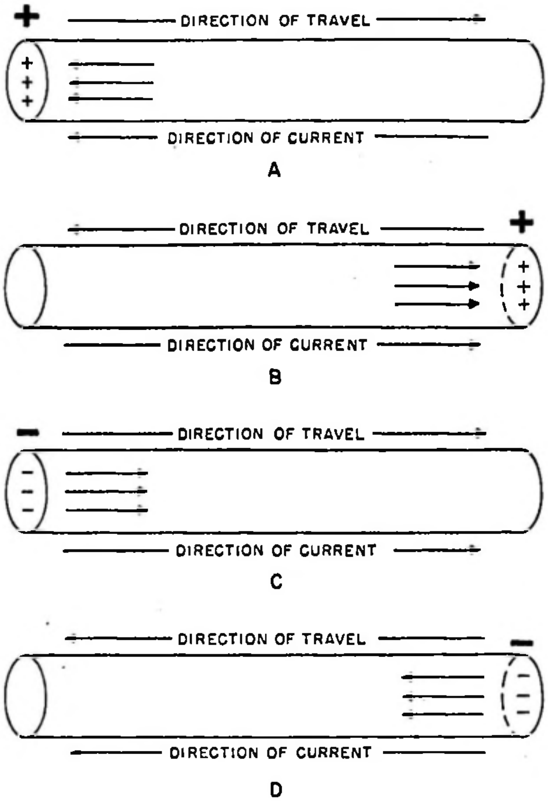


Figure 13-33. Movement of Points of Charge on Wires of Finite Length

right. However, if the positive charge is first established at the right end of the wire, it will move in the opposite direction (that is, to the left), as indicated in part B of the figure. As previously explained, the equivalent of moving a positive charge along a metallic conductor is actually accomplished by moving electrons (current) in the opposite direction. Since the equivalent positive charge shown in part A of the figure is moving in a direction opposite that shown in part B, the current in part A is also the reverse of that in part B. (Arrows are shown within the wires to indicate the direction and approximate position of maximum current.)

13-162. NEGATIVE CHARGE.

13-163. The negative charge first established at the left end of the wire in part C of figure 13-33 moves to the right, and the equal negative charge first established at the right in part D of the figure moves to the left. Since the established charges in parts C and D are negative, the current is in the same direction as the movement of charge or potential.

13-164. OSCILLATION.

13-165. If you compare part A with part B in figure 13-33, it is reasonable to assume that when the positive charge first established at the left in part A reaches the right end of the line, the electrical condition of the line will be essentially the same as that of the line shown in part B of the figure. Under this condition, the positive charge will reverse its direction of travel and move back over the line toward its original position at the left. Following this line of reasoning, it is also reasonable to assume that upon reaching its original position at the left, the positive charge will again reverse its direction of travel and move toward the right end of the line. Assuming that none of its original energy is radiated from the wire or lost within the wire, the positive charge shown in part A, or the negative charge shown in parts C and D, will continue to move back and forth, or oscillate, along the wire forever.

13-166. COMPLETE REFLECTION.

13-167. In this case, the wire is assumed to be a good conductor and the medium surrounding the wire is assumed to be a perfect nonconductor, or insulator. Consequently, as the positive or negative electric charge reaches either end of the wire, it encounters a perfect insulating medium through which it cannot move, and therefore must stop. Since it is assumed that the stopped charge has not lost any energy, its electric field exerts a force on every electron within the wire. This force of the electric field on the electrons now causes a reversal of current within the wire and a consequent movement of the charge back over the wire. Because the reversal of its direction of travel is comparable to the reflection of light from the surface of a mirror, the charge is said to be reflected from the end of the line. The above-described reflection is said to be complete reflection, for all of the energy reaching the end of the line is reflected back over the line; that is, no energy is dissipated in the medium at the end of the line. Whenever the terminating medium absorbs all the energy reaching it from the line, no reflection can take place. In some cases, only a portion of the energy may be absorbed in the terminating medium and the remainder of the energy is reflected back over the line. The various line terminations and their effects on reflection will be discussed in detail later.

13-168. POLARITY OF REFLECTED CHARGE.

13-169. A further examination of figure 13-33 will bring out a very important action which you must thoroughly understand. The equivalent positive charge represents a point of positive potential, or voltage, which moves from the left to the right in part A of figure 13-33. Upon being reflected, however, the point moves from the right to the left, as shown in part B. Note that although this potential reverses its direction of travel, there is no reversal of polarity. In other words, regardless of its direction of travel, the charge is the equivalent of a point of higher potential which is moving back and forth over the line.

13-170. REVERSAL OF CURRENT WITHOUT REVERSAL OF POLARITY.

13-171. If you compare the current flow in part A of figure 13-33 with that of part B, you will see that there is a reversal of current when the excess charge or point of potential is reflected from either end of the line. This reflection of a charge (or potential) without reversal of polarity, but with a reversal of current flow, occurs whenever the line is terminated in a high impedance that is the equivalent of a high resistance. The negative charge (or potential) established at the left in part C of figure 13-33 travels to the right end of the line and then reverses its direction of travel as shown in part D, but in traveling back over the line the negative potential does not reverse its polarity. In this case also, there is a reflection of potential without a reversal of polarity, but with a reversal of current.

13-172. CAPACITANCE OF A WIRE.

13-173. The capacitance of a wire consists of the capacitance between the wire and ground and other objects, and also the capacitance of the wire itself, independent of ground and all other bodies.

13-174. SELF-CAPACITANCE.

13-175. Theoretically, if a wire is moved to an infinite distance from the earth, the capacitance of the wire to ground and to other objects becomes zero. If it were possible to study a wire under these theoretical conditions, it would be found that a charge could be taken from or added to the wire; that is, the wire has self-capacitance. This concept is at first somewhat difficult to understand, because it cannot be visualized directly in terms of the familiar two-plate capacitor.

13-176. SIMPLIFIED EXPLANATION.

13-177. Consider that the wire illustrated in part A of figure 13-34 is isolated in free space and that oscillations can be stopped at the instant the opposite ends contain maximum charge. At this instant, slice end sections 1 and 2 from the wire and place them as indicated in part B of figure 13-34. These end sections and the dielectric mate-

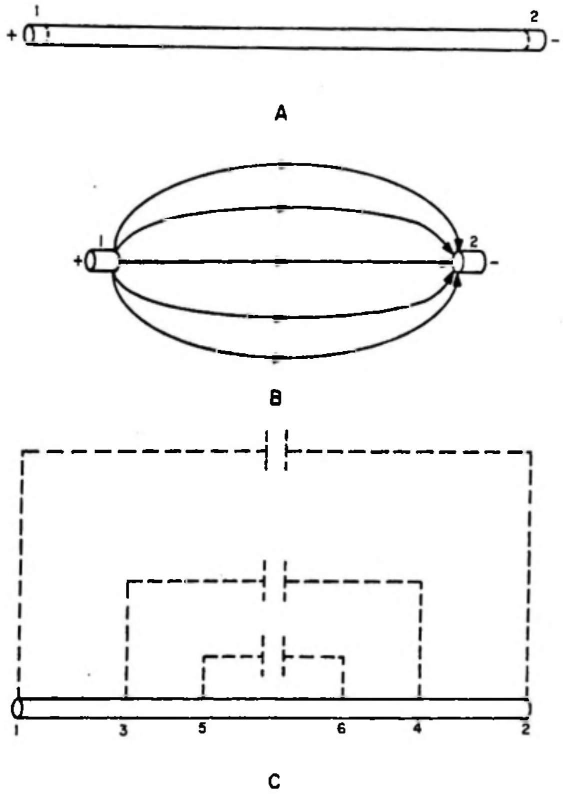


Figure 13-34. Self-capacitance of a Wire

rial about the wire form a very small two-plate capacitor; see points 1 and 2 in part C of figure 13-34. This end capacitance is only a portion of the self-capacitance; there is also capacitance between each section and every other section in the wire. This idea of distributed self-capacitance is shown, within the limitations of a simple drawing, by the dotted-line capacitors between points 3 and 4, and 5 and 6 in part C of figure 13-34.

13-178. INDUCTANCE OF A STRAIGHT CONDUCTOR.

13-179. Every conductor, regardless of its size and shape, possesses self-inductance. This self-inductance may be explained either in terms of the magnetic field or in terms of the electric field. These explanations are equivalent and should cause no confusion if you remember that the magnetic field is an attribute of the electric field. Explanations in terms of the electric field are sometimes more direct. The important point to remember is that every conductor has self-inductance. Thus, the shortest possible piece of wire has self-inductance. Although the amount of self-inductance in a short wire may be extremely small, its effect may be considerable at high frequencies.

13-180. RESONANT WIRE.

13-181. The movement of a point of potential or charge along a wire and reflection from one end to the other have been considered without reference to the inductance, capacitance, or resistance of the wire. These simplified explanations show that a point of charge (or potential) necessarily moves because of the resultant electric field and its effect on charges in the wire. However, the relationship of wire length, frequency, and wavelength can be explained only by including the concept of inductance and capacitance in the discussion.

13-182. COMPARISON OF WIRE AND LC CIRCUIT.

13-183. It is possible to compare a length of wire directly with a coil and capacitor circuit. To show this practical equivalence, assume that the length of wire in figure 13-35 has points of charge that are moving back and forth over the wire. Now consider the total distributed capacitance of the wire as concentrated in the capacitor formed by the end sections, as shown in part B of figure 13-34. This capacitance corresponds to the lumped capacitance, C, in figure 13-35. With a lumped inductance, L, equal to the distributed inductance of the wire, the LC circuit of figure 13-35 can be considered as a practical equivalent of the wire.

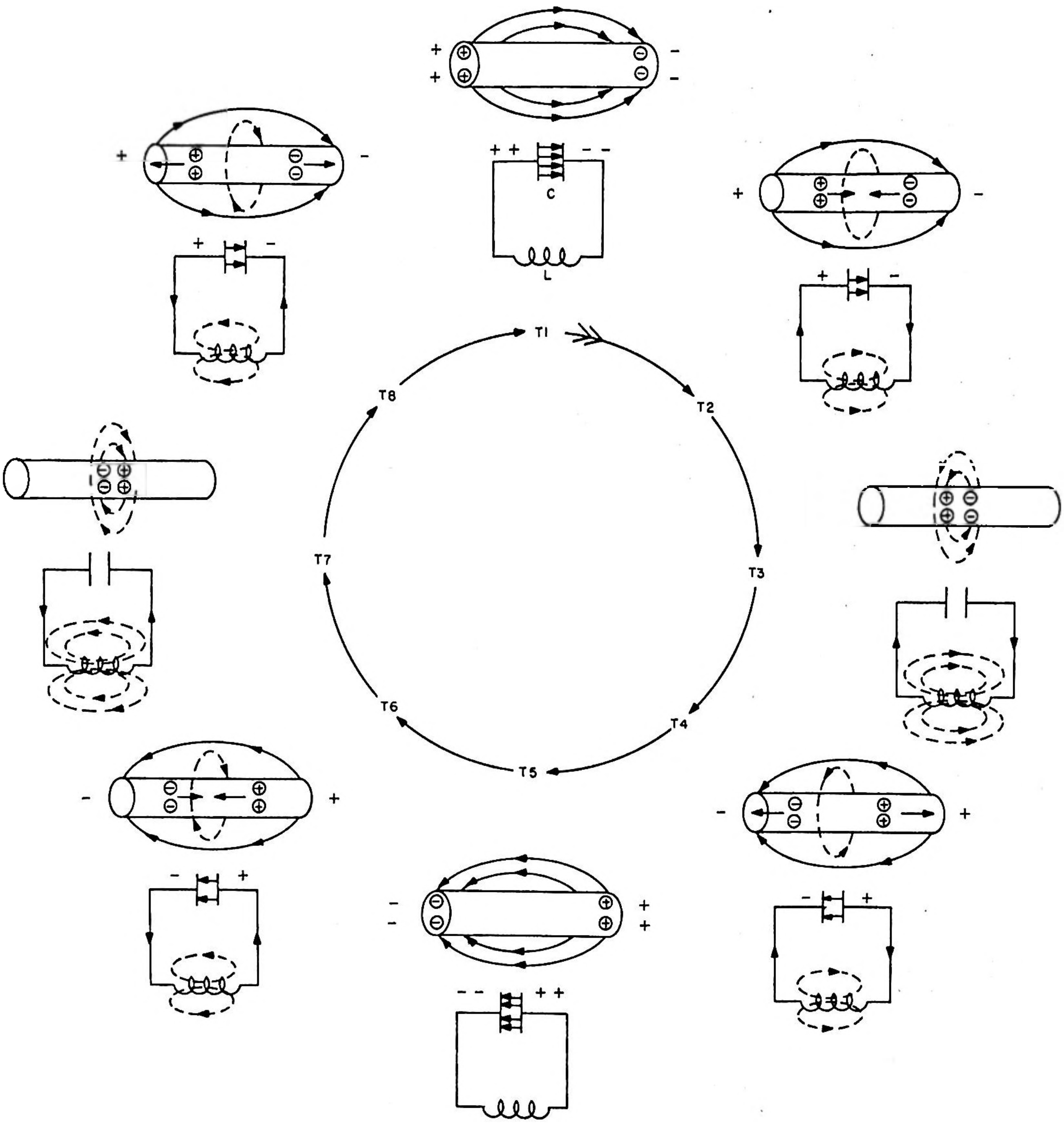


Figure 13-35. Oscillations in a Wire Versus Oscillations in an LC Circuit

13-184. CIRCUIT ACTION.

13-185. Refer to figure 13-35, where the electrical conditions of a wire and its corresponding LC circuits are represented for one cycle of oscillation. Times are given by the time circle T_1 through T_8 (clockwise). The electric and magnetic fields are represented by conventional lines of force. In the wire the point of maximum positive charge is represented by circled plus signs (⊕⊕), and the point of maximum negative charge is represented by circled minus signs (⊖⊖). The short arrows indicate the directions in which these points of charge are moving. The direction of current in the LC circuit is indicated by arrowheads on the connecting wires. The potentials of the wire ends and of the capacitor plates are represented by conventional plus and minus signs.

a. At time T_1 (see figure 13-35), the maximum potential difference exists between the ends of the wire, and the points of charge are also at the ends of the wire. At this instant the distributed capacitance of the wire and the capacitor in the LC circuit are fully charged. Thus, in both circuits all of the energy is in the electric field. At this instant there is no current and, hence, no magnetic field about either circuit.

b. At time T_2 the points of charge are moving from the ends of the wire and are closer together. While the points of charge are in motion, there is current in the wire; consequently, there is a magnetic field about the wire. Note that the potential difference between the ends of the wire is decreasing, because the points of charge are moving closer together. At this instant the voltage across the capacitor in the LC circuit is also decreasing, because the capacitor is discharging through the inductance. This discharge current produces a magnetic field about the inductance, L.

c. At time T_3 the points of charge are together (superimposed) and there is no potential difference between the ends of the wire. At this same instant, the capacitor in the LC circuit is fully discharged, the capacitor voltage is zero, current through the inductance is maximum, and all of the circuit energy is in the magnetic field.

d. At time T_4 the points of charge have passed one another in the wire, and, because these points are no longer superimposed, there is again a potential difference between the wire ends. However, the points of charge have reversed their positions and there is a consequent reversal of polarity. Compare T_4 with T_1 and T_2. Note that although the magnetic field is collapsing, its direction has not reversed. In the LC circuit the magnetic field is also collapsing without a reversal of direction. This collapsing field induces a voltage which causes current to flow in the original direction; consequently, capacitor C is now charging with reversed polarity.

e. At time T_5 the points of charge have reached the ends of the wire, there is maximum potential difference between the wire ends, the current is zero, and the magnetic field is zero. All of the energy is again in the electric field. Conditions in the LC circuit correspond to those in the wire; that is, the capacitor is fully charged, the capacitor voltage is maximum, the current is zero, the magnetic field is zero, and all of the circuit energy is in the electric field.

f. As explained for times T_1 through T_5, the circuit action continues through times T_6, T_7, and T_8, back to T_1, except that the points of charge move in the reverse direction and capacitor C discharges in the opposite direction through inductance L. One cycle of oscillation is completed when the original conditions of T_1 are re-established.

13-186. FREQUENCY, WAVELENGTH, AND WIRE LENGTH.

13-187. Resonant circuit theory shows that the frequency of a resonant LC circuit is a function of L and C. Comparison of a straight wire and the LC circuit of figure 13-35 indicates that the frequency of oscillation in a wire is also a function of L and C. In a wire the inductance and capacitance are distributed, instead of being lumped as in the familiar coil and capacitor circuit.

13-188. EQUATION FOR RESONANT FREQUENCY, F_r.

13-189. The equation for resonant frequency in terms of L and C is:

$$F_r = 1/2\pi\sqrt{LC}$$

where:

L is in henrys

C is in farads

F_r is in cycles

13-190. VARIATION OF L, C, AND F_r WITH WIRE LENGTH.

13-191. With everything else equal, the distributed inductance and capacitance of a given wire are proportional to the wire length. Also, for a given frequency any combination of L and C that will give the required LC product (L x C) may be used to form a resonant circuit at that frequency. Now, if the wire length is doubled, the inductance and capacitance are both doubled and the LC product is four times as great; if the wire length is halved, the inductance and capacitance are both halved and the LC product is one-fourth as great. However, F_r varies inversely as the square root of the LC product. Therefore, if the wire length is doubled, the frequency is halved; if the wire length is halved, the frequency is doubled.

13-192. EXAMPLES.

13-193. When L is in μh (microhenrys) and C is in μf (microfarads), the required LC products for a frequency of 750 kc (kilocycles) is .0450. Assuming that a .0005-μf capacitor is to be used:

$$L = LC/C$$

$$L = .0450/.005 = \frac{45 \times 10^{-3}}{5 \times 10^{-4}} = 90\ \mu h$$

If L and C are both doubled, as is the case when the wire length is doubled:

$$L = 2 \times 90 = 180\ \mu h$$

$$C = 2 \times .0005 = .001\ \mu f$$

and the new LC product is:

$$LC = 180 \times .001 = .180$$

The LC products of .180 will correspond to a frequency of 375 kc. This is one-half the original frequency of 750 kc. If L and C are both halved, as is the case when a wire length is halved:

$$L = 90/2 = 45\ \mu h$$

$$C = .005/2 = .00025\ \mu f$$

and the new LC product will be:

$$LC = 45 \times 25 \times 10^{-5} = 1125 \times 10 = .01125$$

The LC product of .00125 will correspond to a frequency of 1500 kc. This is twice the original frequency of 750 kc.

13-194. VARIATION OF WAVELENGTH WITH WIRE LENGTH.

13-195. As previously discussed, there is a definite relationship between wire length and frequency. Since a definite relationship also exists between frequency and the wavelength of an electromagnetic wave, there must be a relationship among the three factors, wavelength, frequency, and wire length. Consider the wavelength equation:

$$\text{Wavelength} = \frac{\text{velocity}}{\text{frequency}}$$

where:

Wavelength, λ, is in meters
Velocity, V, is the speed of light (approximately 300,000,000 meters/sec)
Frequency F, is in cycles

When wavelength is given in meters and frequency is given in megacycles, this equation becomes:

$$\lambda = 300/F_{mc}$$

Referring to the frequencies discussed in the examples above, when F is 750 kc (.75 mc):

$$\lambda = 300/.75 = 400 \text{ meters}$$

For an F of 375 kc (.375 mc):

$$\lambda = 300/.375 = 800 \text{ meters}$$

For an F of 1,500 kc (1.5 mc):

$$\lambda = 300/1.5 = 200 \text{ meters}$$

13-196. NATURAL FREQUENCY.

13-197. From the foregoing equations it might be thought that for self-oscillations at a frequency of 1500 kc a wire would have to be 200 meters in length. Note, however, that to complete one cycle of oscillation, each point of charge travels the wire length twice (see figure 13-35), once from its original position to the opposite end of the wire, and once more in returning to its original position. Consequently, self-oscillations occur at a frequency which is such that the wire length is one-half the wavelength determined from the basic wavelength formula. The frequency at which self-oscillations occur is called the natural frequency or the natural resonant frequency. The time required for one cycle of oscillation at the natural frequency is called the natural period of the wire.

13-198. CONVENIENT LENGTH EQUATIONS.

13-199. It is sometimes convenient to express wavelength in feet. One meter equals 39.37 inches, or 3.28 feet. Thus, when the frequency is in megacycles, the wavelength, λ, in feet is:

$$\lambda = \frac{3.28V}{F \text{ in mc}}$$

$$= \frac{3.28 \times 300 \times 10^6}{F \text{ in mc}}$$

$$= \frac{984}{F \text{ in mc}}$$

For the half-wavelength ($\lambda/2$), this equation becomes:

$$\lambda/2 = \frac{492}{F \text{ in mc}}$$

These equations give the length in feet when it is assumed that the wave is traveling in free space. In practical work a correction factor must be applied.

13-200. ELECTRICAL VERSUS PHYSICAL LENGTH.

13-201. There are two ways of designating the length of a line:

a. Physical length. This is the length of a line measured in length units such as inches, feet, yards, meters, miles, etc.

b. Electrical length. This is the length of a line measured in terms of wavelength at the frequency normally used on the line.

13-202. LONG AND SHORT LINES.

13-203. Power, telegraph, and telephone lines may be several hundred miles in length. Such lines are long in terms of distance, but are usually short in terms of wavelength. For example, 60 cycles is a commonly used frequency for power transmission. At 60 cycles one wavelength, λ, in meters is:

$$\lambda = \frac{3 \times 10^8}{F \text{ in cycles}} = \frac{3 \times 10^8}{60} = 5 \times 10^6 \text{ meters}$$

and the length in miles is:

$$\lambda = \frac{\text{meters} \times (3.28 \text{ ft/meter})}{5280 \text{ ft}}$$

$$= \frac{5 \times 10^6 \times 3.28}{528 \times 10} = 3100 \text{ miles}$$

Thus, a 100-mile line is only $\frac{100}{3100}$, or 1/31, wavelength at a frequency of 60 cycles. Such a line is long in terms of distance, but short in terms of wavelength. Now consider a line which is 328 feet in length and normally used in 30-megacycle circuits. In terms of distance, this line is extremely short when compared with the 100-mile power line discussed in the previous paragraph. However, one wavelength at an operating frequency of 30 mc is:

$$\lambda = \frac{300}{F \text{ in mc}} = \frac{300}{30} = 10 \text{ meters}$$

The 328-foot line is 328/3.28 = 100 meters in length; therefore, its electrical length is 100/10 = 10 wavelengths. In other words, a 328-foot line used at 30 megacycles is electrically longer than a 100-mile line used at 60 cycles.

13-204. VOLTAGE AND CURRENT DISTRIBUTION.

13-205. Transmission lines cannot be considered as simple series circuits. In series circuits the current is everywhere the same, as indicated by ammeters A, B, and C in figure 13-36. The current is not everywhere the same in the conductors of a transmission line. The constants R, L, C, and G are spread out or distributed along a wire, as was shown in part D of figure 13-6; they are not concentrated as they are in the simple series circuit shown in part A of figure 13-6.

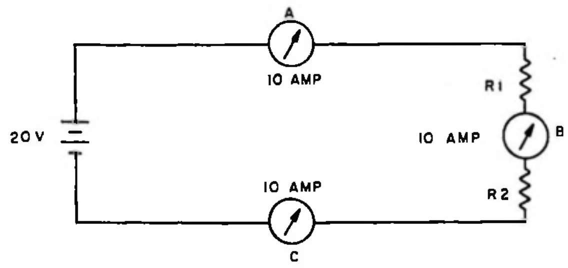


Figure 13-36. Current Everywhere the Same in a Series Circuit

The effects of distributed constants differ from those obtained with lumped constants. Consequently, variations of current and voltage must be determined at each point along a transmission line.

13-206. INSTANTANEOUS VOLTAGE AND CURRENT.

13-207. Although generally considered as an antenna, the resonant half-wave wire provides an excellent example of the effects of distributed constants. Assume that oscillations are occurring in a wire of length A to B as shown in figure 13-37. Also assume that the action can be stopped at any one of times T_1 through T_{17}, so that the instantaneous voltages and currents can be measured at each point along the wire. Voltage and current distribution curves can then be drawn for each time, T, represented.

a. Part A of figure 13-37 shows a sequence beginning at time T_1 where the points of maximum charge are at opposite ends of the wire, indicating that the distributed capacitance is fully charged. (This instant corresponds to time T_1 in figure 13-35). The voltage distribution curve is sinusoidal, with maximum potentials at opposite ends of the wire and zero potential at the center, C. At this instant, there is no current at any point in the wire.

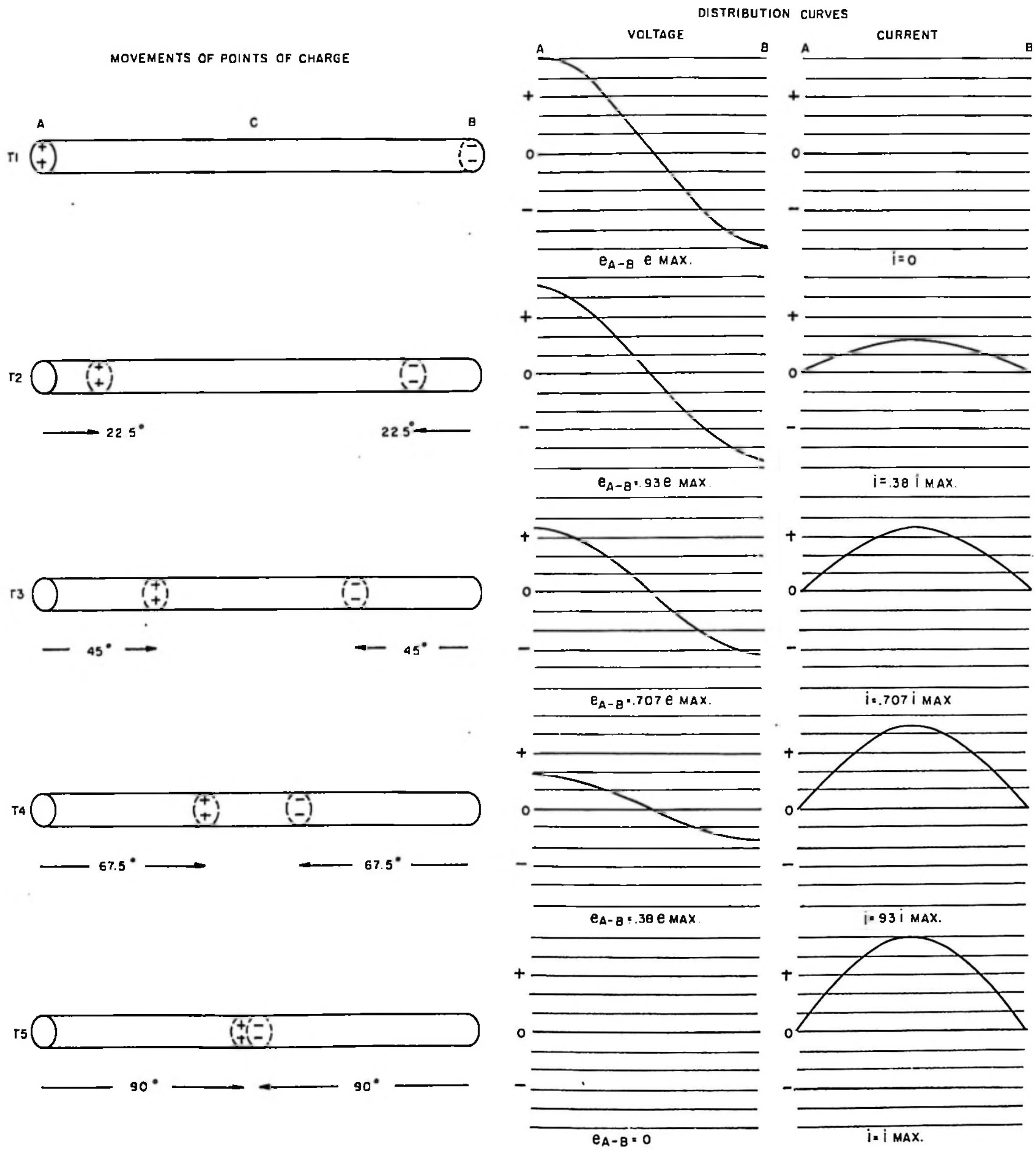


Figure 13-37. Voltage and Current Distribution (Part A)

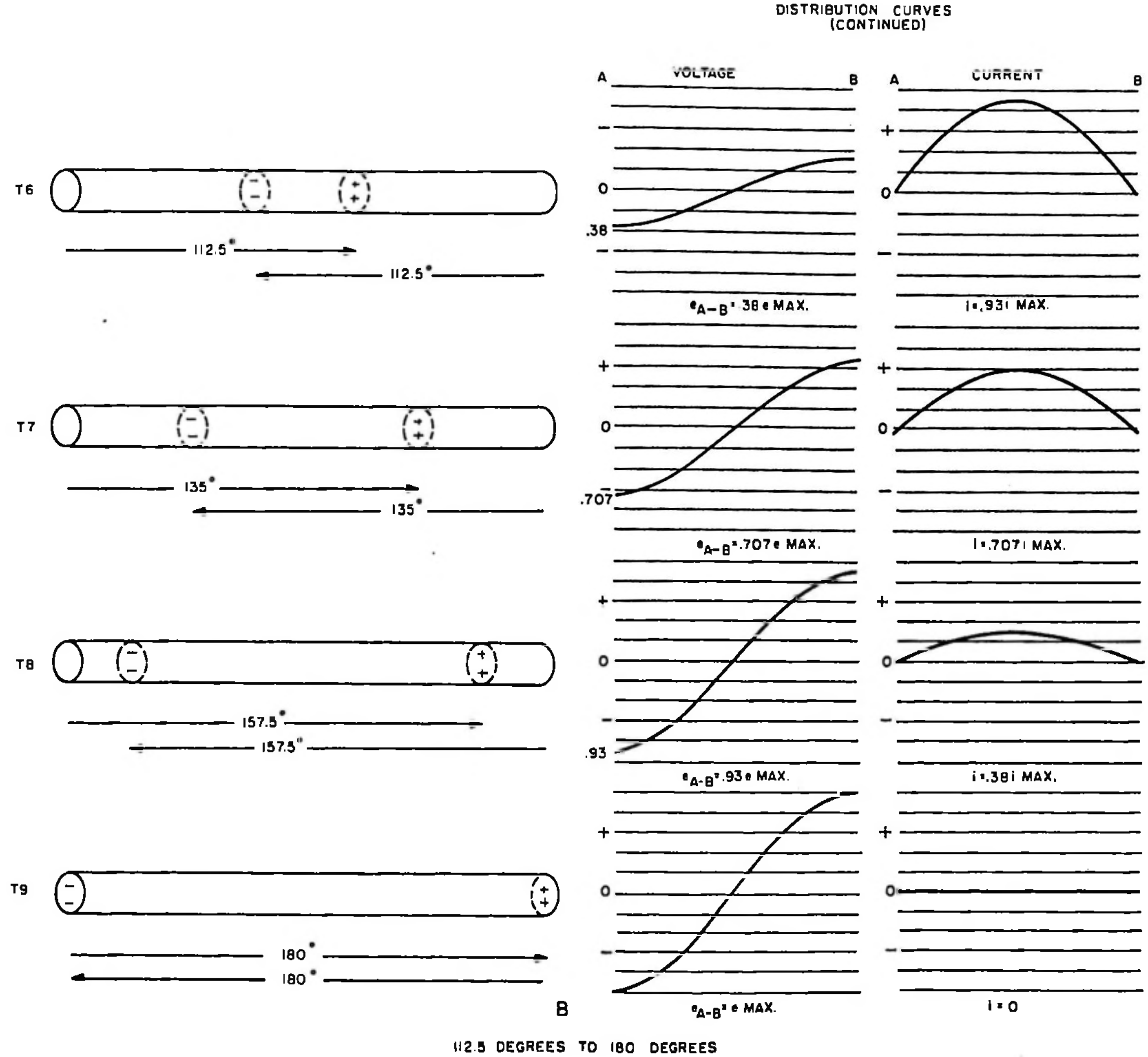


Figure 13-37. Voltage and Current Distribution (Part B)

b. The positive charge at point A and the negative charge at part B are reflected toward the opposite ends of the wire. At time T_2 each point of charge has moved 22.5 degrees from its original position at time T_1. (The arrows and degree markings below the wire indicate the direction and distance (in degrees) that the corresponding point of charge has moved with reference to its original position at time T_1.) The voltage distribution curve for this instant shows that the voltage is falling; the current distribution curve shows that the current is rising. Note that the current distribution is also sinusoidal, and that the maximum current is at the center of the wire.

c. Through times T_2 to T_5 the voltage decrease is in direct proportion to the current increase. Zero voltage and maximum current occur at time T_5.

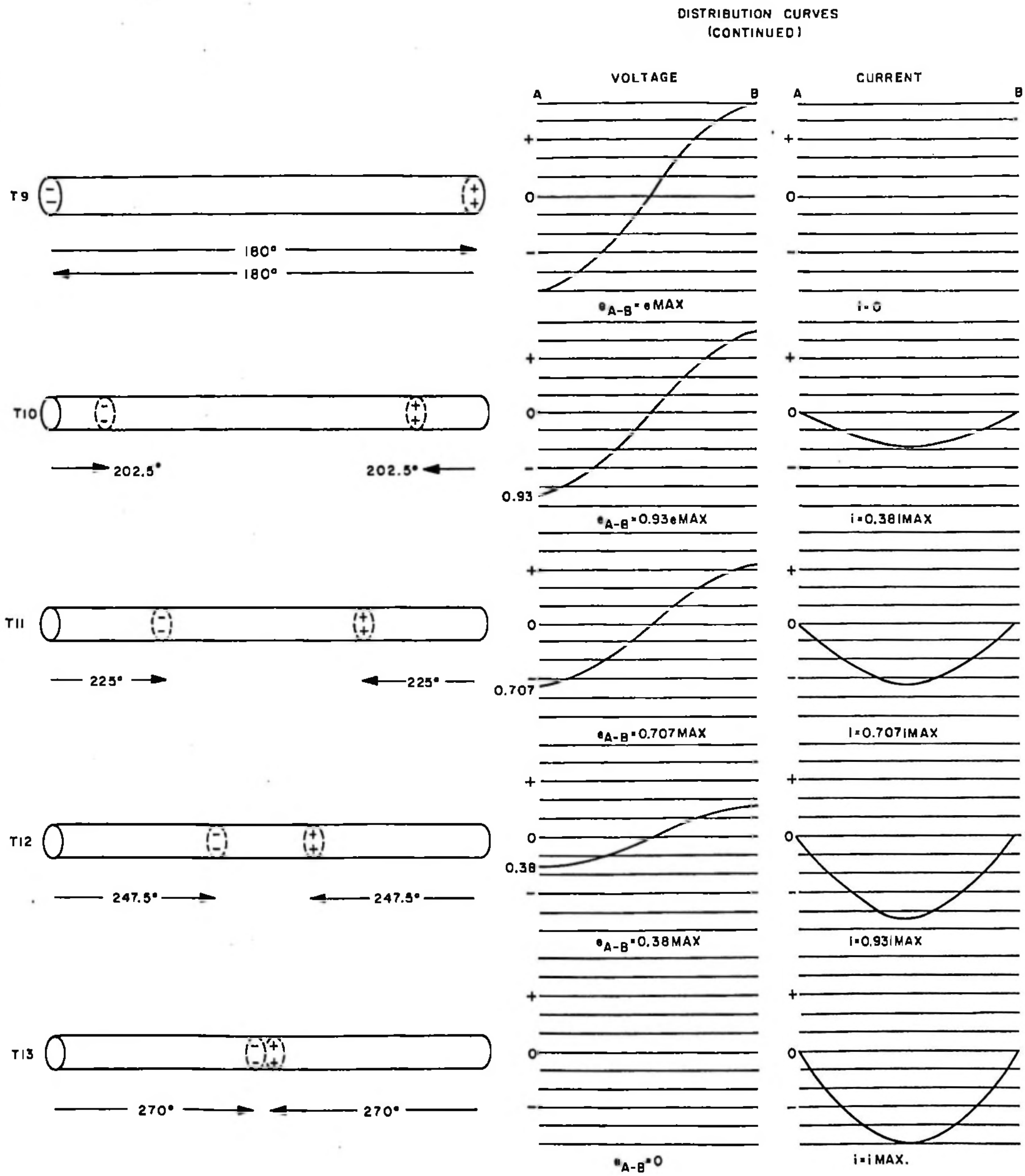


Figure 13-37. Voltage and Current Distribution (Part C)

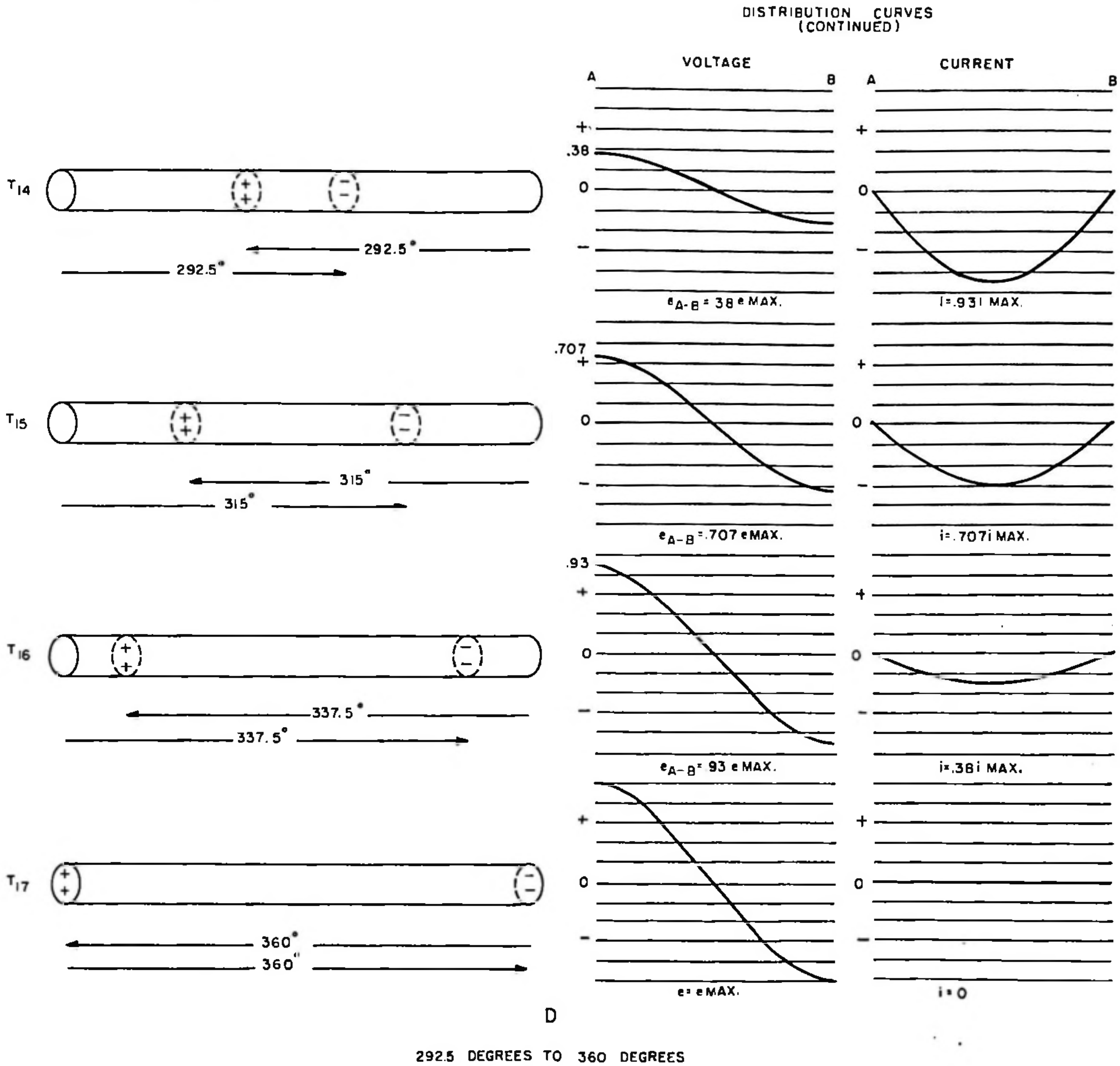


Figure 13-37. Voltage and Current Distribution (Part D)

d. After passing through zero voltage at time T_5 (see part B of figure 13-37), the voltage reverses polarity and begins to rise, reaching its maximum value at time T_9. During this time the current decreases from maximum to zero, but does not reverse. (Note that this action through time T_9 corresponds to times T_1 through T_5 in figure 13-35). The sinusoidal distribution shown in figure 13-37 is due to the distributed inductance and capacitance of the wire. Notice that neither the current nor the voltage is everywhere the same along the wire.

e. At time T_9 (repeated in part C of figure 13-37), the maximum negative charge i at point A and the point of maximum positiv charge is at point B. Reflection again take place and each point of charge moves back over the wire toward its original position a time T_1. Therefore, the voltage now decreases and the current reverses, rising from zero at time T_9 to maximum negative at time T_{13}. From time T_{13} (see part D of figure 13-37), as the current decreases to zero at time T_{17}, the voltage reverses and rises to maximum at time T_{17}.

13-208. TRAVELING WAVES.

13-209. INDIVIDUAL WAVES. At time T1 (see figure 13-37), the point of maximum positive charge is at point A, and the point of maximum negative charge is at point B. If the point of positive charge is considered by itself, it produces at this instant a certain distribution of charge and a corresponding voltage wave, as shown by the dashed-line waveform in part A of figure 13-38. Similarly, considered by itself, the point of negative charge produces a different distribution of charge and, hence, a different voltage wave, as shown by the dash-dot waveform in part A of figure 13-38. Since the points of charge travel along the wire, these two voltage waves also travel over the wire, and at any instant the resultant voltage wave is the sum of these two traveling waves. Assume that the concentration of positive charge at point A makes this point 10 volts positive with respect to the center of the wire. Also, the voltage wave due to the negative charge at point B makes point A an additional 10 volts positive. The resultant potential at point A is the algebraic sum of these two potentials, or 20 volts. By similarly adding the potentials due to the various individual waves, the resultant voltage wave is shown by the solid-line waveform in part A of figure 13-38. At time T_3 in figure 13-37, the points of maximum charge have been reflected and have traveled 45 degrees from the ends of the wire. The individual waves have also traveled 45 degrees along the wire, and are positioned as shown in part B of figure 13-38. Addition of the individual waves at this instant produces a resultant voltage wave of 14 volts peak amplitude. When the points of maximum charge are passing each other, as shown at time T_5 in figure 13-37, the individual voltage waves are exactly 180 degrees out of phase, as shown in part C of figure 13-38. There is, therefore, no resultant voltage on the wire at this instant.

13-210. TRAVELING CURRENT WAVES. Individual and resultant current waves for times T_1, T_3, and T_5 in figure 13-37 are shown in parts D, E, and F, respectively, of figure 13-38.

13-211. SUMMARY.

13-212. The current and voltage in a transmission line must be determined at each point along the line. The current is not everywhere the same as in a simple series circuit. Both the current and voltage are constantly changing at every point on a line. On any line in which reflections occur, the resultant instantaneous voltage, or current, at any point is the algebraic sum of two instantaneous voltages, or currents, at that point. These instantaneous voltages, or currents, are due to two individual waves which travel in opposite directions along the line.

13-213. STANDING WAVES.

13-214. Probably the most important result of reflection is the production of standing waves. Reflection produces two elemental waves that travel in opposite directions over the wire. Because of this phenomenon, the resultant voltage and current distribution differs from that when there is no reflection.

13-215. VOLTAGE STANDING WAVE (VSW).

13-216. As shown in part A of figure 13-39, a few of the instantaneous voltage distribution curves for 1/2 cycle of oscillation in a wire of length A to B are drawn on the same graph. This graph indicates how the voltage at each point varies. Thus, assuming a peak voltage of 10 volts and reading downward along the time axis, the voltage at point A is

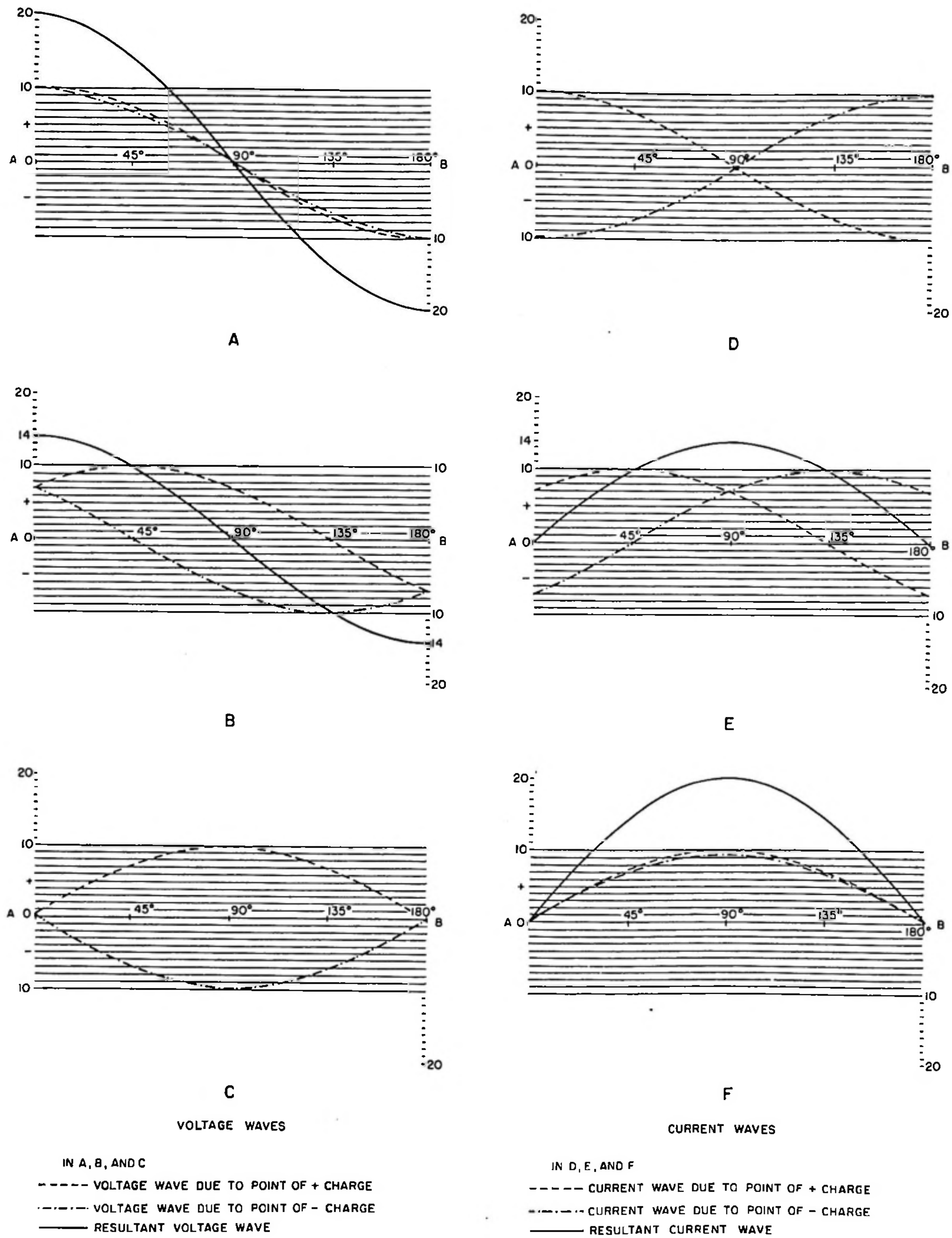


Figure 13-38. Combining of Reflected Waves To Produce Resultant Voltage and Current

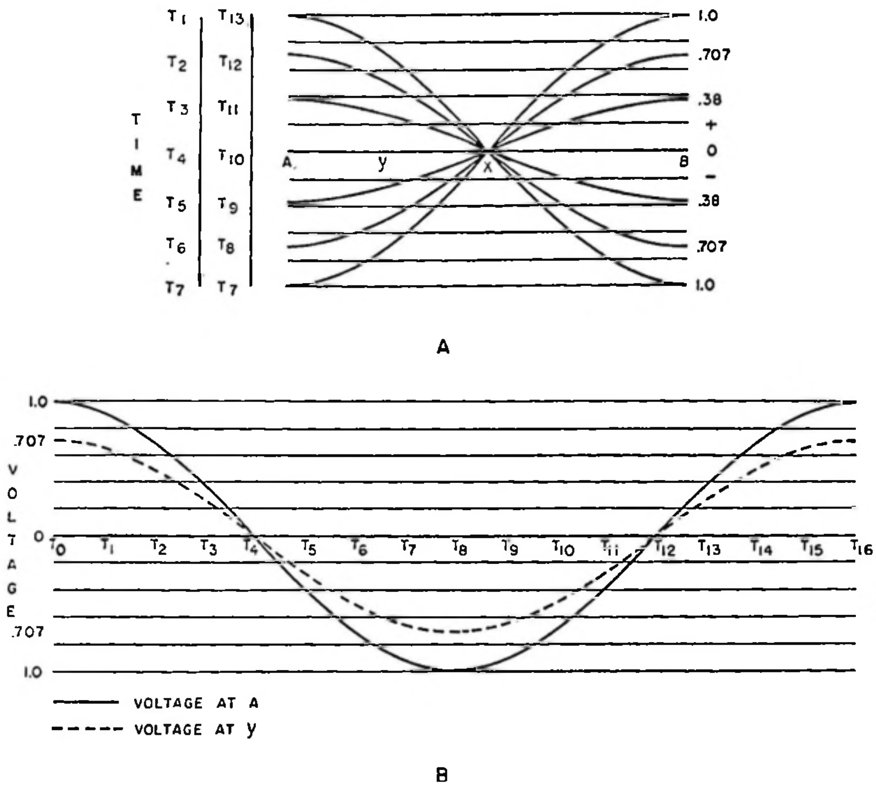


Figure 13-39. Voltage Distribution Drawing and Graph Showing Sinusoidal Voltage at Each Point

+10 volts at time T_1, +7.07 volts at time T_2, +3.8 volts at time T_3, 0 volts at time T_4, —3.8 volts at time T_5, —7.07 volts at time T_6, and —10 volts at time T_7. This completes the first half-cycle; the second half-cycle can be followed by reading upward from time T_7 to T_{13}. If you now fix yourself in position to see only the voltage at point A, you will see a sinusoidal voltage, as shown in part B of figure 13-39. In other words, if the voltage at point A is plotted against time on a separate graph, as in part B of figure 13-39, it will be a sinusoidal voltage. Similarly plotted against time, the voltage at point Y also will be a sinusoidal voltage, but the peak amplitude is only 7.07 volts. By plotting the voltage at each point along the wire, you will find that the voltage at every point is sinusoidal (except, of course, at the reference point X). However, the peak amplitude at any point is a function of the distance of that point from the reference point X.

13-217. METER READINGS. If you use an ac meter to measure the voltage at each point along the wire of length A to B, and if you plot the meter readings against a position on the wire, you will obtain the solid-line voltage curve, E, of figure 13-41. An ac meter does not show polarity; therefore, this curve is drawn without reference to polarity indications. Also, because the measurements are usually made with a meter that reads in effective (root-mean-square) values, the amplitudes indicated in figure 13-41 are only .707 of those indicated in part B of figure 13-39.

13-218. STANDING WAVES. The solid-line curve in figure 13-41 represents a standing wave of voltage. In other words, as measured with an ac meter, the voltage along the half-wave wire seems to be a wave that remains motionless. Actually, as indicated in parts A and part B of figure 13-39, the voltage at each point is constantly varying. As will be explained later, the presence of standing waves on a transmission line is a positive indication that reflection is occurring at some point on the line (usually at the load). In transmission line applications, standing waves are of considerable importance; they enable measurements of frequency and wavelength, are an indication of the amount of radiation that may be expected from the line, and are an excellent indication of how well the load is matched to the line.

13-219. STANDING WAVES OF CURRENT.

13-220. Again referring to the length of wire from A to B of figure 13-39, you can see that a number of the instantaneous current distribution curves are shown in part A of figure 13-40. By plotting the current at points X and Y against time, you will find that the current at each point in the wire varies sinusoidally, as shown in part B of

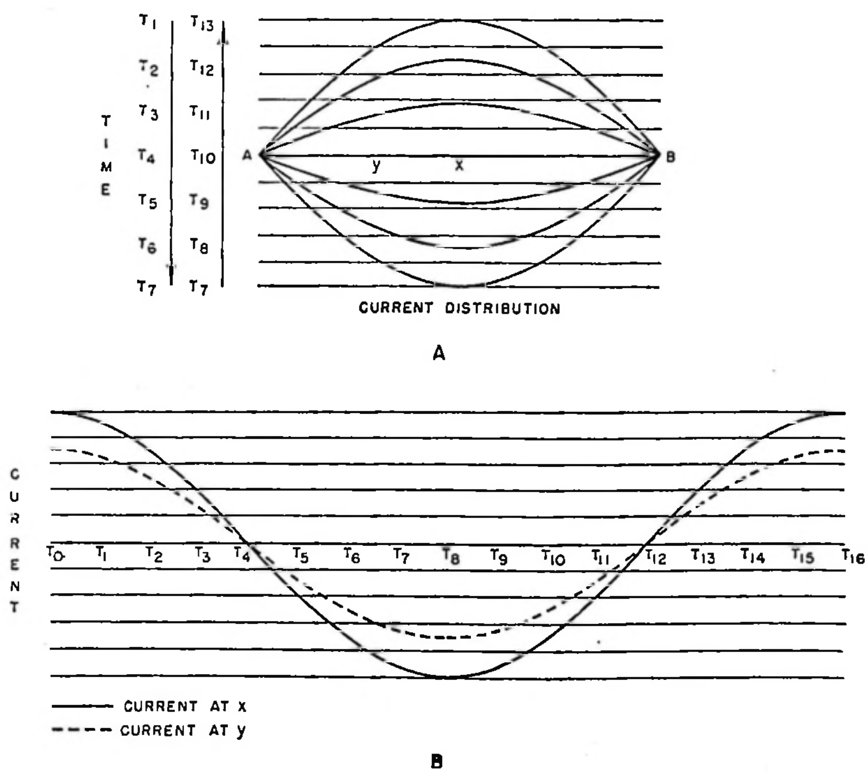


Figure 13-40. Current Distribution Drawing and Graph Showing Sinusoidal Current at Each Point

figure 13-40. Measurements with an ac meter result in the dashed-line current standing wave shown in figure 13-41.

13-221. You should remember that the half-wave wire (figures 13-39, 13-40, and 13-41) is used here as a means of explaining the cause and meaning of standing waves. You should also remember that these explanations of waves are simplified. In actual transmission lines the exact distribution of standing waves may differ considerably from that shown in these illustrations. Later discussions will show that the phase and amplitude relationships of voltage and current waves, as well as their positions on the line, are a function of the type of line, the type of load connected to the line, impedance matching, and other characteristics of the particular circuit arrangement.

13-222. TRANSMISSION OF ENERGY.

13-223. In any electric circuit the ability to do work (energy) is in the electric field. Basically, then, the problem in transmitting electric energy over appreciable distances is to bring the generated field to the load. When the load is connected directly to the generator, as shown in figure 13-23, the explanation that electrons are forced through the wires by electrical pressure is reasonably satisfactory. This explanation is, of course, the familiar water analogy. A pump (electrical pressure) forces water (current) through pipes (wires). Such an analogy implies a continuous force through the wire and load, but there is actually no such continuous force over a transmission line. Consequently, transmission over distances must be explained in terms of electric waves or, more precisely, electromagnetic waves.

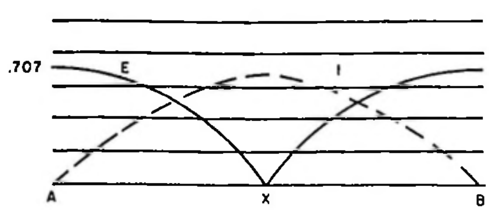


Figure 13-41. Standing Waves of Voltage and Current on Half-Wave Wire

13-224. RIBBON WAVE.

13-225. In figure 13-42, a battery is connected through a relatively long two-wire transmission line to a load at the far end of the line. At the moment the switch is closed, neither voltage nor current is distributed over the line. Point A becomes a point of positive potential, and point B becomes a point of negative potential. As previously explained, these points of potential move down the line. However, as the initial points of potential leave points A and B, they are followed by new points of potential which the battery adds to A and B. This is merely saying that the battery maintains a constant potential difference between points A and B. A short time after the switch is closed, the initial points of potential have reached points A' and B', and the wire sections A to A' and B to B' are at the same potential as A and B, respectively. The points of charge are represented by + and — signs along the wires, the currents in the wires are represented by short arrows, and the direction of travel is indicated by an

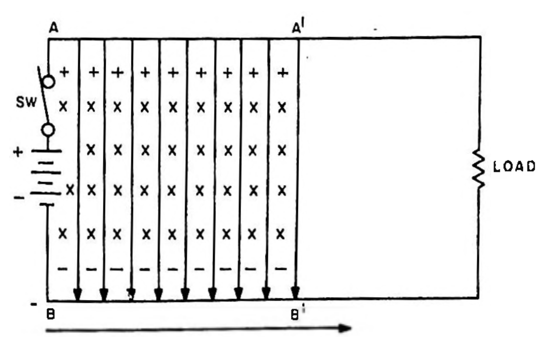


Figure 13-42. DC Voltage Applied to Line

arrow below the line. Conventional lines of force represent the electric field that necessarily exists between the opposite kinds of charge on A to A' and B to B'. Crosses (tails of arrows) indicate the magnetic field created by the electric field moving down the line. The moving electric field and its accompanying magnetic field constitute an electromagnetic wave that is moving from the generator (battery) toward the load. This type of electromagnetic wave is sometimes called a ribbon wave, because it moves down the line much as a ribbon would if inserted at the battery and pulled toward the load. This ribbon wave travels approximately at the speed of light in free space and, when it reaches the load, the energy reaching the load is equal to that developed at the battery. (This assumes no losses in the transmission line.) Assuming that the load absorbs all of the energy, there will be no reflection, and the current and voltage will be evenly distributed along the line.

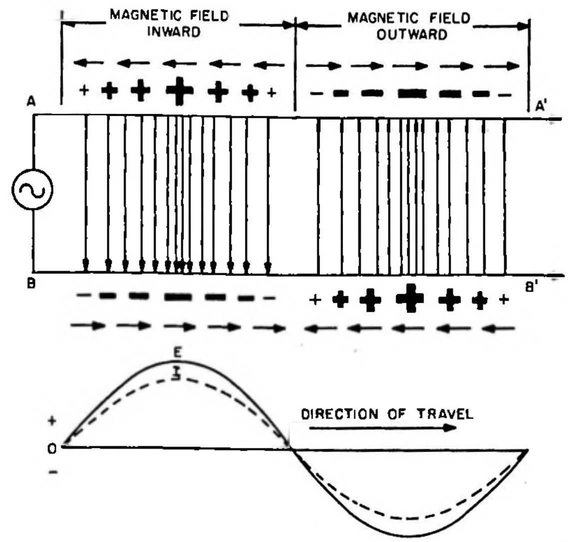


Figure 13-43. AC Voltage Applied to Line

13-226. SINUSOIDAL AND INCIDENT WAVES.

13-227. When the battery of figure 13-42 is replaced by an ac generator, as in figure 13-43, each successive instantaneous value of the generator voltage is propagated down the line at the velocity of light. The action is like that described for the ribbon wave, except that the applied voltage is sinusoidal instead of constant. Neglect possible transients and assume that the switch is closed at the moment the generator voltage is passing through zero to make point A become positive. At the end of one cycle of generator voltage, the current and voltage distribution will be as shown in figure 13-43. In this illustration conventional lines represent the electric fields. For simplicity, the magnetic fields are not shown, but their directions are indicated by legends in the drawing. Points of charge are indicated by + and — signs, larger signs indicating points of higher amplitude of both voltage and current. Short arrows indicate directions of current (electron movement). The waveform drawn below the line represents the voltage, E, and current, I. It is assumed that the line is infinite in length so that there is no reflection. Thus, traveling sinusoidal voltage and current waves continually travel in phase from the generator toward the load, or far end of the line. Waves traveling from the generator to the load are called incident waves.

13-228. DISTRIBUTION OF ENERGY.

13-229. The discussion of oscillations in a half-wave wire showed that at certain instants all of the energy is in the electric field and that at certain other instants all of the energy is in the magnetic field. Examination of the fields between the conductors in figures 13-42 and 13-43 shows that part of the energy is in the electric field and the remaining energy is in the magnetic field.

13-230. DIVISION OF ENERGY.

13-231. The maximum energy, W, stored in the inductance (magnetic field) per unit loop length is $W_L = LI^2/2$; the maximum energy stored in the capacitance (electric field) per unit length is $W_C = CE^2/2$. It can be shown that the impedance, Z_0, of a line is:

$$Z_0 = E/I = \sqrt{L/C}$$

Squaring:

$$E^2/I^2 = L/C$$

Cross-multiplying:

$$CE^2 = LI^2$$

Since $CE^2 = LI^2$, $W_L = LI^2/2 = CE^2/2 = W_C$. This means that the energy is equally divided between the electric and magnetic fields (assuming high-Q lines; that is, WL/R is high).

13-232. VOLTAGE AND CURRENT.

13-233. The foregoing equations indicate that the energy is in the electromagnetic field. This seems at variance with more familiar explanations in terms of voltage, current, and resistance. It might be well, therefore, to consider the relationship of fields and voltage and current.

a. Electrons and protons are particles of electricity; they cannot be created or generated by man.

b. Man can furnish energy to separate charges. The energy thus expended is then in the force (electric field) between the separated charges. Electric potential is a measure of the amount of work done in separating the charges, and the field will do the same amount of work in re-establishing an electrically neutral condition. Thus, voltage measurements in a transmission line are indirect measurements of the electric fields along the line.

c. Energy in the electric fields about points of charge causes currents in the wire conductors. Thus, the potential differences and currents along a line are functions of the electromagnetic waves moving along the line.

d. It is extremely difficult to make measurements in the electromagnetic waves themselves. However, line voltages and currents can usually be measured directly or calculated from measurements made with relatively simple equipment. Consequently, for most practical applications, line theory can be studied in terms of voltages and current.

13-234. SUMMARY.

13-235. ELECTROMAGNETIC WAVE.

13-236. The action of any transmission is due to the basic phenomena that produce electric forces when electric charges are separated. If points of charge are established on a line, the resultant field causes current in the line, and the points of charge move along the line. The moving field is an electric field in motion, and produces a magnetic field; therefore, an electromagnetic wave travels along the line. Because most of the energy is in the electromagnetic wave that is guided from generator to load, a transmission line is in effect a waveguide.

13-237. INCIDENT, REFLECTED, AND STANDING WAVES.

13-238. When all of the energy in incident waves is absorbed by the load, no reflection can take place. However, if all of this energy is not absorbed by the load, the unabsorbed portion is reflected and travels back over the line toward the generator.

Under this condition there are two waves on the line; incident waves travel from the generator toward the load, and reflected waves travel from the load toward the generator. These incident and reflected waves combine to produce standing waves on the line.

13-239. FREQUENCY, WAVELENGTH, AND WIRE LENGTH.

13-240. Because of the distribution of L, C, and R along a line, there is a definite relationship among frequency, wavelength, and wire length. These relationships and the production of standing waves form a basis for many of the practical applications of transmission lines. Thus, lines are used as frequency and wavelength measuring devices, as one-to-one step-up, step-down, and impedance-matching transformers, as resonant circuits in oscillators and amplifiers, and as excellent insulators in radio-frequency circuits.

13-241. PRACTICAL APPLICATIONS.

13-242. Despite the fact that the energy is in the electromagnetic waves that the line is guiding, line operation is usually explained in terms of line voltages and currents. This is because it would be difficult to solve complex line problems by field theory, and the voltage and current equations are sufficiently accurate for engineering purposes. Throughout the remainder of this manual, therefore, lines will be discussed in terms of voltages and currents. The brief, simplified discussions in this chapter should provide a better understanding of the why and how of line action. A reasonably good mental picture of the invisible action which occurs is always an aid to understanding the operation of any electrical device.

SECTION IV

TRANSMISSION LINES

13-243. GENERAL.

13-244. Sections I through III discussed transmission-line theory in terms of elemental charges, electric fields, magnetic fields, and electromagnetic waves. In this section the infinite line, characteristic impedance, lumped and distributed constants, and practical applications are considered in terms of line voltages and currents, and voltage and current waves.

13-245. THE INFINITE LINE.

13-246. The theoretical behavior of two parallel-wire conductors of infinite length forms a starting point or basis for studying practical lines. Stated simply, this theoretical line is of importance for the following reasons: (1) many lines, particularly power and telephone lines, are so long that they act approximately as infinite lines; (2) the transmission of electric waves over any line can be analyzed in terms of an infinite line; (3) any properly terminated line will act as an infinite line; and (4) the infinite line is the simplest to study.

13-247. TERMINOLOGY.

13-248. Every transmission line has two ends, the end to which power is applied and the end to which the load is connected to receive power. The first is called the sending end, the input end, or the generator end; the second is called the receiving end, the output end, or the load end. The opposition of a line to current from a generator is called the input impedance, Z_{in}. Input impedance is expressed by the equation $Z_{in} = E/I$, where E is the voltage and I is the current at the input terminals of the line. The symbol Z_r is used to indicate the impedance of the load at the receiving or load terminals of the line. Other terms and symbols will be introduced as required in the text.

13-249. CHARACTERISTIC IMPEDANCE.

13-250. The input impedance of any real line is a function of frequency and the load impedance. For any line there is one, and only one, value of load impedance that will make the input impedance equal to the load impedance. The required value of load impedance is equal to the input impedance of an infinite length of the particular line. This value of impedance is called the characteristic impedance. The concept of characteristic impedance is of particular importance. When terminated at its load end by a value of Z_r equal to its characteristic impedance, Z_o, any line has an input impedance equal to its characteristic impedance; that is, $Z_{in} = Z_r = Z_o$. When $Z_{in} = Z_r = Z_o$, the impedance (looking toward the load end) at every point in a line is also equal to Z_o, the load impedance absorbs all of the energy reaching it from the generator, there are no reflections, and, hence, no standing waves on the line. These ideas of impedance matching are of considerable importance and will be considered throughout this chapter.

13-251. FUNDAMENTAL CONCEPT.

13-252. According to the concept of infinity, an infinite line is so long that energy applied at the near end will never reach the load or far end of the line. If none of the input energy ever reaches the load end of a line, it is feasible to assume that, regardless of its value, the load impedance, Z_r, will have no effect whatsoever on the input impedance, Z_{in}. Suppose, for example, that a length of 500 miles is cut from either end of an infinite line. Whether the length is cut from the near end or from the far end of the line, the generator sees no difference in the line because the line is still infinite in length. Since neither the load impedance nor the cutting away of a length has any effect, we can deduce that the input or characteristic impedance of an infinite line is a function of line construction and frequency only. Carrying this deduction a little further, we can conclude that the characteristic impedance of any line is determined entirely by frequency and the line construction, that is, the size, shape, spacing, number, and electrical characteristics of the conductors, and the electrical properties of the dielectric material.

13-253. DC VOLTAGE APPLIED TO INFINITE LINE.

13-254. The development of a ribbon wave has been discussed in terms of the electric field and its effects on charges in the wires. We will now consider this same wave in terms of voltage and current.

13-255. PROPAGATION.

13-256. The distributed inductance and capacitance of a line are represented in part A of figure 13-44 by a series of extremely small lumps. The line resistance, R, and the shunt conductance, G, are neglected, and the line is assumed to be infinite in length.

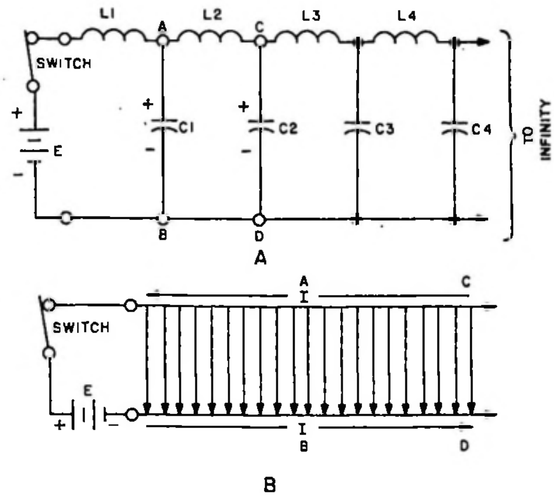


Figure 13-44. Propagation of Voltage and Current Along an Infinite Line

13-257. VOLTAGE WAVE. As illustrated in figure 13-44, at the moment the switch is closed, the battery voltage, E, is applied to the input terminals of the line. At this same instant capacitor C_1 contains no charge; therefore, it is a short circuit across the points A to B. Consequently, current I begins to flow through L_1 into C_1, but not beyond the limits of A to B. As current I flows into C_1, this capacitor becomes charged and a voltage appears across A to B. Note that in this lumped circuit, L_1 delays the charging of C_1 and no current can flow beyond the limits of A to B until C_1 acquires some charge. As C_1 charges, the voltage across it increases and causes current to flow through L_2 into C_2. As C_2 charges, its voltage increases and causes current to flow through L_3 into C_3, and this action continues down the line. First C_1 charges to voltage E, then C_2, then C_3, etc. In other words, there is a voltage wave traveling along the line. This wave starts at the sending end and travels with a certain velocity, V. All of the line that the wave has reached is charged to voltage E, as shown in part B of figure 13-44. Beyond the wavefront, the line is not charged.

13-258. CURRENT WAVE. The voltage wave is necessarily accompanied by a current wave, I, as shown in part B of figure 13-44. To charge C_1, current I must flow through L_1; to charge C2, current I must flow through L_2; etc.

13-259. VOLTAGE AND CURRENT WAVES SUSTAIN ONE ANOTHER. For the voltage wave to travel down the line, current I must flow through each L into the following C, but for current I to flow through the next L a voltage must first be built up across each C. This means simply that the voltage and current waves are mutually sustaining. This idea is in agreement with the theory that a moving electric field creates a magnetic field, and a moving magnetic field creates an electric field. This concept also agrees with the earlier statement that a magnetic field is an attribute or a different aspect of an electric field.

13-260. CONCLUSION.

13-261. The previous discussion has been simplified by neglecting the line resistance and shunt conductance. Certain effects that occur in lumped circuits also have been neglected. For example, the voltage across C in a lumped circuit does not build up to E and remain constant; damped oscillations occur and the voltage across C is alternately greater and smaller than the battery voltage, E. However, these effects become negligible as real line conditions are approached, because of the reduction in size and increase in the number of lumps.

13-262. DISTRIBUTED CONSTANTS.

13-263. LINE CHARACTERISTICS.

13-264. When a generator is connected directly to a power-consuming load, the load may be represented as a simple resistor, as in figure 13-45. This is because only the resistance of the load is capable of dissipating power. The power consumed in the resistor is equal to the power supplied by the generator. Assume that the same power-consuming device is connected to the generator by means of a long transmission line, as in figure 13-46. Although this circuit might appear to be identical with the directly connected circuit of figure 13-45, it is now found that in the second circuit the power consumed in the load is less than the power supplied by the generator. Thus, the transmission line has characteristics that cause a power dissipation (loss) between the generator and the load. This power dissipation does not take place at any one point, but occurs equally along the entire length of the line. The electrical characteristics of the line that cause the power loss are distributed uniformly over the entire line. For this reason, the characteristics are called the distributed constants of the line. If the power dissipation took place at a distant point on the line because of concentrated characteristics, as a coil concentrates

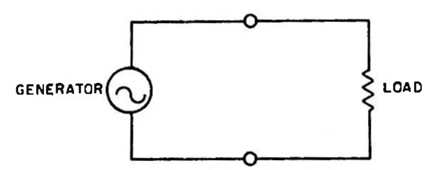


Figure 13-45. Generator and Load Connected Directly

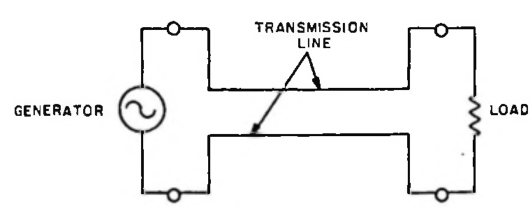


Figure 13-46. Generator and Load Connected by Long Transmission Line

inductance or a capacitor concentrates capacitance, the characteristics would be called lumped constants. There are four distributed constants: series resistance per unit length, series inductance per unit length, shunt capacitance per unit length, and shunt conductance (more commonly known as leakage) per unit length.

13-265. SERIES RESISTANCE PER UNIT LENGTH. The total resistance of a transmission line is found by multiplying its series resistance per unit length by the total length of the line. Thus, if the series resistance of a line is 10 ohms per mile, and the length is 100 miles, its total resistance is 1000 ohms. The distributed resistance of a transmission line depends on the size of the wires and the frequency of the traveling wave. The variation of resistance with frequency results from the phenomenon called skin effect. The resistance increases as the diameter of the wire decreases, and as the frequency of the transmitted wave increases. It is represented by the symbol R, and is expressed in ohms per loop mile. A loop mile (see figure 13-47) is the total length of both lines between two points 1 mile apart. Thus, a loop mile of line is 2 miles long, and the resistance in ohms per loop mile is twice the resistance of one conductor per mile. The series resistance per loop mile may be represented in an equivalent circuit (see part A of figure 13-48) where half the resistance is considered to be lumped in series with each wire; or (see part B of the figure) where the entire resistance is concentrated in one conductor only.

13-266. SERIES INDUCTANCE PER UNIT LENGTH. Self-inductance is that property of a circuit which causes a countervoltage to be induced in the circuit by a change of current in the circuit. In a transmission line through which a changing current is flowing, a voltage is induced all along the line. This indicates that series inductance is distributed over the entire length of the line. The magnitude of this series inductance is determined by the size of the wires and their separation. It increases as the center-to-center distance between the wires increases and as the diameter of the wire decreases. The distributed inductance is represented by the symbol l, and is expressed in henrys per loop mile. The total inductance per loop mile may be represented also in two halves, each half representing the inductance in one of the two conductors (see part A of figure 13-49), or it may be considered as concentrated in one conductor (see part B of the figure). The series inductance of a transmission line causes an opposition in the form of inductive reactance to the alternating currents. Inductive re-

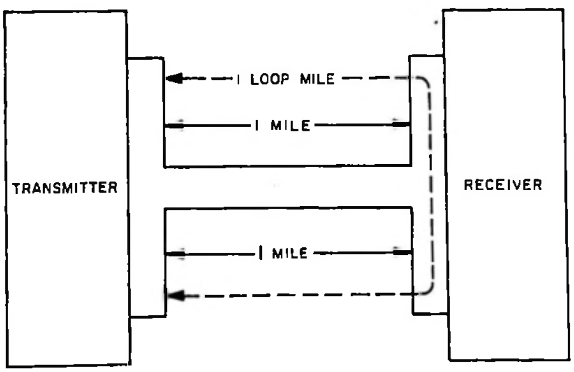


Figure 13-47. One Loop Mile of Transmission Line

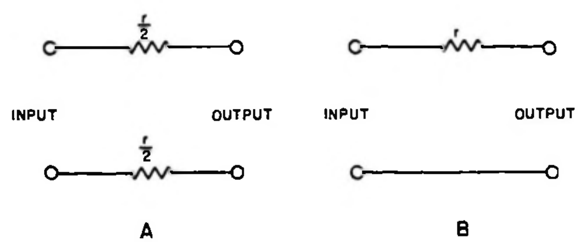


Figure 13-48. Distributed Series Resistance

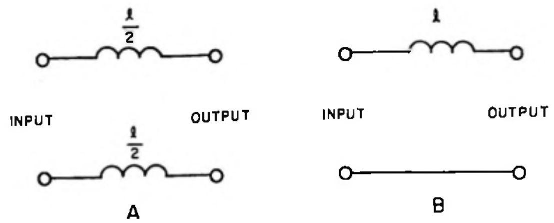


Figure 13-49. Distributed Series Inductance

actance, X_L, is a function of frequency and is expressed by the formula $X_L = 2\pi fL$; therefore, as the frequency increases, the inductive reactance, or current-opposing effect, increases.

13-267. SHUNT CAPACITANCE PER UNIT LENGTH. A capacitor consists of two metallic conductors separated by a nonconducting substance, such as air or some other dielectric. The capacitance is large when the area of the conductors is large, as in the case of two flat plates, and the capacitance increases when the distance between the plates is made smaller. In a capacitor, the separation of the plates is deliberately made a small fraction of an inch. A transmission line also consists of two metallic conductors separated by a dielectric. As a result, the line capacitance is distributed over the entire length of the line. However, the area of the surface of a length of transmission line is much less than the area of a conventional capacitor plate, and the distance between the lines is much greater than the separation between the plates of a capacitor; the distributed capacitance of a transmission line appears between the adjacent wires, and is called the shunt capacitance. The shunt capacitance per unit length is determined by the size of the wires, the distance between the wires, and the nature of the dielectric material between them. The capacitance increases as the diameter of the wires increases, as the center-to-center separation of the wires decreases, and as the dielectric constant increases. The distributed capacitance is represented by the symbol C; it is expressed in farads per loop mile, and is indicated as a capacitor shunted across the two conductors (see figure 13-50). The important thing to keep in mind is that the distributed capacitance causes capacitive reactance to develop across, or in shunt with, the transmission line. This causes a shunting of the alternating current, and the result is that less alternating-current energy reaches the receiving end. Since capacitive reactance decreases when the frequency increases ($X_C = \frac{1}{2\pi fC}$), the shunting effect becomes greater as the frequency increases.

13-268. SHUNT CONDUCTANCE PER UNIT LENGTH. Because the dielectric between the two wires of a transmission line is not a perfect insulator, a leakage current exists between the wires. The dielectric separates the wires of the transmission line over its entire length because leakage exists at every point along the line. In open-wire lines the dielectric between the conductors is air. Although dry air is an almost perfect insulator, outdoor air is seldom dry, and its conductivity increases greatly in damp weather. In cables the dielectric consists of the insulation surrounding the individual conductors. The best insulators conduct extremely small amounts of current. Although cable leakage is unaffected by dampness of the outdoor air (which is excluded by the outer covering of the cable), it does vary somewhat with temperature. Since the leakage takes place through a conducting path between

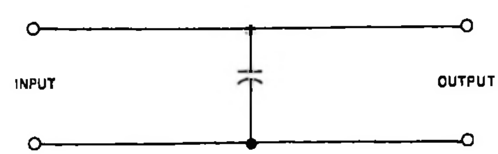


Figure 13-50. Distributed Shunt Capacitance

the wires, this corresponding line characteristic may be called shunt conductance, or leakage. It is represented by the symbol G; it is expressed in mhos per loop mile, and is indicated by the reciprocal of the resistance between the two wires (see figure 13-51). Note that line leakage acts as a shunt to the flow of alternating currents in the transmission line, and, as a result, current shunted across the line cannot reach the receiving terminal.

13-269. REPRESENTATION OF LINE BY EQUIVALENT NETWORK.

13-270. A unit length of transmission line, or line section, may be represented for convenience as an equivalent network composed of all four distributed constants of the actual line, as shown in figure 13-52. Three alternative representations of the actual line section are shown. The equivalent tee section, in C, is formed by placing the series constants (resistance and inductance) in one line, half on each side of the shunt constants (capacitance and conductance). It must be remembered that in figure 13-52 the lumped elements, R, L, C, and G, are merely convenient representations of properties that are actually distributed over the whole length of the section. Their values are such that they would have substantially the same effect on a transmitted signal, and on any circuit connected to the line section, as the actual distributed constants of the section.

13-271. A transmission line which is more than one section long may be represented by an equivalent network formed by connecting two or more unit line sections, as in figure 13-53. The electrical behavior of the line may then be studied by analyzing the behavior of the equivalent network.

13-272. RADIATION FROM LINE.

13-273. Under certain conditions, a transmission line acts as an antenna and radiates power. Such power, of course, is not delivered to the load. The input power to the transmission line, therefore, must be increased to make up for the power lost by radiation, just as it must be increased to make up for the power dissipated by the distributed line constants. At low frequencies, such as the frequencies used in normal telephone transmissions, the amount of radiation is so small that it can be neglected. At higher frequencies, however, such as those used in radio transmission and some earlier telephone transmissions, the radiation is greater, and continues to

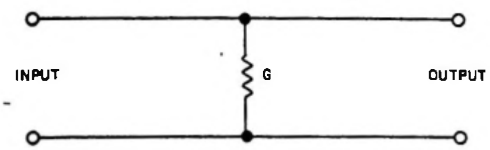


Figure 13-51. Distributed Shunt Conductance

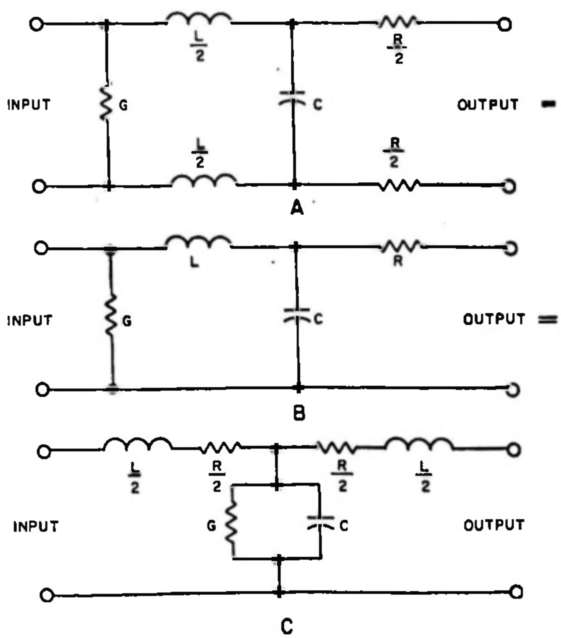


Figure 13-52. Line Section of Unit Length

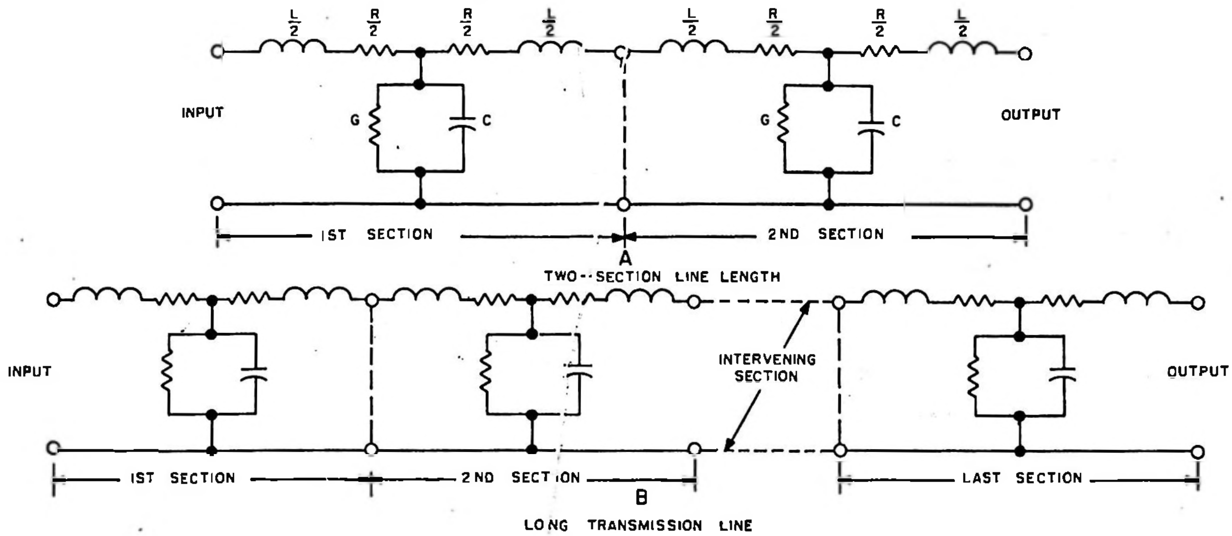


Figure 13-53. Equivalent Network of Transmission Line

increase as the frequency increases. The radiation also increases with an increase in the length of the line. Usually, radiation is not considered as a line constant.

13-274. CHARACTERISTIC IMPEDANCE OF LINE.

13-275. The distributed constants of a transmission line determine its operating characteristics, which, in turn, affect the signal being transmitted over the line. The most important operating characteristic of a transmission line is its characteristic impedance, Z_O. For a given frequency the characteristic impedance of a transmission line is a property of the line itself, dependent only on its distributed constants. It is entirely independent of the length of the line, the internal impedance of the generator supplying the line, and the magnitude of the load placed across the terminals of the line. The characteristic impedance does, however, vary with the frequency of the transmitted signal. The importance of characteristic impedance lies in the requirements it imposes on the generator and load impedance to obtain maximum transfer of power to the load.

13-276. CALCULATION OF LINE IMPEDANCE.

13-277. To demonstrate the meaning of characteristic impedance, the tee section representation shown in part A of figure 13-53 is transformed first to a pi section, composed of three resistances, as shown in figure 13-54. Resistances alone are used, because the mathematics necessary to derive the value of a true characteristic impedance, including inductance and capacitance, of an actual transmission line is beyond the scope of this manual. Using only resistances, the mathematics is much simpler, and the meaning of the results is the same as far as characteristic impedance is concerned.

13-278. Let the three resistances of the pi section shown in figure 13-54 be 300 ohms each, as shown in part A of figure 13-55. The input impedance of this line section is 200 ohms, as illustrated. Now connect a

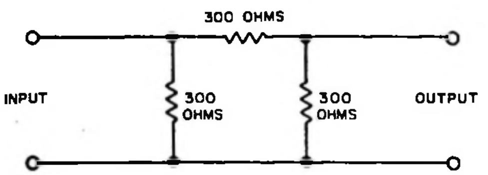


Figure 13-54. Unit Line Section Represented as a Pi Section

second section ahead of the first, making the line two sections long, as shown in part B of the figure. The first section acts as an equivalent 200-ohm resistor across the second. Combining the resistance in the manner shown, it is found that the input impedance of a line two sections long is 175 ohms. Next, let the line be three sections long, as shown in part C. The first two sections act

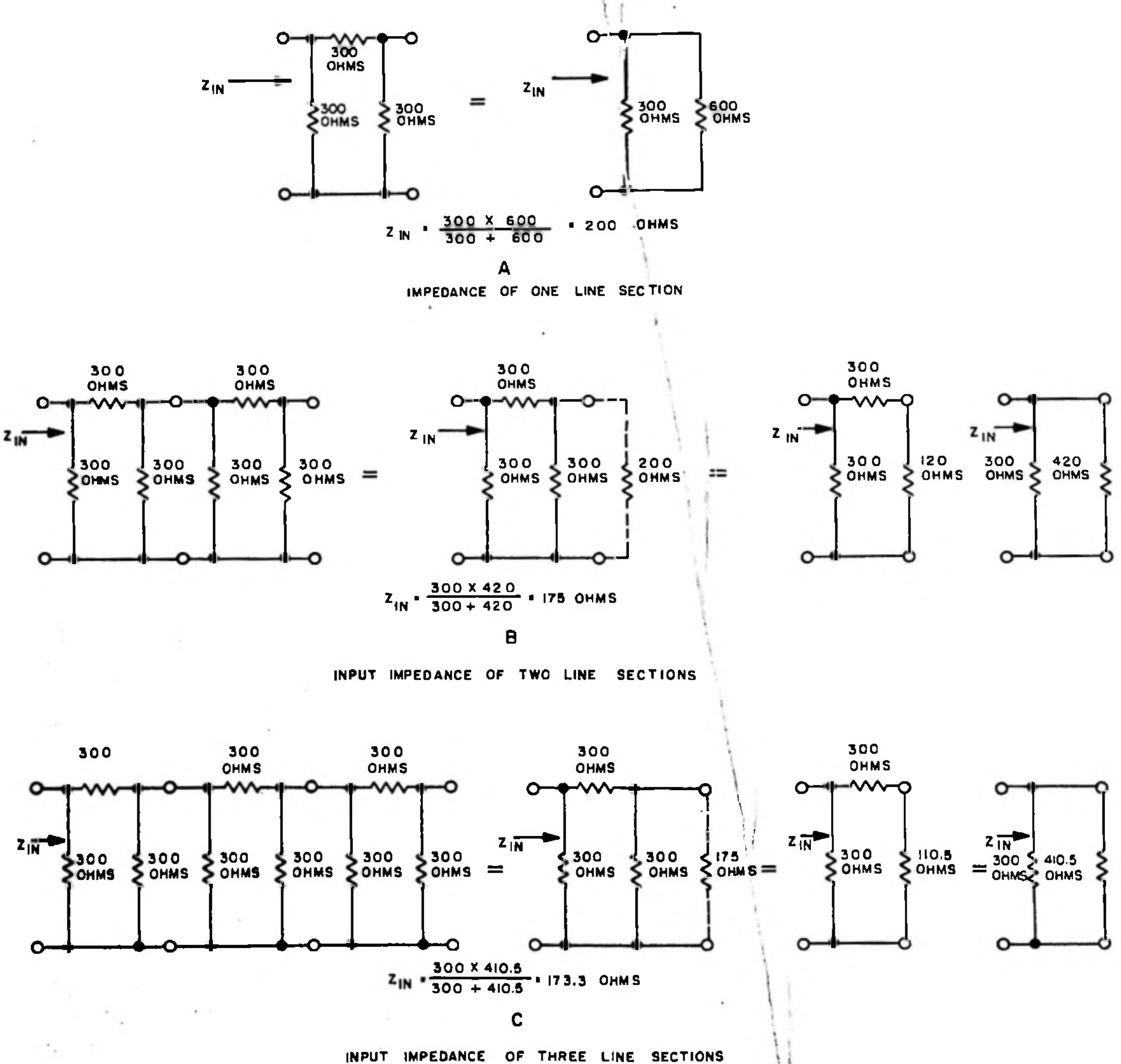


Figure 13-55. Calculation of Input Impedance

as an equivalent 175-ohm resistor across the third section. The resistances are combined in the manner illustrated, resulting in an input impedance of 173.3 ohms for a line three sections long. If the line were composed of four sections, the first three would act as an equivalent 173.3-ohm resistor across the fourth, and the input impedance would then be 173.2 ohms.

13-279. The values of input impedance found in paragraph 13-278 are plotted in figure 13-56. The graph shows that the input impedance of the transmission line decreases toward a fixed value as the length of the line increases. Note that this is true because the amount of decrease becomes smaller and smaller as more and more sections are added. The input impedance cannot decrease to zero because of the distributed series resistance. When the line is very long (in consideration of the frequency), the addition of more sections causes practically no change in the input impedance; that is, the input impedance takes on a constant value. This constant input impedance of a very long line is called the characteristic impedance of the line. Theoretically, the transmission line must be infinitely long in order for the input impedance to be equal to the characteristic impedance of the line. But the graph shows that (for the example chosen), after taking only four sections, the line has an impedance value which is very nearly equal to its characteristic impedance.

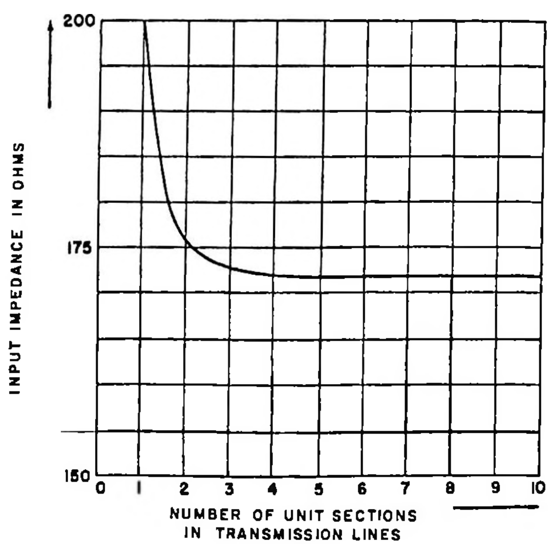


Figure 13-56. Input Impedance Versus Transmission Line Sections

13-280. For an actual line with distributed inductance and capacitance, the input impedance also approaches the characteristic impedance as additional sections are connected, but more slowly.

13-281. In the example previously discussed the inductance and capacitance of the line were omitted for the sake of simplicity. For an actual transmission line, however, their presence causes the characteristic impedance to vary with frequency. If the characteristic impedance for a transmission line is assumed to be 600 ohms at 1000 cycles, it will be slightly lower at frequencies below 1000 cycles. In most cases, the differences in the values of the characteristic impedance are not great enough to be a source of difficulty.

13-282. TERMINATING IMPEDANCE.

13-283. The input impedance of a transmission line is also equal to the characteristic impedance of the line when, no matter how short or how long, it is terminated by an impedance equal to its characteristic impedance.

13-284. To demonstrate this, a unit line section will again be represented as a pi section composed of three 300-ohm resistors, as shown in figure 13-54. Again, this is not a true representation of a transmission-line section; however, the meaning of the results achieved is the same as for the more difficult problem involving shunt capacitance and series inductance. Let the load across the line section be a variable resistor, as shown in part A of figure 13-57. The input impedance of the section is equal to the characteristic impedance of the line when the terminating impedance also is equal to the characteristic impedance of the line (in this case, 173.2 ohms).

13-285. When the load resistance is 50 ohms, the resistances may be combined as shown in part B of the figure, and the input resistance will be only a little less than 160 ohms. The graph in figure 13-58 shows how the input impedance varies as the load resistance ranges from 0 ohms to 300 ohms. When the load resistance is 173.2 ohms, the input impedance may be calculated in the same manner, and is found also to be equal to 173.2 ohms. In figure 13-56, however, it is shown that the line section of figure 13-54 had a characteristic impedance of 173.2 ohms. The obvious conclusion is that when a line section is terminated by an impedance equal to its characteristic impedance, the input impedance is equal to the characteristic impedance.

13-286. Any length of the same type of line may be connected to the input of the section represented in figure 13-57, and the input impedance of the entire line will still be equal to the characteristic impedance of the line. In part A of figure 13-59, the first section represents one section of the new line connected to the input of the second section. The latter is the terminated line section of figure 13-57. Its input impedance has already been calculated to be 173.2 ohms. This is the impedance in which the first, or new, section is terminated. The new section acts exactly as though it were terminated in a 173.2-ohm resistor, as shown in the equivalent circuit, at the right in part A of figure 13-59. It has been shown, however, that a section of line so terminated has an input impedance of 173.2 ohms. If another section of line is connected to the output of this terminated line, making a line three sections long, as shown in part B of the figure, the same result will be obtained. The newly added section is terminated in the input impedance of the old sections. Since this was equal to the characteristic impedance of the line, the new input impedance of

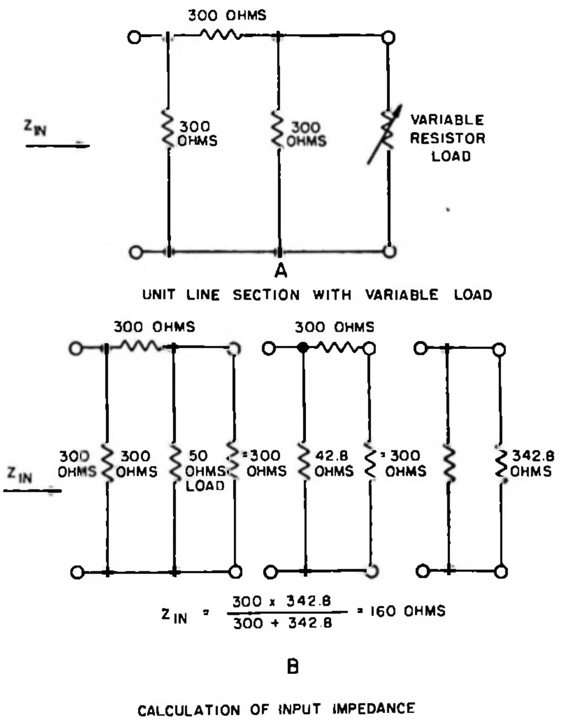


Figure 13-57. Variation of Input Impedance

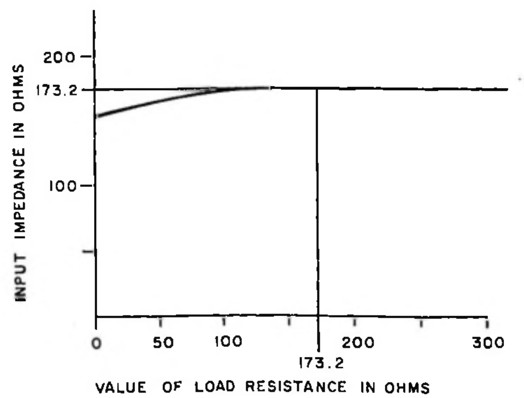


Figure 13-58. Input Impedance Versus Load Resistance

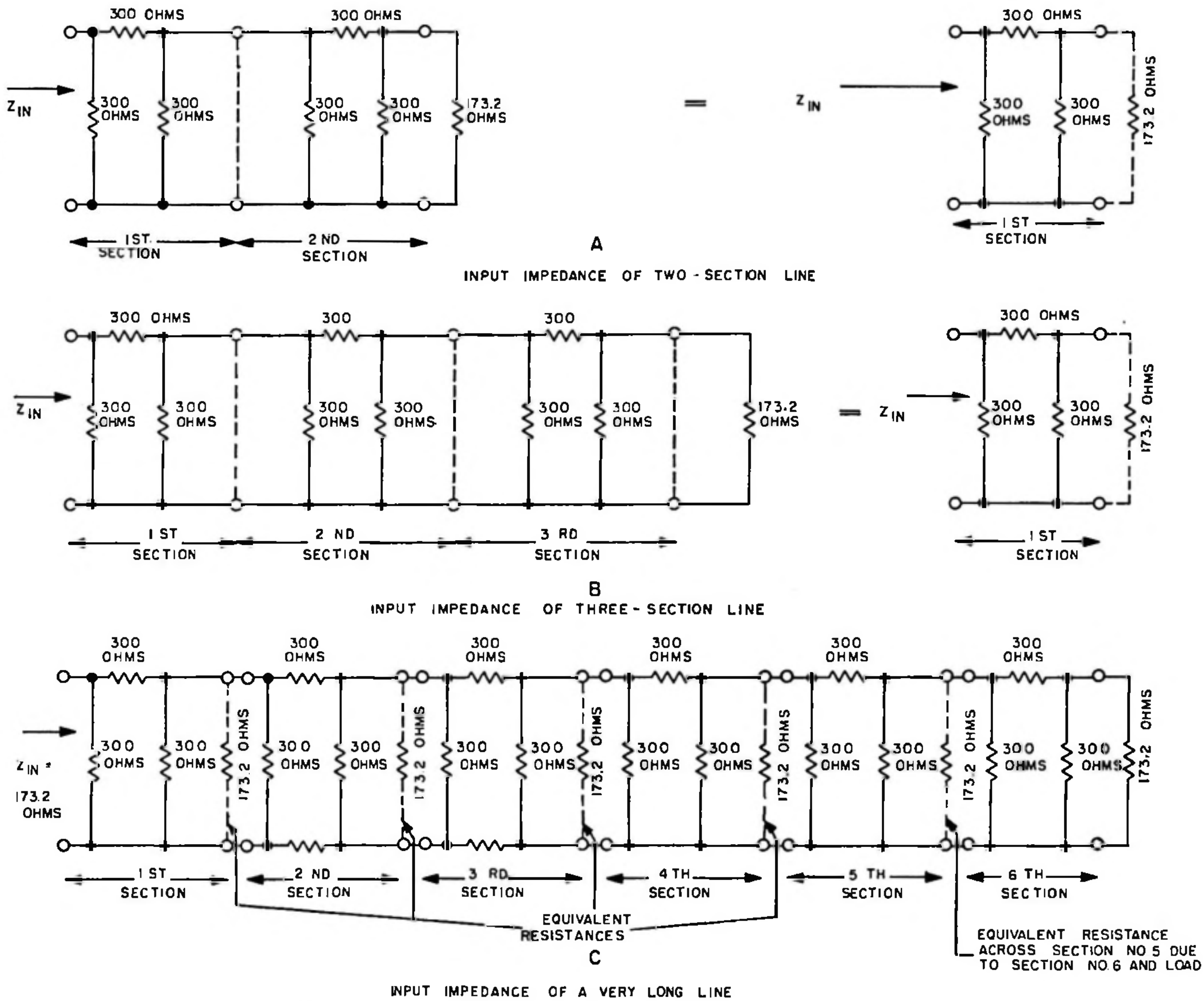


Figure 13-59. Input Impedance of Line Terminated by Z

the three sections is also equal to the characteristic impedance of the line. As the line increases in length, as shown in part C of the figure, each newly added section is terminated in an impedance equal to its characteristic impedance; therefore, the input impedance of the entire line is also equal to the characteristic impedance of the line. Also, the impedance looking toward the load from any point on the line is equal to the characteristic impedance.

13-287. The input impedance of a very long line is substantially equal to its characteristic impedance. This is true even when the long line is open-circuited or short-circuited at its load end.

13-288. The preceding facts may be summarized as follows: the input impedance of a transmission line is equal to its characteristic impedance (1) when the line is very long, no matter what its termination, or

(2) when it is terminated, no matter what its length, in a resistance equal to its characteristic impedance.

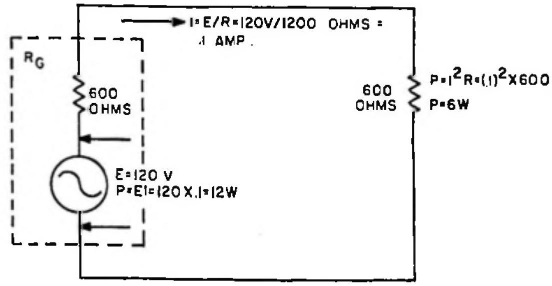
13-289. FACTORS AFFECTING CHARACTERISTIC IMPEDANCE.

13-290. For a given frequency, the value of the characteristic impedance of a transmission line is determined by the values of the distributed constants of the line. These are determined by the size of wire used, the distance between the wires, and the nature of the dielectric between the wires. The value of the characteristic impedance increases as the diameter of the wire decreases, as the center-to-center separation increases, and as the value of the dielectric constant decreases. Thus, the characteristic impedance is determined by the construction of the transmission line and is in no manner dependent on the physical length of the line. In addition, characteristic impedance varies with frequency.

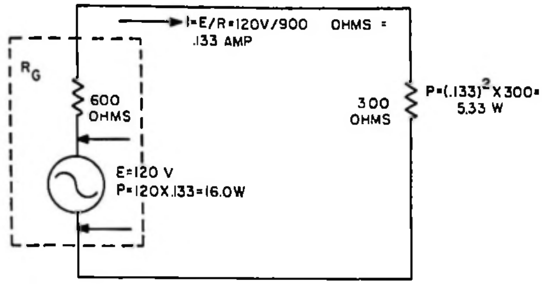
13-291. POWER TRANSFER.

13-292. The primary purpose of a transmission line is to transfer a maximum amount of power to the receiving end of the line from the transmitter antenna. The maximum power transfer occurs when the input impedance of the receiver (antenna) and the internal resistance of the generator (transmitter) are equal to the characteristic impedance of the transmission line. For this reason, the value of the characteristic impedance of a transmission line is important.

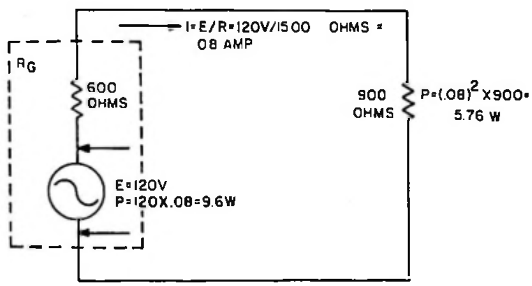
13-293. Consider the circuit in part A of figure 13-60, which shows a generator delivering power to a 600-ohm resistor. Let the internal resistance of the generator be 600 ohms, and let its voltage be 120 volts. The calculations show that the total output power of the generator is 12 watts, and that the power delivered to the 600-ohm load



A
USING A 600-OHM LOAD RESISTOR



B
USING A 300-OHM LOAD RESISTOR



C
USING A 900-OHM LOAD RESISTOR

Figure 13-60. Generator Operating into Variable Load

resistor is 6 watts. In part B of the figure the load resistor is changed to 300 ohms, causing the total output power of the generator to change to 16.0 watts, and the power of the load to change to 5.33 watts. In part C of the figure the load resistor is changed to 900 ohms, causing the total output power of the generator to change to 9.6 watts, and the power to the load to change to 5.76 watts.

13-294. The calculations of the values of power delivered to the load resistor in figure 13-60, supplemented by additional calculations to provide enough points for a smooth curve, are plotted in the graph of figure 13-61. The graph shows that maximum power is transferred from the generator to the load when the input impedance of the load equals the internal impedance of the generator.

13-295. Conclusions to be reached, based on the preceding discussion, are related to the characteristic impedance of a transmission line as follows: To the generator supplying power to a transmission line, the line is the load; consequently, the generator will deliver maximum power when the input impedance of the line matches the internal impedance of the generator. At the receiving end of the line, the line acts as the generator, delivering power to the load. The internal impedance of the line at the generator end is the characteristic impedance of the line. Therefore, the generator will deliver maximum power to the load when the load impedance is also equal to the characteristic impedance of the line. The input impedance of the line is also equal to its characteristic impedance, requiring that the generator impedance also be equal to the characteristic impedance of the line. Thus, a 600-ohm line should be supplied from a 600-ohm generator and be terminated in a 600-ohm load; maximum power transfer can then take place at both ends of the line, and, therefore, from the actual generator to the actual load. Under these conditions, the generator and the load are said to be matched to the line and to each other.

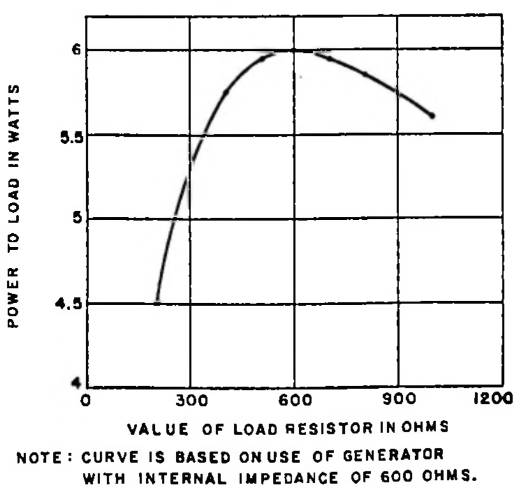


Figure 13-61. Power Transfer to Variable Load

13-296. TRANSMISSION LINE ATTENUATION.

13-297. A wave traveling along a transmission line is attenuated as it moves away from the generator. The attenuation amounts to a gradual diminishing of the amplitude of the current and voltage waves.

13-298. ATTENUATION OF CURRENT AND VOLTAGES.

13-299. Consider a transmission line terminated by a load having an impedance equal to the characteristic impedance of the line, as shown in figure 13-62. Because of the presence of the distributed series resistance of the line, there will be a series voltage

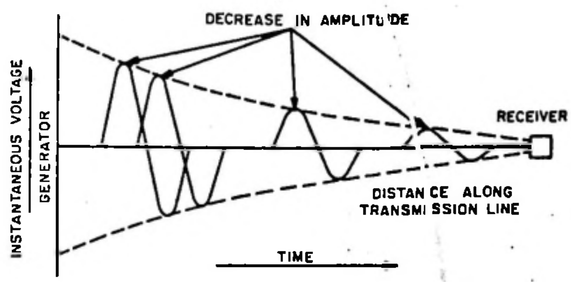


Figure 13-62. Attenuation of Current Voltage Waves Along a Transmission Line

drop (IR) along the entire length of the line. In addition, some of the current will not travel over the entire line and reach the load, because of the distributed leakage and capacitance between the wires, through which part of the current will return to the generator. The magnitude of both the current and the voltage waves, therefore, is diminished, and, since the power in the line is a product of the current and the voltage, the power gradually falls off as the distance from the generator increases. Figure 13-62 shows how the amplitude of the voltage, or the current, of the transmitted wave is decreased by the series resistance and shunt leakage.

13-300. If the distributed constants of the line are uniform over the entire length of the line (that is, if the distributed constants have the same values for every unit section), the amplitudes of the voltage and current waves decrease at the same rate over the entire length of the line. Figure 13-63, represents a line of at least four unit line sections. The impedance of the line is 600 ohms at all points on the line. The voltage at the input terminals of the line is 600 millivolts, and at the output terminal of the first section it is 60 millivolts. The decrease in voltage amplitude produced by the line section represents a ratio of 10 to 1. (This ratio was selected as an example.) The decrease in current amplitude represents same ratio. Thus, if the input current has a value of 1 milliampere, the output current will have a value of 0.1 milliampere. For the second line section, the input and output voltage are 60 millivolts and 6 millivolts, respectively, and the input and output currents are 0.1 milliampere and 0.01 milliampere, respectively. The figures show that in the first line section the voltage amplitude is decreased by 540 millivolts, and in the second line section it is decreased by 54 millivolts; thus, in both sections the ratio of input and output voltages is 10 to 1. In the current-wave amplitude, the amount of decrease is 0.9 milliampere in the first section and 0.09 milliampere in the second section; thus, the ratio of input and output currents is also 10 to 1 for both sections. In other words, the amount of decrease differs from section to section, but the ratio of input and output values of current and voltage is the same for all sections, as long as they have the same characteristics.

NOTE

The ratio of 10 to 1 used here is not the same for all types of lines; the ratio varies widely according to the properties of the lines as well as the length of the individual sections.

13-301. ATTENUATION OF TRANSMITTED WAVE.

13-302. In part A of figure 13-64, values of the voltage amplitude at different points

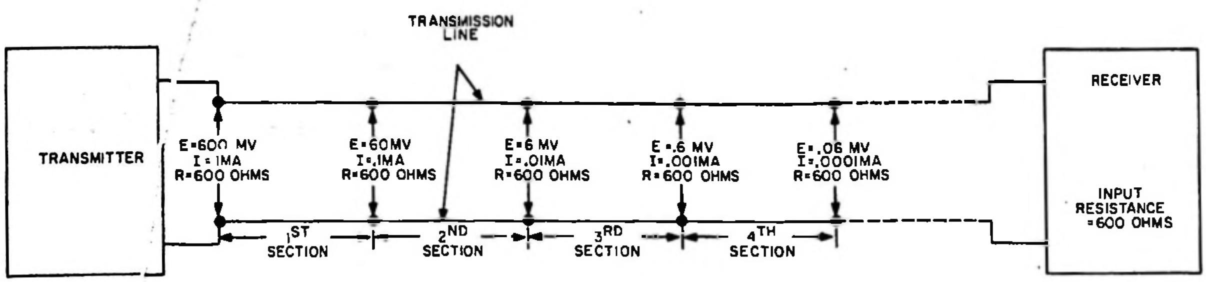


Figure 13-63. Attenuated Voltage and Current Points Along a Transmission Line

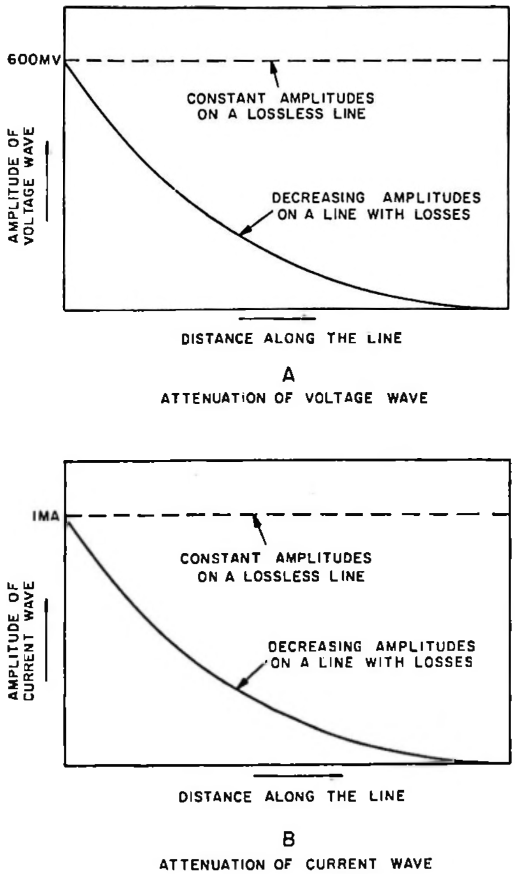


Figure 13-64. Attenuation of Transmitted Wave

along the line are plotted and connected by a solid line. In part B of the figure, values of the current amplitude are treated in the same way. Note how the current and voltage amplitudes decrease as the wave travels toward the load. The broken lines represent the values of voltage and current amplitude that would exist on a theoretically ideal line. An ideal line does not exist, because it is impossible to construct a line without losses; the concept, however, serves as a comparison reference for all practical cases. Also, under certain conditions, an actual line may approach the performance of a lossless (ideal) line. Such a line has little series resistance or shunt conductance; consequently, there is practically no power loss. As a result, the voltage and current amplitudes will remain constant over the entire length of the line, as indicated by the straight broken lines in parts A and B of figure 13-64. This figure shows, then, that if there were no series resistance or shunt conductance in the line, the transmitted wave would be of constant peak-to-peak amplitude; unfortunately, however, the presence of series resistance and shunt conductance causes the amplitude of a transmitted wave to be attenuated at a constant rate, as shown by the solid line in each part of the figure.

13-303. TRANSMISSION LINE PHASE SHIFT.

13-304. The transmission line also causes a continuously increasing shift in the phase (phase shift) of the transmitted wave as it travels along the line toward the load.

13-305. PHASE VELOCITY.

13-306. When a wave is introduced on a transmission line at the sending end, a very small but definite amount of time goes by before it reaches the receiver. The wave moves along the line with a very high, but definite, velocity. Refer to the sine-wave signal in part A of figure 13-65. Point X represents a definite phase of the wave (in this case, the positive maximum phase). X is not a fixed point, but rather an imaginary one that should be thought of as being attached to the wave itself. This is called the phase velocity, because it is the velocity with which a point representing a phase of the wave moves forward. Further, it is the velocity with which a carrier signal moves along a transmission line. The phase velocity is not the same in all lines; it is determined by the characteristics of a particular line.

13-307. PHASE SHIFT CONSTANT.

13-308. Figure 13-65 illustrates how point X, the peak value of the wave, moves down

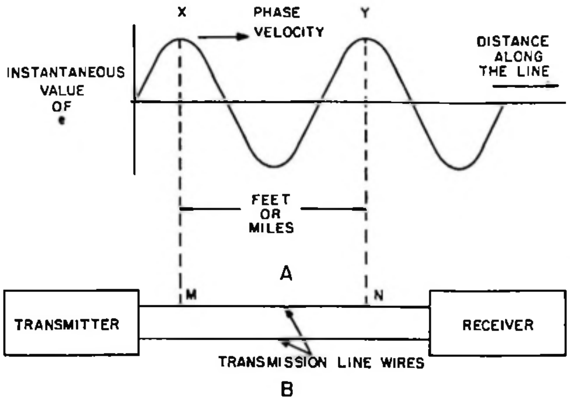


Figure 13-65. Sine Wave Motion

the transmission line. It is important to remember that points X and Y, in part A of the figure, move with the wave, whereas points M and N, in part B, are fixed on the line. The waveform shown in part A represents an instantaneous snapshot at a particular moment of time (that is, a graph of instantaneous voltage at a particular instant, plotted against distance along the line). A short time later, X will occupy the previous position of Y, whereas the instantaneous value of the signal at part M will pass through a complete cycle. Thus, the phase represented by point X will progress from M to N. Meanwhile, Y will move to a new position farther along the line, or, with respect to the transmission line shown in figure 13-65, it will be absorbed in the receiver. This means that, during the period considered, the wave moves forward a distance exactly equal to the distance between two maximum points (that is, the distance between two points of similar phase).

13-309. The distance between these two points of maximum amplitude is called the wavelength of the transmitted wave. The phase difference over this distance is 360 degrees. If it is assumed, for example, that the wavelength is 10 miles, there is a phase shift of 360 degrees over that length of line. In 1 mile, the total phase shift would be 36 degrees for a 10-mile-long wavelength. The phase shift constant (also called phase constant) is the phase shift in degrees per unit length of line. The phase shift constant in the preceding example is 36 degrees per mile.

13-310. PHASE DIFFERENCE.

13-311. If the distributed constants of a line are uniform over the entire length of line, the phase shift constant is also uniform. Furthermore, if the entire length of the line is not an exact multiple of the phase shift per unit length, there will be a phase difference between the sending and receiving ends. For example, if the phase shift constant is 36 degrees per mile in a uniform line 24 miles long, then the total phase shift is 24 times 36 degrees, or 864 degrees. Dividing this by 360 degrees gives two complete cycles of the wave, with a remainder of 144 degrees. This 144 degrees represents the phase difference between the transmitted wave and the received wave.

13-312. VARIATION OF PHASE SHIFT CONSTANT.

13-313. The phase shift constant is determined by both the frequency and the phase velocity of the transmitted wave. The way in which the phase shift constant is affected by the frequency of the transmitted wave will be considered in the following paragraphs.

13-314. In part A of figure 13-66, the sine-wave signal has a frequency of 10,000 cycles per second (cps). Point M on the wave moves along the line to point N with a velocity that is characteristic of the properties of the ideal line. In part B of the figure, the sine-wave signal has a frequency of 20,000 cps. Point X moves down the line to Y with the same velocity that point M moves to point N, since the velocity is fixed by the properties of the line.

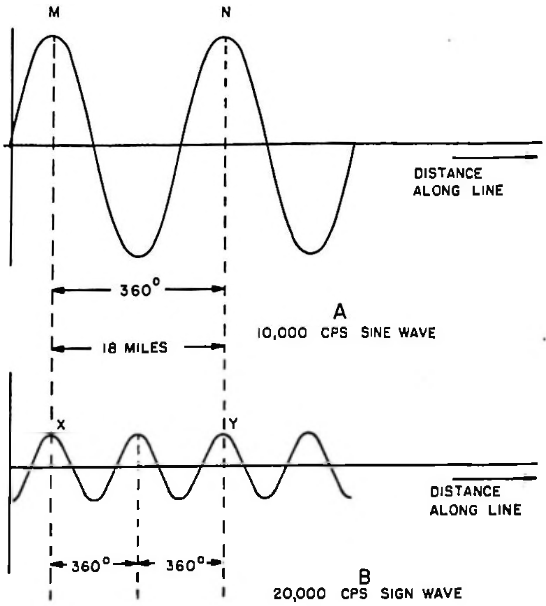


Figure 13-66. Phase Shift Constant

13-315. When the frequency is 20,000 cycles, two cycles occur in the same time that one cycle requires at a frequency of 10,000 cycles. This means that during the time required for one cycle to develop from M to N, as shown in part A of figure 13-66, two cycles will develop over the same distance, from X to Y, as shown in part B of the figure. Assume that end of the distances M to N and X to Y is 18 miles, and that the phase shift constant is 20 degrees per mile. In part B, there is a phase shift of 720 degrees in 18 miles; therefore, the phase shift constant is 40 degrees per mile.

13-316. These results show that when the frequency of the transmitted wave on an ideal line is doubled, the phase shift constant is doubled also. If the frequency is tripled, the phase shift constant is also tripled. On the ideal line, then, the phase shift constant varies directly with the frequency of the transmitted wave.

13-317. The phase shift constant can also be determined by the phase velocity of the transmitted wave. This is shown easily by setting up and examining the relationship as follows:

a. Let V = the phase velocity in number of miles per second.

b. Then:

(1) VB = the number of degrees per second.

(2) 360 degrees x f = the number of degrees per second.

c. For example, if f = 1 you will have 1 cycle per second and, therefore, 360 degrees per second, since there are 360 degrees in each cycle.

d. If f = 2, you will have 2 cycles per second and, therefore, 360 degrees times 2, or 720 degrees per second.

e. Equating (1) and (2), $VB = 360^{o}f$.

Therefore: $B = \frac{360^{o}f}{V}$

where B = phase shift constant
V = phase velocity
f = frequency

13-318. This simple equation immediately shows the relationship between B, f, and V. V depends only on the distributed constants of the line; f is the frequency of the applied signal. For example, if f is a constant, and V increases, B decreases. Furthermore, since the phase velocity, V, is determined by the distributed constants of the line, the phase shift constant, B, is also affected by the distributed constants of the line.

13-319. The equation $B = \frac{360^{o}f}{V}$ also shows that the phase shift constant, B, is proportional to the frequency, f, as previously discussed.

13-320. GROUP VELOCITY.

13-321. In an ideal line, by definition, there is no attenuation, and the phase shift constant, B, is directly proportional to the frequency of the transmitted wave. From the equation $B = \frac{360^{o}f}{V}$, it can be seen that if B is proportional to f, then V must be a constant. (Looking at this conversely, if V is a constant, then B must be proportional to f.) Therefore, on an ideal line, the phase velocity, V, is the same for all frequencies.

13-322. IDEAL LINE.

13-323. Assume that a complex wave composed of more than one frequency moves along a line on which the phase shift constant is directly proportional to the frequency. Such a case is illustrated by two sine waves, f_1 and f_2, in part A of figure 13-67. Adding these waves point by point yields the resultant wave, F. The two component waves, bearing the phase relationship during time t_1, move down the line a distance, d_p. During time t_2, the same phase relationship still exists; that is, the component waves, of different frequency, move down the line at the same velocity. Thus, the resultant is unchanged, since the phase relationship between the two component waves is still the same during time t_2. Thus, the envelope of the wave travels a distance, d_g, which is identical with d_p. The velocity with which an individual wave moves down the line is termed the phase velocity. The velocity with which the resultant, F, moves down the line is termed the group velocity. On the ideal line, where the phase shift constant varies directly as the frequency of the transmitted wave, the resultant and the component waves will have the same velocity. In such a case, there is no separate group velocity; it is equal to the phase velocity.

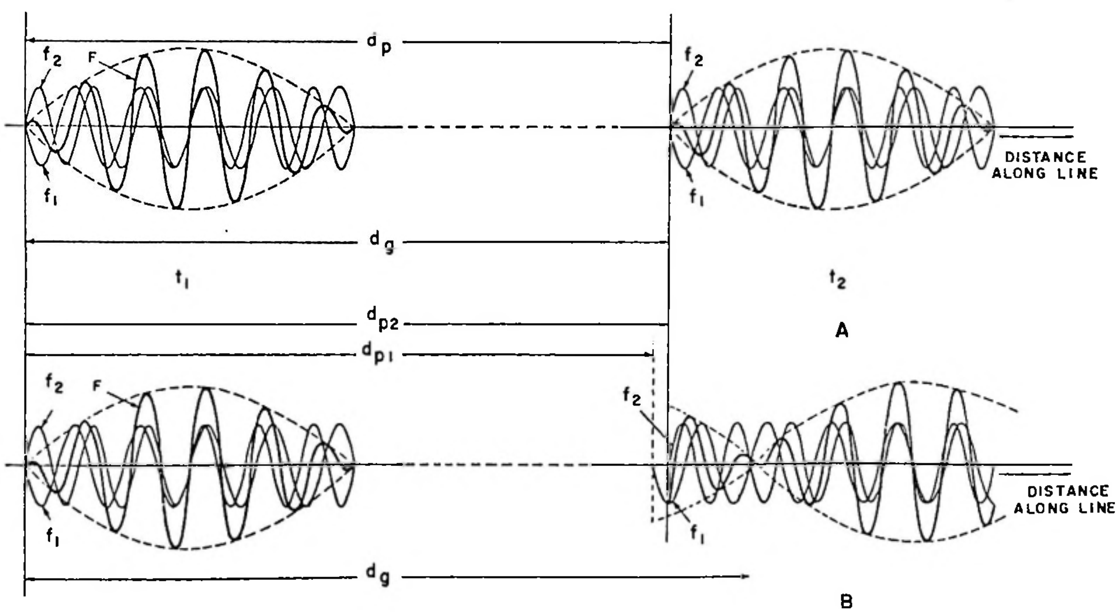


Figure 13-67. Phase Velocity Versus Group Velocity

13-324. ACTUAL LINE.

13-325. Consider the case of an actual, rather than the ideal, transmission line. Because of the distributed constants of the line, the phase shift constant is not directly proportional to the frequency. Therefore, waves of different frequency do not move along the line with the same velocity. Rather, the highest-frequency wave moves with the greatest velocity. This is because of the skin effect, which reduces the inductive reactance, and therefore the inductance, at the higher frequencies. In part B of figure 13-67, the transmitted waves, f_1 and f_2, and the resultant wave, F, start with the same relationships, as shown in part A of the figure. However, during a later time, t_2, wave f_2 has traveled a distance, d_{p2}, but f_1 has traveled a shorter distance, d_{p1}. Thus, the phase of f_1 has gradually dropped behind that of f_2. When f_1 and f_2 are added, the resultant, F, has its maximum amplitude at a point different from that on the ideal line. Its envelope, denoted by the broken line, has progressed a distance, d_g, now greater than d_{p1} or d_{p2}. It has thus moved farther and faster than either of the component waves, f_1 and f_2. The velocity of the envelope is the group velocity of the complex wave. When it is not the same as the phase velocity, the group velocity is always greater than the phase velocity of any component of the complex wave on the transmission line. (For waveguides, the reverse is true.)

13-326. Although the group velocity may be thought of as the velocity of the envelope of the wave, it is not the velocity with which the intelligence is conveyed by a complex signal, such as an amplitude-modulated carrier. The intelligence is conveyed by the sidebands, at a velocity which is the phase velocity of the sideband components. Neither the energy of the wave nor the intelligence it represents can travel faster than the fastest component of the wave.

13-327. It must be pointed out that, for simplicity, the attenuation of the waves has been ignored. Attenuation does occur, however. As a matter of fact, the effects of attenuation and phase shift upon the transmitted signal are combined into a single constant, y, called the propagation constant, where y = oc + jB, with oc and B representing the attenuation and phase shift, respectively. The symbol j represents an angular rotation of 90 degrees counterclockwise. The propagation constant, y, thus describes the effect that the transmission line has on the transmitted wave.

13-328. DISTORTION IN LINE.

13-329. DESIGN CHARACTERISTICS.

13-330. As a wave travels along a transmission line, its amplitude is attenuated and its phase is shifted. For the wave to be transmitted without distortion, it is necessary that each component of the wave be attenuated in the same proportion, and that the phases of the component waves be shifted by amounts directly proportional to their frequencies, no matter what the frequency components may be.

13-331. Figure 13-68 illustrates the motion of an undistorted wave. The dotted line in part A represents a 500-cycle wave with an amplitude of 10 volts. The dot-and-dash line represents a 1500-cycle wave with an amplitude of 8 volts. The solid line represents the resultant algebraically combined wave, which is composed of the 500-cycle and 1500-cycle waves.

13-332. Part B of figure 13-68 represents the same wave after it has passed through a transmission line that has an attenuation ratio assumed to be 2 to 1. As a result, the 500-cycle wave (dotted line) now has an amplitude of 5 volts, as compared with the 10 volts it had before transmission. Likewise, the 1500-cycle wave (dot-and-dash line) is

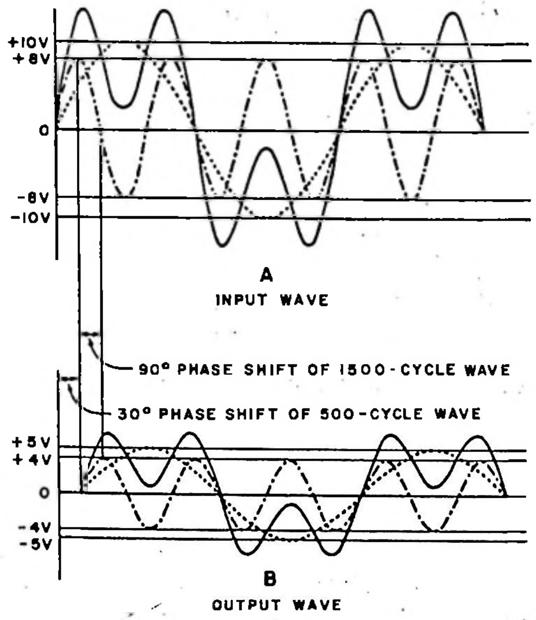


Figure 13-68. Desirable Transmission Conditions

attenuated from 8 volts to 4 volts, which is also a ratio of 2 to 1. Thus, the amplitude of the transmitted wave (solid line) also has been attenuated, but note that its shape remains the same; that is, the wave is not distorted.

13-333. Note also that the 500-cycle wave, shown in part B, has also had its phase shifted by an amount assumed to be 30 degrees. The 1500-cycle wave must have its phase shifted by 90 degrees in order that the phase shifts reflect the same proportion as the frequencies. This is what occurs in this case: $1500:500 = 90^{\circ}:30^{\circ} = 3:1$. The phase of the resultant transmitted wave (solid line) necessarily has been shifted also, but note that the shape of the wave remains the same; it has not been distorted by its motion along the line.

13-334. FREQUENCY DISTORTION.

13-335. In actual transmission lines, not all waves are attenuated the same amount; the attenuation increases as the frequency of the transmitted wave increases. The solid-line curve in figure 13-69 demonstrates how attenuation increases with frequency. This results from the effects of the capacitance, leakage, inductance, and resistance of the line. The lower broken line represents the constant attenuation which is desirable.

13-336. The effect of the variation in attenuation on the shape of the transmitted wave is represented in figure 13-70. The input wave, shown in part A of figure 13-70, is the same as for part A of figure 13-68. Again, after moving along the line some specific distance (see part B of figure 13-70), the 500-cycle wave (dotted line) has an amplitude of 5 volts because it has been attenuated in the ratio 10:5, or 2:1. However, the 1500-cycle wave (dot-and-dash) line now has an amplitude of only 2 volts. Because the attenuation characteristics of the line are not ideal, there has been a greater attenuation; in this example, a ratio of 8:2, or 4:1, has been selected. Thus, the amplitude of the transmitted wave (solid line) has also been attenuated. Note, however, that the shape of the transmitted wave has been changed; that is, the wave has been distorted.

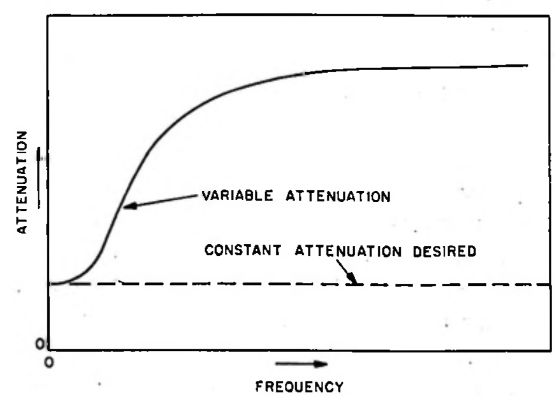


Figure 13-69. Variation of Attenuation with Frequency

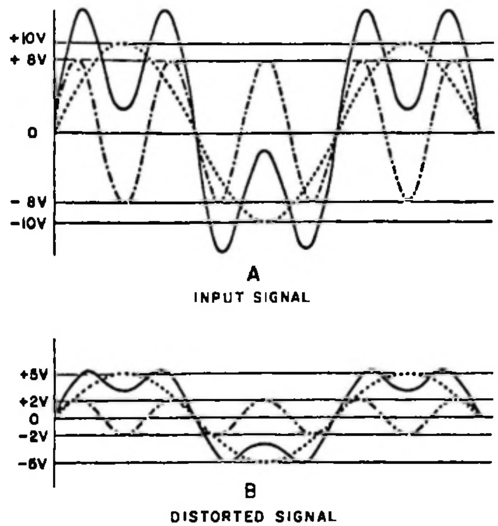


Figure 13-70. Frequency Distortion

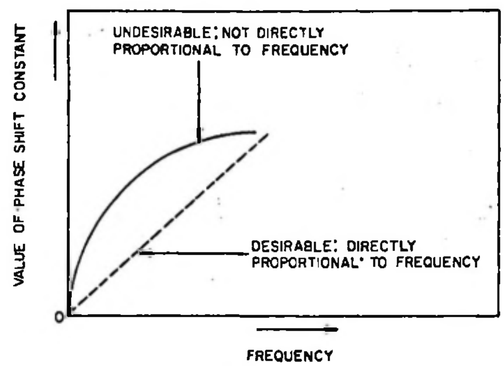


Figure 13-71. Variation of Phase-Shift Constant and Frequency

This type of distortion, which is caused by the variation of attenuation with frequency, is often called attenuation distortion (or, sometimes, amplitude versus frequency distortion).

13-337. PHASE DISTORTION.

13-338. On actual transmission lines, the phase shift constant does not vary directly as the frequency of the transmitted waves, since the phase velocity varies with frequency. The broken-line graph in figure 13-71 shows how the phase shift constant varies in the desirable (or ideal) case; the variation is directly proportional to frequency. The solid-line graph shows how the phase shift constant deviates from the ideal in an actual case; there is more deviation at the lower frequencies than at the higher frequencies. This results from a change in the values of reactance of either the distributed series inductance or the shunt capacitance. These distributed constants affect the phase velocity, V. Since $B = \frac{360^{\circ}f}{V}$, it is evident that the phase shift constant, B, will change as the frequency changes. The effect of this variation of the phase shift constant is illustrated in figure 13-72. Again, the input wave shown in part A of figure 13-72 is identical with that shown in part A of figure 13-68. In part B of figure 13-72, the 500-cycle wave (dotted line) has had its phase shifted 45 degrees lagging. However, since the 1500-cycle wave travels faster than the 500-cycle wave, the 1500-cycle wave lags less than 135 degrees, and in this case the 1500-cycle wave (dot-and-dash line) has had its phase shifted only 90 degrees lagging. The phase of each component of the transmitted wave has been shifted, but not by the desired amount. As a result, the shape of the transmitted wave (solid line) has been changed in its motion along the line; that is, the wave has been distorted. This type of distortion, which results from the fact that the phase shift constant is not directly proportional to the frequency of the transmitted wave, is called phase distortion.

13-339. REFLECTION.

13-340. REFLECTIONS ON AN OPEN LINE.

13-341. CAUSE OF REFLECTION. As has been previously explained, the impedance at

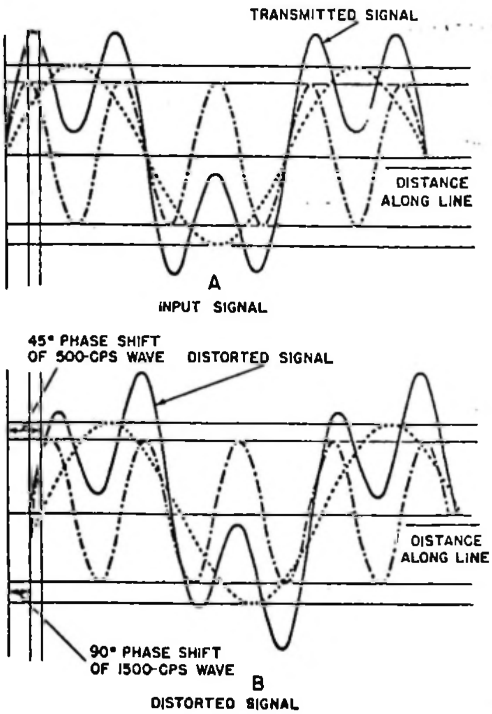


Figure 13-72. Phase Distortion

the output of an open-circuited or open-end line can be considered as infinite. When energy is first applied to the generator end of the line, a wave of current and a wave of voltage sweep down the line. These initial current and voltage waves eventually reach the infinite impedance at the open end of the line. In reaching this open circuit, the current must collapse to zero, because electrons cannot move beyond this point. When the current wave collapses, its magnetic field must also collapse. The collapsing magnetic field crosses the conductor near the end of the line and induces a voltage across the line. This induced voltage acts as a reverse generator, setting up new current and voltage waves, which travel back over the line toward the generator. Waves traveling from the generator toward the load are called <u>incident waves</u>; waves traveling back over the line toward the generator are called <u>reflected waves</u>.

13-342. WATER ANALOGY. Reflection on a line may be compared with the reflection of water waves in a trough. When energy is applied to the water by a flat paddle at one end of the trough, a wave travels toward the opposite end of the trough. If the trough extended to infinity, the waves would travel forever without bouncing back. However, the trough ends at a splash board and the wave of water splashes up on this splash board, raising a certain amount of water above the average level. As the water falls, its energy creates a new water wave, which travels back toward the paddle. The electric-wave action and the water-wave action both produce reflected waves. Since the current becomes a minimum and the voltage becomes a maximum at the output end of an open-circuited line, the resultant voltage and current waves are 90 degrees out of phase. Figure 13-73 shows this 90-degree phase difference between the voltage and current after reflections are set up in an open-circuited line.

13-343. STANDING WAVES OF VOLTAGE.

13-344. Incident and reflected waves produce standing waves, as shown in figure 13-73. Incident voltage and current waves travel down the line with a certain velocity, and the reflected waves travel back at the same velocity. Incident voltage waves are reflected without a reversal of polarity. At each point on the line the incident and reflected voltage waves combine, or add, to produce a resultant voltage which is the sum of the two. The exact manner in which the two waves combine at any point is a function of the distance of that point from the point at which reflection starts. If the resultant voltages are measured and plotted along the line, the curve of voltage values forms a standing wave of voltage. Actually, the voltage at each point (except the zero voltage points) is alternating. However, the rms (root mean square) value of the alternation at any point remains constant. An ac voltmeter, then, reads a series of rms values

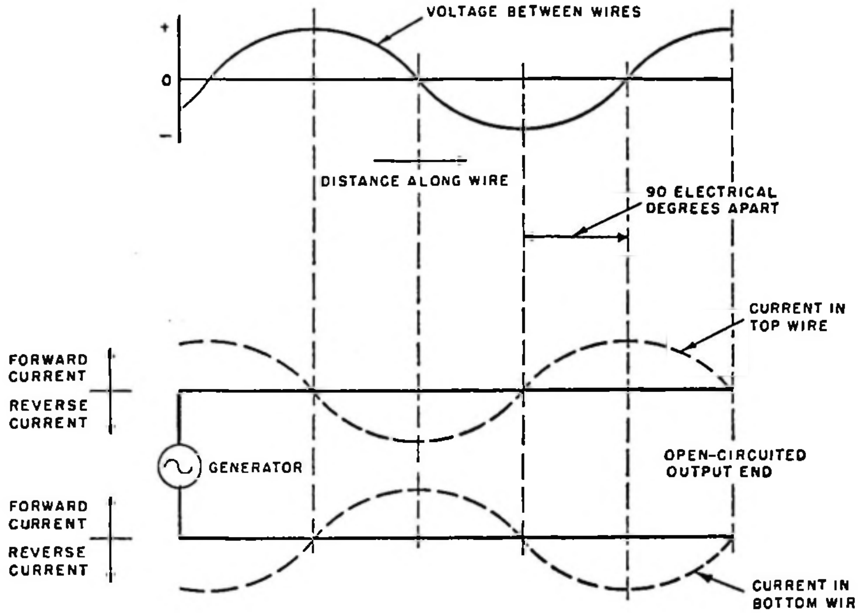


Figure 13-73. Reflected Waves

along the line. When these rms values are plotted, the curve represents a wave which appears to be stationary, or standing, on the line. Figure 13-74 shows a series of different positions of the generated (incident) and reflected waves which are added (heavy lines) to show how the standing waves are produced. Whenever two sine waves are added together, the sum is a sine wave. Therefore, the standing wave is a sine wave.

13-345. STANDING WAVES OF CURRENT.

13-346. A condition similar to that for the voltage also exists for the current, except that current waves are reflected with a reversal of polarity. The incident and reflected current waves combine to produce standing waves of current. The maximum points of the standing waves of current are 90 degrees, or one-quarter wavelength, apart from the maximum values of the voltage standing waves (see figure 13-75).

13-347. REFLECTIONS ON A SHORT-CIRCUITED LINE.

13-348. The voltage and current relations on a short-circuited, or closed-end, line are shown in figure 13-76. Since a short circuit is a condition of zero impedance, the current at the closed end of the line is maximum and the voltage is minimum. Reflection occurs on the closed-end line for the same reason that it occurs on the open-end line; that is, none of the energy in the initial wave is absorbed by the load. The high current in the short circuit represents energy which the zero-impedance load cannot absorb. The only place this current can go is back over the line; consequently, a wave of current is reflected from the load toward the sending end of the line.

13-349. REPRESENTATION OF STANDING WAVES.

13-350. Standing waves of voltage can be measured with an rf voltage indicator. As

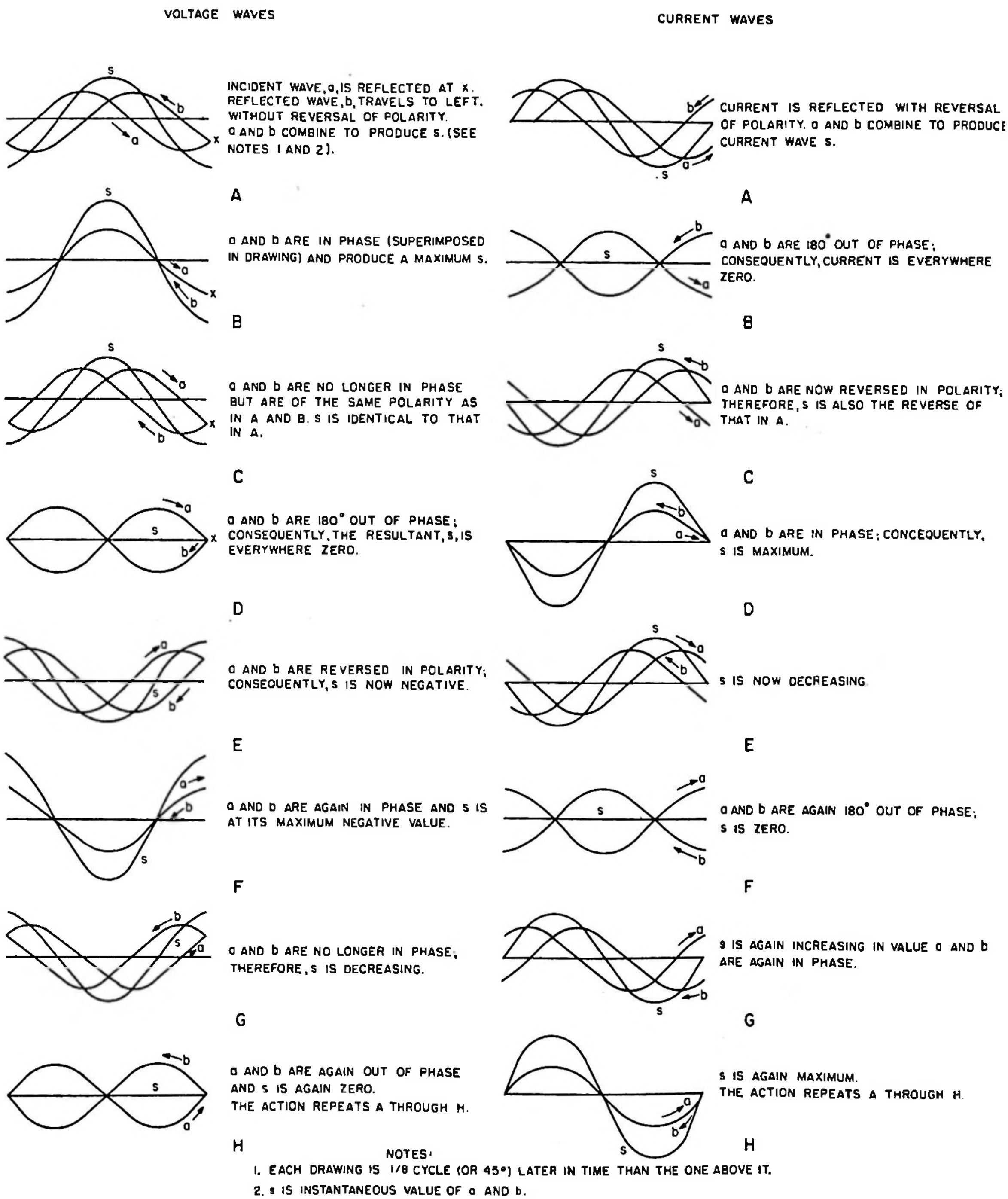


Figure 13-74. Combining of Incident and Reflected Waves To Produce Standing Waves

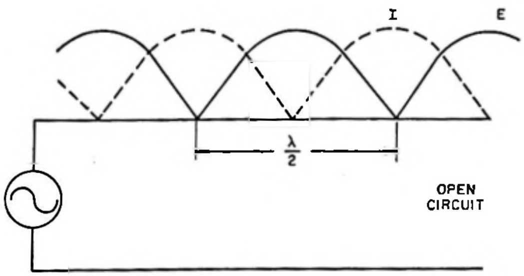


Figure 13-75. Voltage and Current Standing Waves

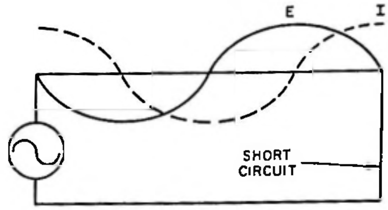


Figure 13-76. Standing Waves on Short-Circuited Line

shown in figure 13-74, high-voltage and low-voltage points do exist. The minimum points are called nodes, and the maximum points are called antinodes. Theoretically, on a line with no losses, the minimum voltage or current is zero. However, there are losses in any real line, and some small amount of current must flow to supply the lost energy. Consequently, no true zero points exist on a real line. Few indicators are arranged to detect whether a standing wave is positive or negative. However, standing waves of voltage and current may be shown on one side of the base line (see figure 13-75). For convenience in drawing, most curves are made to show the voltage between the two conductors, and the current in one of the two conductors. In general, the top conductor (in the drawing) is used as a base line, or reference, for the waves, as shown in figure 13-76.

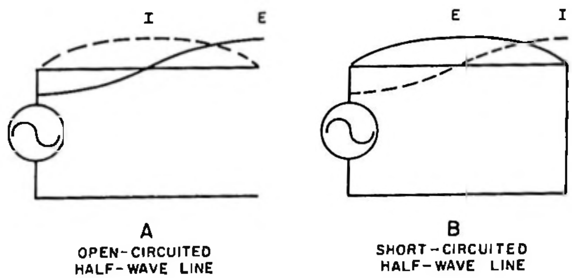


Figure 13-77. Voltage and Current on Open- and Short-Circuited Half-Wave Lines

13-351. COMPARISON OF OPEN-CIRCUITED AND SHORT-CIRCUITED LINES.

13-352. The voltage and current relationships for open and closed lines are the opposite of each other, as illustrated by the half-wave lines of figure 13-77.

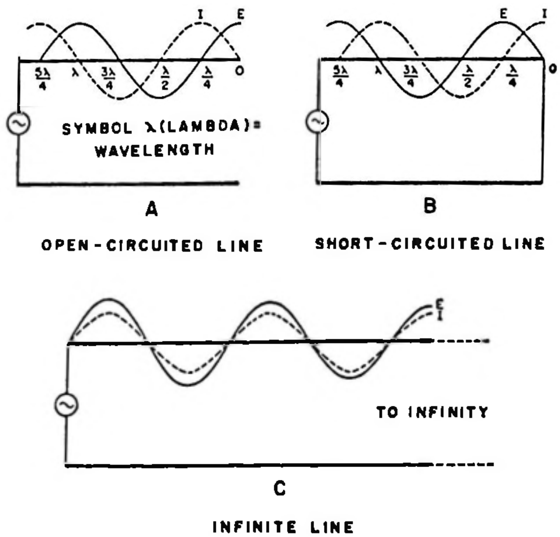


Figure 13-78. Comparison of Open and Closed Line Conditions

13-353. The interrelationships between open-circuited, short-circuited, and infinite-line conditions is shown in figure 13-78. A comparison of parts A and B of the figure shows that, for any length of line, the voltage and current relations in a closed line are the opposite of those in an open line. The points of maximum and minimum voltage and current must be determined from the output end of the line, because reflection always begins at the output end. A line does not have to be any particular length to produce standing waves. Standing waves occur on any line that is not terminated by a resistance equal to the characteristic impedance.

SECTION V

RESONANT LINES

13-354. GENERAL.

13-355. On a line which has standing waves of current and voltage, energy from the generator is surging back and forth along the line, and most of this energy is in the electric and magnetic fields about the line. As waves of current travel along the line, the current points are surrounded by magnetic fields. When the current is reduced to zero at the load end of an open-circuited line, the magnetic field collapses, but the energy cannot just disappear. Because there is no load to absorb it, this energy is transferred to the electric field. The increased electric field is evidenced by an increased voltage at the open end of the line. Energy given to the line by the generator can serve only one of two purposes; it either adds energy to the electric and magnetic fields or supplies line losses resulting from heat and radiation. If there were no losses whatever in the line, the generator could be removed and the line inductance and capacitance would discharge through each other in an oscillatory, self-sustained manner. From ac theory we know that, in the theoretical LC circuit without resistance, this discharge of inductance and capacitance through each other would continue indefinitely, once started, even with the applied voltage removed. With line resistance and leakage of appreciable magnitudes, the line inductance and capacitance discharge through each other against the line impedance, and the oscillations decay with the removal of the applied voltage. This corresponds to the similar decay of oscillation in an ordinary LC circuit with resistance. We can now say that a transmission line can act as a resonant circuit.

13-356. NONRESONANT LINES.

13-357. At this point consider briefly what is meant by a nonresonant line. A nonresonant line may be defined simply as a line which has no standing waves of current and voltage. Such a line is either the theoretical infinite line or a real line which is terminated in a resistance equal to its characteristic impedance. The absence of standing waves is an indication that all of the energy reaching the load is absorbed by the load.

13-358. PRINCIPLES.

13-359. In a section of transmission line of finite length which is not terminated in its characteristic impedance and cannot absorb all of the energy fed into it, reflections will occur. This causes stored energy to exist in the form of standing waves of voltage and current. Since this storage of energy causes the section of transmission line to act as a resonant circuit, it may be put to use like any similar LC combination. A line of any wavelength can produce standing waves; however, to offer a resistive impedance at a particular frequency (become resonant), it must have an electrical length that is some multiple of a quarter wavelength. When this condition is met, the inductive and capacitive reactances cancel and the section behaves as either a series-resonant or parallel-resonant circuit at the applied frequency. Only quarter-wave line sections or some multiple of quarter-wave line sections at the working frequency are considered as resonant line sections.

13-360. COMPARISON OF LINE AND LC CIRCUIT.

13-361. A resonant transmission line has many of the characteristics of a lumped resonant LC circuit composed of coils and capacitors. The more important characteristics that resonant lines have in common with the more familiar lumped resonant circuits are given below.

a. Series resonance.

(1) Resonant rise of voltage across circuit elements.

(2) Low impedance across resonant circuit.

b. Parallel resonance.

(1) Voltage not in excess of applied voltage.

(2) High impedance across resonant circuit.

13-362. DIAGRAMS.

13-363. In studying the applications of resonant transmission lines, you should carefully examine figures 13-79 and 13-80. These illustrations show the relationships of voltage, current, and impedance for various lengths of open- and short-circuited transmission lines. The impedance which the generator sees for various lengths of line is shown directly above the generator on the charts. The curves above the letters of various height indicate the relative values of the impedance presented to the generator as it moves from right to left, and the circuit symbols indicate the equivalent (lumped) electrical values for rf transmission lines of particular lengths. The standing waves of voltage, E, and current, I, applied to the impedance, Z, are shown above each line. These waves are shown on the chart as having the same maximum heights and as going to zero at the same minimum points. The last part of this statement actually would be the case only on an ideal line, that is, a line having no loses. Figures 13-81 and 13-82 also show equivalent circuits of particular lengths of lines which have been taken from figures 13-79 and 13-80.

13-364. RESONANCE IN OPEN LINES.

13-365. ODD QUARTER-WAVELENGTHS.

13-366. The open-end line can be studied with the aid of figures 13-79 and 13-81. At all odd quarter-wave points ($\lambda/4$, $3\lambda/4$, etc) measured from the output end, the current is maximum and the impedance minimum. In addition, there is a resonant rise of voltage from the odd quarter-wave point toward the output end. Thus, at all odd quarter-wave points the open-end transmission line is acting as a series-resonant circuit. Part C of figure 13-81 shows the equivalent circuit for a quarter-wave line or any odd multiple thereof ($3\lambda/4$, $5\lambda/4$, etc). The impedance is a very low resistance, and would be zero if there were no losses in the line.

13-367. EVEN QUARTER-WAVELENGTHS.

13-368. At all even quarter-wave points ($\lambda/2$, $3\lambda/2$, etc), figure 13-79 shows that the voltage is maximum. The voltage on an even quarter-wave line never exceeds the applied voltage. Comparison of the transmission line with an LC resonant circuit demonstrates that at even quarter-wavelengths an open-end line acts as a parallel-

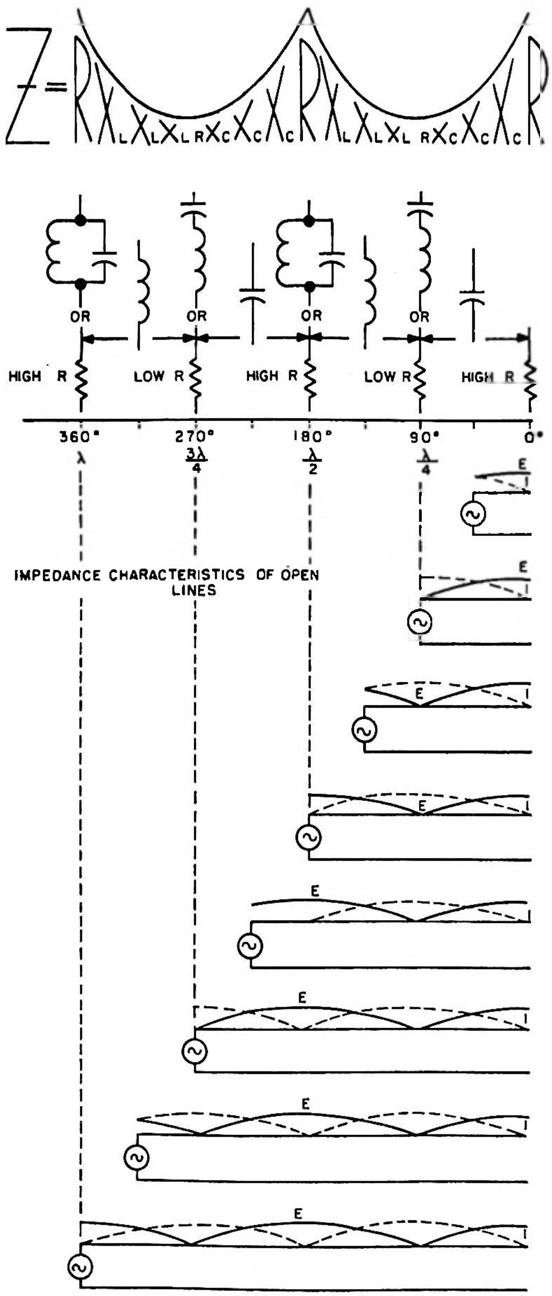


Figure 13-79. Characteristics of Open Lines

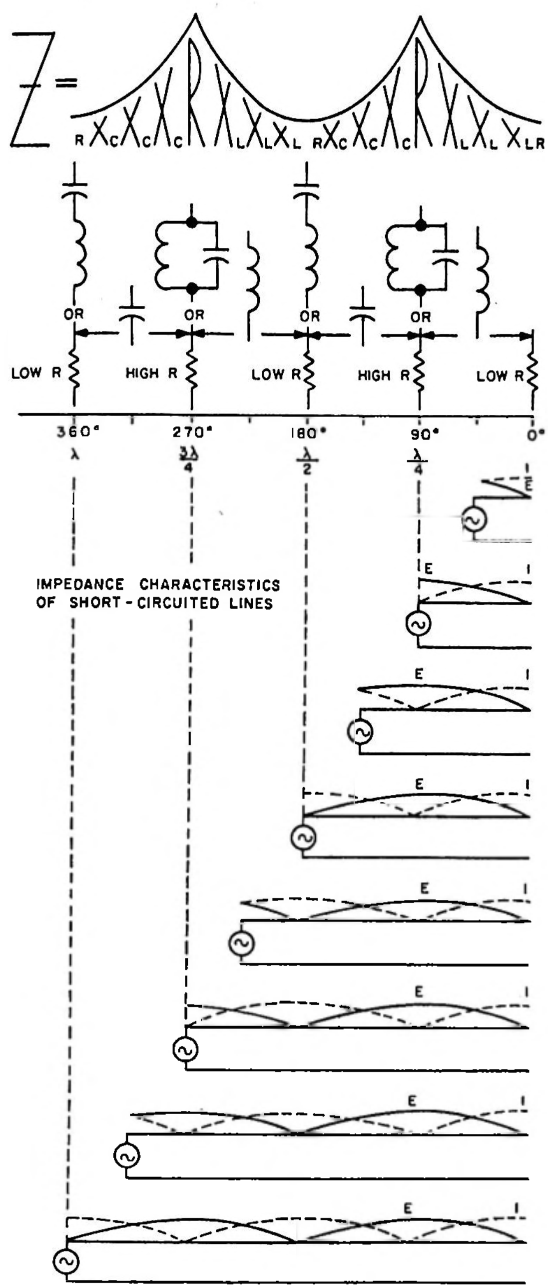


Figure 13-80. Characteristics of Short-Circuited Lines

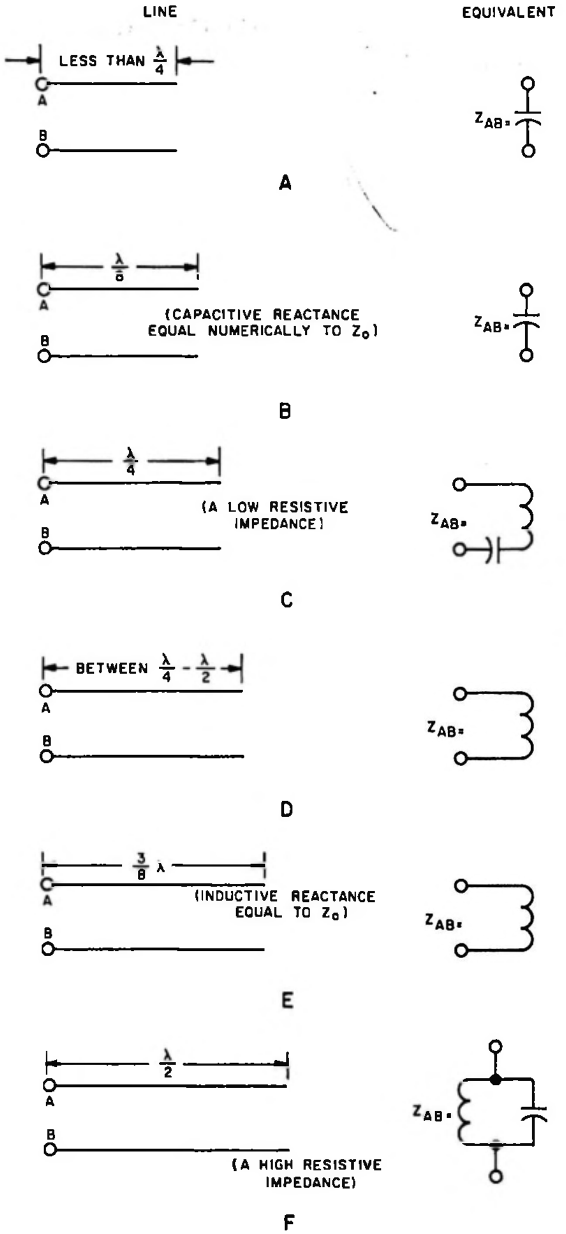


Figure 13-81. Open Lines and Corresponding Lumped Circuits

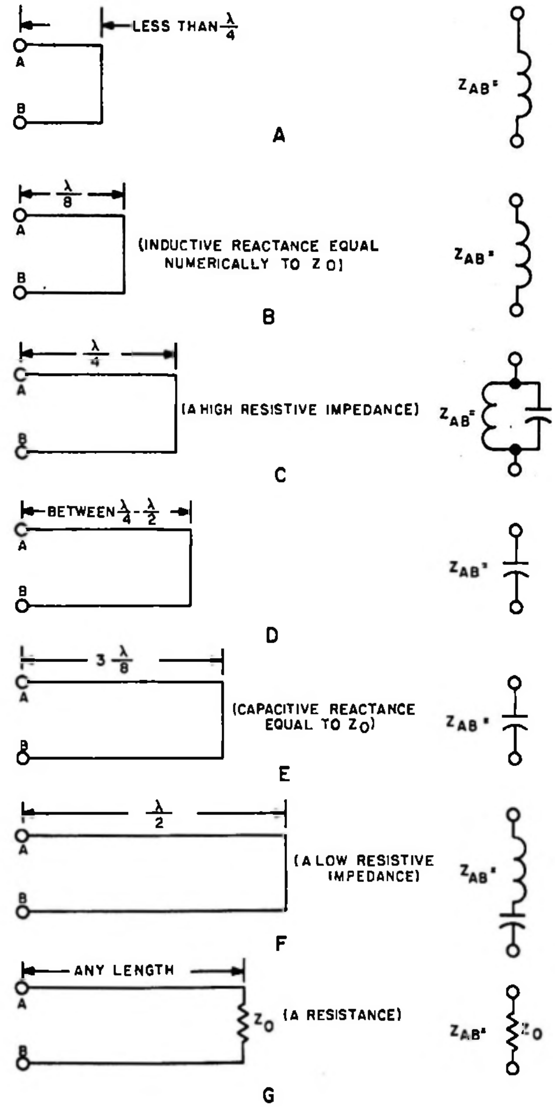


Figure 13-82. Short-Circuited Lines and Corresponding Lumped Circuits

resonant circuit, as shown in part F of figure 13-81, which reflects the equivalent circuit for a half-wave line or any even multiple of a quarter-wave line ($\lambda/2$, λ, $3\lambda/2$ etc). The impedance is an extremely high resistance.

13-369. LINE AS INDUCTANCE OR CAPACITANCE.

13-370. In addition to acting as LC resonant circuits, resonant open-end lines may also act as nearly pure capacitances or inductances. Figure 13-79 shows that an open-end line less than a quarter-wavelength long acts as a capacitance (also see part A of figure 13-81); from $\lambda/4$ to $\lambda/2$ long, as an inductance (also see part D of figure 13-81); from $\lambda/2$ to $3\lambda/4$ long, as a capacitance; and from $3\lambda/4$ to λ long, as an inductance. Part B of figure 13-81 shows that an eighth-wavelength open line acts as a capacitive reactance numerically equal to the characteristic impedance, Z_0, and part E of the figure shows that a three-eighths-wavelength open line acts as an inductive reactance numerically equal to the characteristic impedance, Z_0.

13-371. RESONANCE IN SHORT-CIRCUITED LINES.

13-372. ODD QUARTER-WAVELENGTHS.

13-373. The closed line can be studied with the aid of figures 13-80 and 13-82. At odd quarter-wavelengths the voltage is high, the current is low, and the impedance is high. Also, at no point on such a line does the voltage exceed the applied voltage. Since these conditions are similar to those in a parallel-resonant circuit, the shorted transmission line of odd quarter-wavelength acts the same as a parallel-resonant circuit. Part C of figure 13-82 illustrates the equivalent circuit for a quarter-wavelength shorted line.

13-374. EVEN QUARTER-WAVELENGTHS.

13-375. At the even quarter-wave points, the voltage is minimum, the current is maximum, and the impedance is minimum. Since this action is similar to series resonance in an LC circuit, a shorted transmission line of even quarter-wavelengths acts the same as a series-resonant circuit. Also, looking from any even quarter-wave point toward the output end, there is a resonant rise of voltage to the adjacent quarter-wave point. Consequently, the standing waves of voltage may be of considerably greater amplitude than the applied voltage. Part F of figure 13-82 illustrates the equivalent circuit for a half-wavelength line.

13-376. LINE AS INDUCTANCE OR CAPACITANCE.

13-377. Resonant closed-end lines, like the open-end lines, also act as nearly pure capacitances or inductances. Figure 13-80 shows that a closed-end line of less than $\lambda/4$ acts as an inductance (also see part A of figure 13-82), from $\lambda/4$ to $\lambda/2$ long, as a capacitance (also see part D of figure 13-82), etc. Part B of figure 13-82 shows that a $\lambda/8$ line acts as an inductive reactance numerically equal to the characteristic impedance Z_0, and part E shows that a $3\lambda/8$ line is equal to a capacitance numerically equal to the characteristic impedance, Z_0. In part G of figure 13-82, termination in Z_0 is illustrated. The input impedance in this case is a resistance equal to Z_0.

13-378. OTHER LINE TERMINATIONS.

13-379. EFFECT OF TERMINATING IN A RESISTANCE EQUAL TO Z_0.

13-380. Any line, even if cut to a particular fraction of a wavelength, loses its resonant characteristics when it is terminated in a resistance equal to Z_0.

13-381. Since a quarter-wavelength open-end line has low E and a high I, as shown in figure 13-79, there is a low impedance across points A to B in part A of figure 13-83. Conversely, since a quarter-wavelength closed-end line has a high E and a low I, as shown in figure 13-80, there is a very high impedance across points A to B in part B of figure 13-83. Now if this same quarter-wavelength line is terminated in its characteristic impedance, Z_0, it immediately becomes a nonresonant line and presents an impedance across points A to B equal to Z_0. In figure 13-84, no difference can be detected across terminal points A to B, whether the line is a quarter-wavelength (as in part A) or an infinite line (as in part B). In other words, the quarter-wavelength

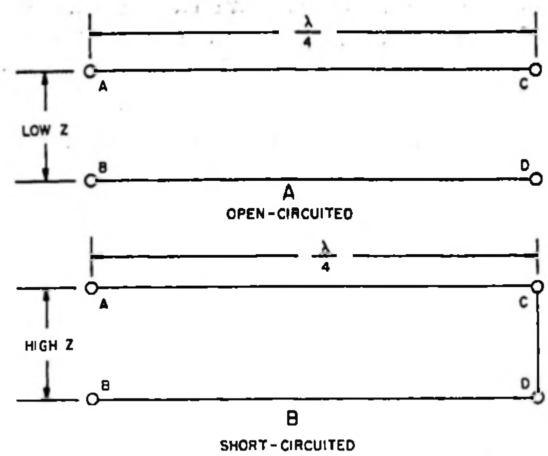


Figure 13-83. Quarter-Wave Lines

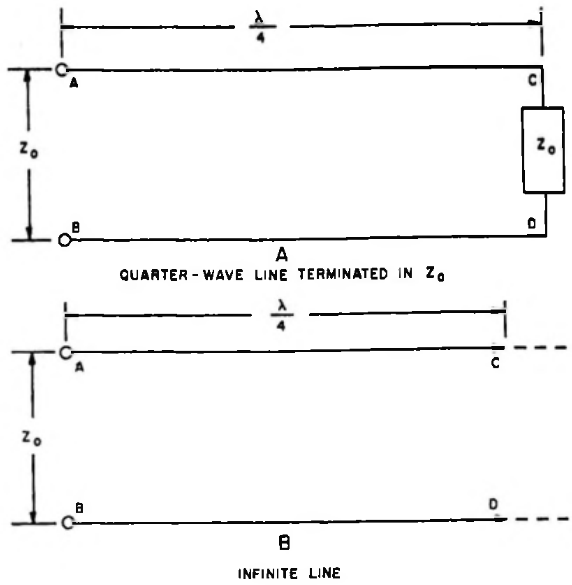


Figure 13-84. Similarity of Quarter-Wave Line Terminated in Z_0 and Infinite Line

line in part A has none of the characteristics of a quarter-wavelength resonant line (see figure 13-83); it is just a nonresonant line.

13-382. LINE TERMINATED IN A REACTANCE.

13-383. A line terminated in a resistance equal to its characteristic impedance has no reflections present. However, if a line is terminated in a reactance of any value, standing waves are present. Part A of figure 13-85 shows the standing waves on a line terminated in a capacitive reactance equal to the characteristic impedance. Part B of figure 13-85 shows the standing waves on a line terminated in an inductive reactance equal to the characteristic impedance.

13-384. With a capacitive-reactance load, as shown in part A of figure 13-85, the first minimum point of voltage is closer than a quarter-wavelength to the output end of the line. Similarly, in part B of figure 13-85, the first minimum point of voltage is more than a quarter-wavelength from the output end of the line. With capacitive termination the voltage and current distributions has essentially the same characteristics as with the open-circuit output, except that the curves

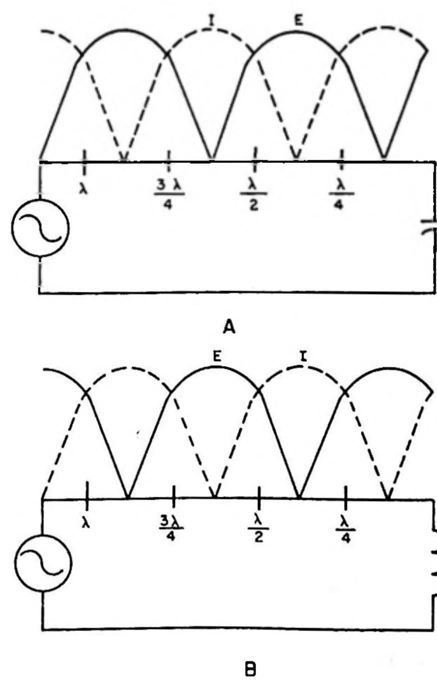


Figure 13-85. Standing Waves Produced by Terminating Line in Reactance Equal to Z_0

are shifted toward the output end of the line by an amount that increases as the capacitive reactance is decreased, that is, as the line approaches the closed-line condition.

13-385. With inductive termination the voltage and current distribution has essentially the same characteristics as with the short-circuited output, except that the curves are shifted toward the output and by an amount that increases as the load reactance approaches infinity, that is, as the line approaches the open-line condition.

SECTION VI

TYPES OF LINES

13-386. GENERAL.

13-387. Transmission lines are usually classified according to their construction; a distinction is made between open-wire lines and the different types of cables. It is important for you to know where these different types of transmission lines are used. You should also understand the construction and characteristics of different types of lines, and realize the correlation between line characteristics and individual application. The four general types of transmission lines are:

a. The open two-wire or parallel-conductor line

b. Cables or shield pair

c. Concentric (coaxial) line

d. The twisted pair or field wire

13-388. OPEN TWO-WIRE LINE.

13-389. CONSTRUCTION OF OPEN TWO-WIRE LINE.

13-390. An open two-wire transmission line consists of two parallel conduction wires maintained at a fixed distance by means of insulating spacers or spreaders at suitable intervals, as shown in figure 13-86. This line is used because of its ease of construction, its economy, and its efficiency. In practical applications, open-wire transmission lines are used for commercial power lines, telephone lines, telegraph lines, and connecting links between an antenna and a transmitter or an antenna and a receiver.

13-391. The choice of spacing between the two conductors of a pair, and between separate pairs, depends on the carrier frequency. The higher the frequency, the smaller the spacing between the conductors of a pair, and the greater the spacing between pairs. The conductors are usually made of copper, or an alloy of copper and steel.

13-392. ATTENUATION IN OPEN TWO-WIRE LINE.

13-393. Signal attenuation produced by open-wire lines is very low as compared with that produced by cables and field wires; it varies between .035 and .085 db (decibel) per mile, depending on the frequency, the wire size, the weather conditions, and the type of circuit. As the frequency increases from 500 to 5000 cycles, the amount of attenuation increases by about 50 percent. However, the signal attenuation increases by about 200 percent when the frequency increases from 5000 cycles to 50,000 cycles. At a carrier frequency of 140 kc, the amount of attenuation is approximately 0.38 db per mile. Loading (a method for reducing attenuation by increasing the line inductance) is never used with open-wire lines, because adverse weather conditions affect open-wire loading detrimentally. The attenuation char-

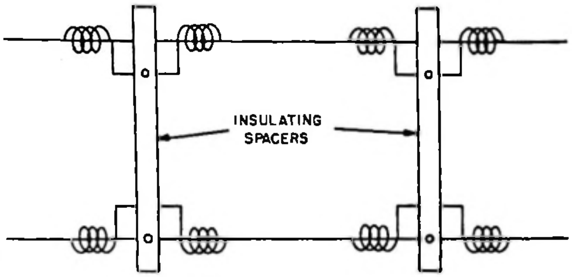


Figure 13-86. Construction of Two-Wire Line

acteristics of the line change with weather (the amount of attenuation increases in wet weather). Repeaters, which amplify the signal, are placed along the line when it is desired to increase the length of the telephone circuit. Because the attenuation on an open-wire line is low, the repeaters can be spaced relatively far apart, making the open-wire line a less expensive form of transmission equipment to install. Maintenance, however, is more expensive to perform than it is for cable, especially under combat conditions. Table 13-1 lists different types of open-wire lines, with their various characteristics.

13-394. SHIELDED PAIR.

13-395. CONSTRUCTION OF SHIELDED PAIR.

13-396. The shielded pair, shown in figure 13-87, consists of two parallel conductors separated from each other and surrounded by an insulating dielectric material, such as plastic copaline. The conductors are contained within a copper-braid tubing, which acts as a shield. The assembly is covered with a rubber or flexible composition coating to protect the line against moisture and friction. Outwardly, it looks very much like an ordinary power cord for an electric motor.

13-397. The outstanding advantage of the shielded pair is that the two conductors are balanced to ground; that is, the capacitance between each conductor and ground is uniform along the entire length of the line, and the wires are shielded against pickup of stray fields. This balance is effected by the

Table 13-1. Open-Wire Lines and Their Characteristics
(Based on Standard Spacing of 8 Inches)

WIRE SIZE	DC RESISTANCE (ohms per loop mile)	CAPACITANCE (μf per loop mile)	ATTENUATION AT 1 KC (db loss per mile)
#12 gage (104)	10.15	0.00905	0.070
#10 gage (128)	6.74	0.00944	0.049
#8 gage (165)	4.11	0.00996	0.032

Notes:
1. All values are for dry-weather conditions.
2. All measurements are made at 68^{0} Fahrenheit.
3. Parenthetical number in WIRE SIZE column is diameter of wire in thousandths of an inch.

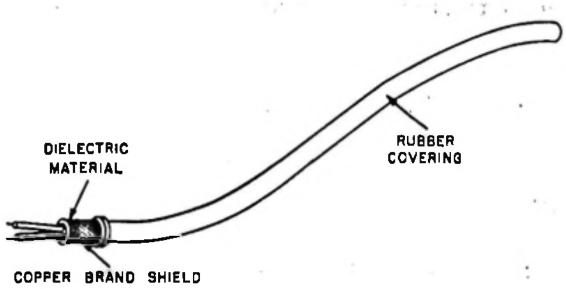


Figure 13-87. Shielded Pair

grounded shield which surrounds the conductors at a uniform spacing throughout their length. If radiation from an unshielded line is to be prevented, the current flow in each conductor must be equal in amplitude in order to set up equal and opposite magnetic fields which will cancel out. This condition may be obtained only if the unshielded line is well in the clear of all obstructions, and if the distance between the wires is small. If an unshielded line runs near either a grounded or conducting surface, one of the two wires is closer to the obstruction than the other. The capacitance between the closer conductor and the conducting surface is greater than the capacitance between the farther conductor and the conducting surface. These unequal capacitances act as unequal conducting paths for each half of the line, causing a greater current in the conductor closer to the nearby conducting surface. Because of these unequal line currents, radiation from the line is increased. The shielded line reduces these undesirable line radiations by maintaining balanced capacitances to ground.

13-398. CONSTRUCTION OF CABLES.

13-399. Although it is customary to distinguish between lead-covered, rubber-covered, and submarine cables, the three types are basically the same in construction. Some cables, although not all, contain insulated conductors in the form of quads. The cable conductors are transposed by twisting the two wires of each pair together and then twisting the pairs of each group of four wires together to form quads, and by spiraling the quads in opposite directions about the core of the cable. The number of quads varies from seven (14 pairs) to 150 (500 pairs) per cable, or more. American-manufactured cables use the multiple-twin quad or bell quad, which consists of two twisted pairs, twisted together in a long twist. Cables of foreign manufacture use either the multiple-twin quad or a spiral-four quad, also called a star quad. This type consists of four wires laid together and twisted as a group in a long twist, the diagonally opposite wires being used as a pair. The conductors are usually either 19- or 16-gage solid copper. Some long circuits use each pair of a quad as a single conductor to decrease the signal attenuation. On short local circuits you should use each pair of a quad as a single conductor to decrease the signal attenuation. However, on short local circuits, two-wire operation is permissible, each wire of a pair being used as a separate conductor. Table 13-2 lists various types of cables, with their characteristics.

13-400. LEAD-COVERED CABLE. You can use this type of cable in a permanent installation to provide a comparatively large number of voice-frequency circuits. It may be laid above ground or buried. Other applications of this type include use as an entrance cable between the end of an open-wire line and office equipment, or as an intermediate cable between two sections of an open-wire line. The cable consists of many quads (up to 150 or more) encased in a flexible lead pipe. The close spacing between the conductors of a pair causes excessive attenuation. The attenuation varies between 2.0 and 3.8 db per mile at voice frequencies, depending on the size of the wire. The use of loading coils decreases the attenuation so that it varies between 0.15 and 0.5 db per mile. At carrier frequencies, it is not

Table 13-2. Cable Conductors and Their Characteristics

SIZE OF CABLE CONDUCTOR	DC RESISTANCE (ohms per loop mile)	CAPACITANCE (μf per loop mile)	ATTENUATION AT 1 KC (db loss per mile)
#26 gage	440	0.069	2.67
#24 gage	274	0.072	2.14
#22 gage	171	0.082	1.79
#19 gage	86	0.084	1.26

customary to use loading. Instead, closely spaced repeaters are used along the line, to make up for the relatively large amount of attenuation.

13-401. RUBBER-COVERED CABLE. In this type of cable, the insulation around the individual wires is rubber, or a rubber substitute, as is the covering over the quads. Since rubber-covered cables are comparatively light and flexible, they may be installed aerially, laid on the ground, or buried. The use of rubber insulation prevents the absorption of moisture which would occur with paper insulation, with consequent increase of attenuation. The cable assembly diagramatically illustrated in figure 13-88 is an example of rubber-covered cable. It consists of four rubber-insulated, stranded copper conductors twisted to form a spiral-four quad. The rubber insulation of one pair of conductors is white, and the rubber insulation of the other pair is colored. A basket-weave steel-wire braid covers the paper, and a rubber jacket covers the entire core. The cable is available in 1/4-mile lengths, with male and female connectors at opposite ends for rapid interconnection. Each connector assembly contains a 6-millihenry loading coil. This cable is used for four-channel carrier operation over intermediate distances; it may also be used on spaced poles over distances up to 150 miles.

13-402. Spiral-four Cable Assembly CX-1065/G is made from spiral-four Cable WM-8/G, which is a recent development. WM-8/G cable is smaller and lighter than previously made spiral-four cable, and is also capable of use over wider frequency and temperature ranges. In the construction of WM-8/G cable, plastic materials have been used instead of rubber for the insulation and the jacket (see figure 13-89). Four stranded-copper conductors are insulated separately with polyethylene and cabled around a polyethylene core. The cabled conductors are covered by an inner jacket of polyethylene, a carbon-cloth stabilizing tape, a stainless steel braid, and a thermoplastic outer jacket.

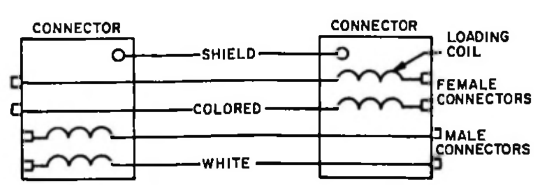


Figure 13-88. Rubber-Covered Cable Assembly

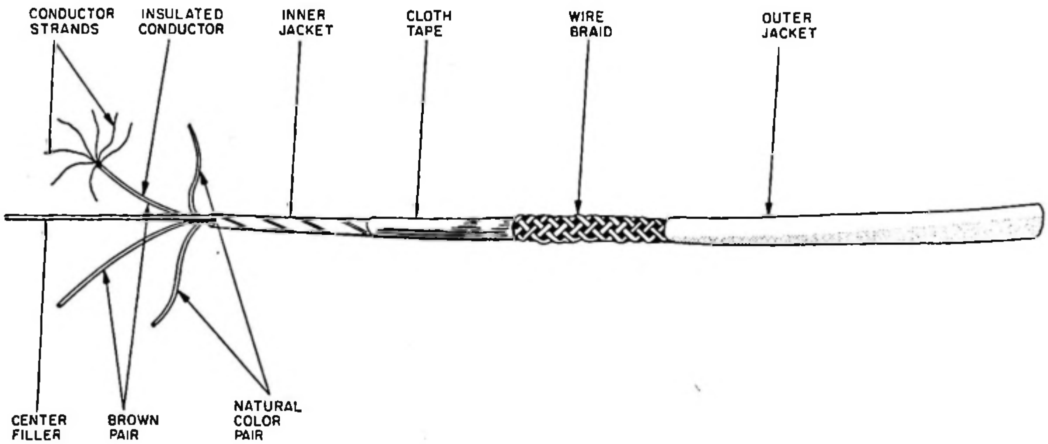


Figure 13-89. Plastic-Covered Cable, Showing Construction Details

13-403. Cable Assembly CX-1065/G consists of approximately a 1/4-mile length of Cable WM-8/G with a universal connector at each end. Loading coils are not built into the connectors as they were in earlier versions of spiral-four cable. Instead, Telephone Loading Coil Assembly CU-260/G may be inserted between the connector faces when required by the carrier equivalent design.

13-404. Rubber-covered cables also come in 5- and 10-pair types for use in voice-frequency communication. An example of 5-pair rubber cable is shown in figure 13-90. These cables are used principally for short loops where a number of circuits are required. They are available with or without connectors at the ends. The conducting, or semiconducting, sheath about the cable stabilizes it. Cables are said to be non-stabilized if their distributed characteristics (capacitance, conductance, and resistance) increase by large amounts from dry to wet weather. The sheath reduces the effects of these changes. Five- and ten-pair cables are fairly stable, when in good condition, if water does not reach the conductors.

13-405. SUBMARINE CABLE. The submarine cable is used as an insert in either open-wire lines or cable when the line must traverse a body of water too wide to be spanned between poles. The submarine cable consists of paper-insulated conductors, lead-covered and wire-armored (that is, encased in a wire mesh). It is suitable for use in water to depths of 250 feet without danger of collapse of the sheath. Excessive water cannot penetrate along the cable, even though the sheath should fail at some point. At depths greater than 250 feet, there is danger of gradual collapse of the sheath; also, excessive water pressure may push the wires through the insulation, causing gradual failure of the cable. However, submarine cables are also made with extra-

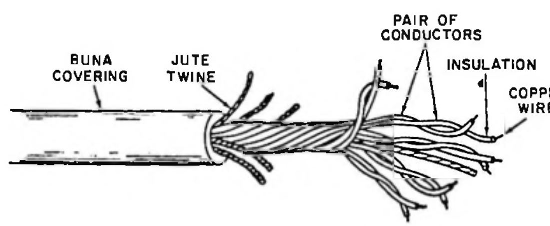


Figure 13-90. Five-Pair Rubber Cable

dense cores and thicker sheaths for use in depths up to 1000 feet. A rubber-covered cable may also be used for submarine applications if there is no danger that the cable will be damaged by excessive abrasion. When a rubber-covered cable is used under water, it is important that all connections and splices be made watertight.

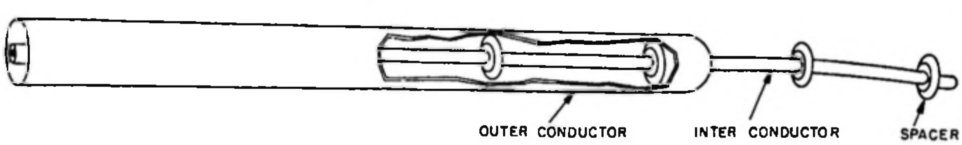
13-406. CONCENTRIC LINE.

13-407. CONSTRUCTION OF CONCENTRIC LINE.

13-408. The concentric or coaxial line has advantages which make it very usable for efficient operation at high frequencies. It consists of a wire within a tubular outer conductor, as shown in figure 13-91. In some cases the inner conductor is also tubular. The inner wire or conductor is insulated from the outer conductor by insulating spacers or beads spaced at regular intervals. The spacers are made of pyrex, polystyrene, or some other material having good insulating qualities and low loss at high frequencies. Flexible coaxial cables also are made with the inner conductor consisting of flexible wire insulated from the outer conductor by a solid and continuous insulating material. Greater flexibility is gained by using a metal braid as the outer conductor. The losses in this type of line are higher than those in a well-designed tubular type of concentric line.

13-409. The electromagnetic fields about a two-wire open line extend into space; consequently, radiation losses may be high and the lines are likely to peak up interference from external sources. In a coaxial line, however, the electromagnetic fields are confined to the space between the inner and outer conductors; that is, the coaxial line is almost a perfectly shielded line.

13-410. Because of the near perfect shielding, there is very little radiation loss from a coaxial line, and the line may be run close to, or attached to, metallic objects without causing any appreciable loss. In fact, one of the principal advantages of the coaxial line is that it may be laid over the most convenient route without increasing the losses due to attenuation within the line. The greatest disadvantage of a coaxial line is its relatively high attenuation, particularly at higher frequencies. The interior of tubular lines must be kept dry to prevent excessive leakage between the conductors. To prevent the condensation of moisture within them, tubular lines may be filled with dry nitrogen at pressures ranging from 3 to 35 pounds per square inch. The nitro-



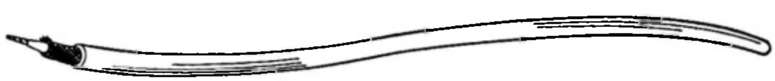


Figure 13-91. Construction of Concentric (Coaxial) Line

gen is used to dry the interior of the line, and the pressure is maintained to insure that any nitrogen leakage will be outward.

13-411. CONCENTRIC CABLE ASSEMBLY.

13-412. Figure 13-92 shows a typical concentric or coaxial cable assembly, containing two (there can be four or even eight) coaxial cables, in addition to several conventional pairs used for supervisory purposes. In carrier service, as many as 480 two-way conversations may be carried simultaneously on any one such line.

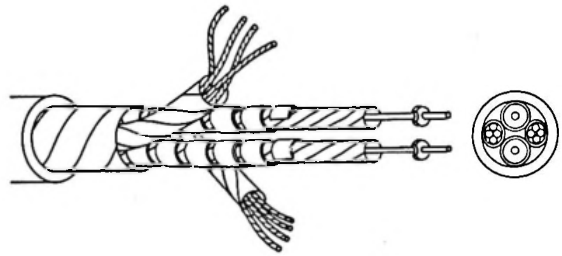


Figure 13-92. Typical Concentric (Coaxial) Cable Assembly

13-413. TWISTED PAIR.

13-414. CONSTRUCTION OF TWISTED PAIR.

13-415. The twisted pair, as the name implies, consists of two insulated wires twisted to form a flexible line without the use of spacers (see figure 13-93). It is generally used as an untuned (nonresonant) line for low-frequency transmission. It is not used at the higher frequencies, because of the high losses occurring in the rubber insulation. Its chief advantage is that it may be used over short distances where more efficient lines would not be feasible because of mechanical considerations.

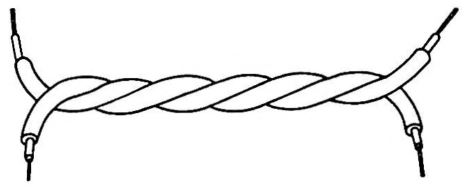


Figure 13-93. Twisted Pair

13-416. FIELD WIRES.

13-417. Field wires are a special type of twisted pair designed for use in situations where communication lines must be set up at great speed. Desirable transmission properties are sacrificed in favor of light weight and strength. The line consists of a twisted pair made of strong, light conductors insulated with rubber or a rubber substitute. Signal attenuation in this type of line is high (1.5 db per mile if the wire is wet, at a frequency of 1000 cycles per second). Field wire lines, in general, are operated both without loading and without repeaters, so that their use is restricted to short distances. The capacitance, leakage, and loss of this wire increase considerably with a change from dry to wet weather. The usable range may be increased, in some cases, by the use of loading or voice-frequency repeaters. Repeater design for such lines is difficult, however, because of the sharp increase in attenuation when the wire becomes wet. Therefore, field wires are used for voice-frequency circuits where an appreciably long talking range is not required. Figure 13-94 shows field wire type WD1/TT, and gives some details concerning its construction.

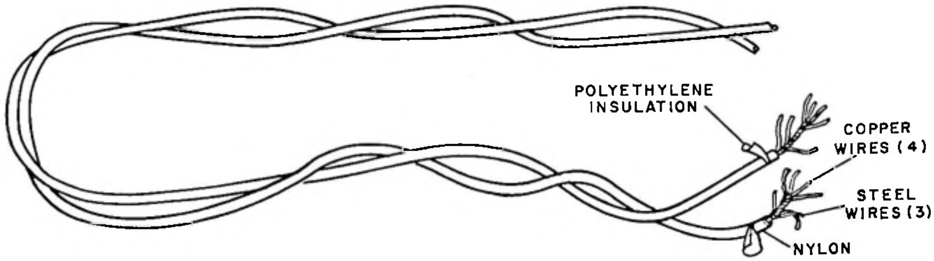


Figure 13-94. Field Wire

SECTION VII

TRANSMISSION LINE PRACTICE

13-418. LINE PROBLEMS.

13-419. ATTENUATION IN LINE.

13-420. The primary purpose of a transmission line is to transfer maximum power to the receiver at the receiving end of the line. Achievement of this purpose involves two separate factors. First, the line itself must absorb as little power as possible, delivering maximum power to the receiving end. Second, the power that arrives at the receiving end should all be absorbed by the load, and there should be no reflections back along the line. The former requirement is met by keeping signal attenuation, due to the line, at the lowest possible value. The latter is met by proper matching of the load at the line.

13-421. INTERFERENCE IN LINE.

13-422. In communication circuits, it is also necessary that the transmission line transmit only the desired signal. Interfering signals, represented by reflections back along the line, may distort the desired signal or even override it so that it is rendered unintelligible at the receiver.

13-423. CAUSES OF ATTENUATION.

13-424. The characteristics of a transmission line which attenuate the signal may be summarized as follows:

a. A mismatch between the impedances of the transmitter and receiver and the characteristic impedance of the transmission line will reduce the amount of power transferred from the transmitter to the receiver.

b. Reflection of power because of a mismatch at the joint of two dissimilar line sections will result in a decrease in the power transferred to the receiver.

c. Power losses will occur in pieces of equipment that are part of the communication circuit. Such losses are not really line losses, but they may be reduced by proper choice and maintenance of the line. The power losses due to mismatches between the line and the equipment may be reduced to a minimum by careful matching; losses in the equipment itself cannot be avoided.

d. The attenuation per unit length of line and the length of the line determine the total line attenuation. To this must be added the losses inherent in the equipment, in order to obtain the total attenuation of the circuit.

e. If the characteristic impedance of the line varies with frequency, the match between the line and the receiver is not constant; thus, there is a variation in the power transfer.

f. Frequency distortion reduces the power transfer by amounts varying with frequency, reducing the power of the higher frequencies more than that of the lower frequencies. Consequently, a signal containing components of different frequencies will have a distorted waveshape. Obviously, these power losses may be reduced only by altering the characteristics of the transmission line.

13-425. LOADING OF LINE.

13-426. The loading of a transmission line changes its characteristics and thus reduces the total attenuation, causing the characteristic impedance and attenuation per unit length to remain substantially constant over a desired frequency range.

13-427. EFFECT OF LOADING.

13-428. Which of the four distributed constants of the transmission line can be changed for loading purposes? Reducing the distributed resistance lowers the attenuation, but it may be accomplished only by increasing the size of the wires. This, in turn, increases the amount of copper used and thus decreases the number of circuits per given cable size. Reducing the distributed capacitance by increasing the separation between conductors reduces the attenuation, and this, in turn, also reduces the number of circuits per cable. The solution of the problem lies with the fourth constant, the series inductance. In practice, the value of this inductance is increased deliberately, whereas the other three constants are kept as small as possible. This corrective measure is called loading. The effect of loading on attenuation is illustrated in the following example: In part A of figure 13-95, assume that 5 watts is the input power, P, to a transmission line having a characteristic impedance of 500 ohms. The ratio of input voltage to input current must be equal to the characteristic impedance, because the line is terminated with an impedance equal to the characteristic impedance of the line. As shown, the input voltage must be 50 volts, and the input current 0.1 ampere. Now, assume that the characteristic impedance of the line is increased by the inductive loading, as indicated in part B of figure 13-95. The characteristic impedance is now 600 ohms. Since the input power is held constant, the input voltage is now 54.8 volts, and the input current 0.091 ampere. This reduction in

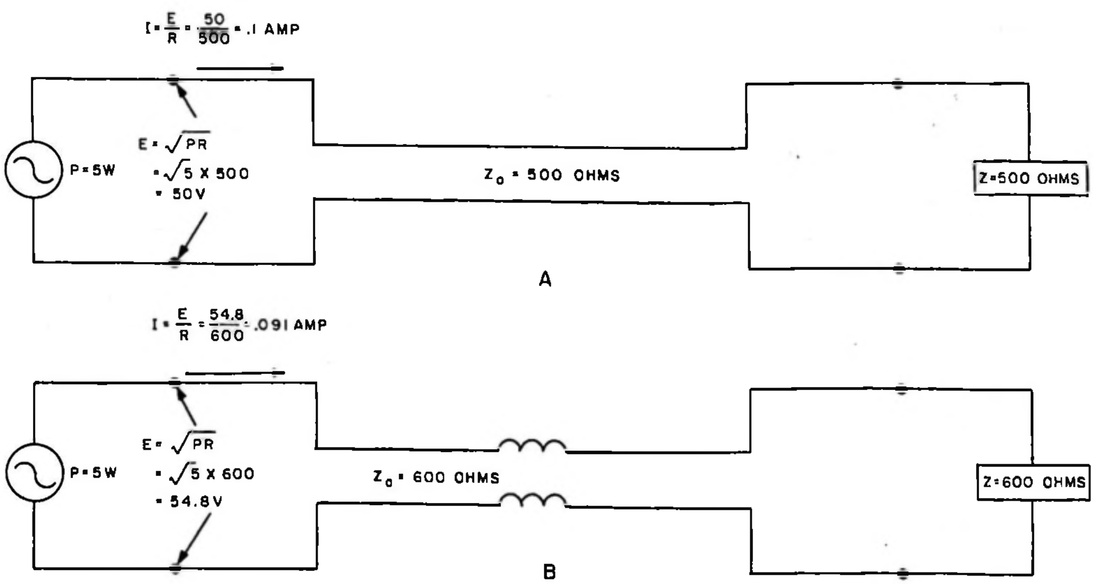


Figure 13-95. Effect of Loading

current decreases the I^2R losses along the line, so that if loaded and unloaded circuits receive the same amount of input power, more power is delivered to the receiver by the loaded circuit than by the unloaded circuit.

13-429. Figure 13-96 shows that the attenuation over a wide frequency range is much less with loading than without loading. The same curves show that with loading the attenuation is practically constant over a wide range of frequencies, whereas without loading the attenuation increases as the frequency increases. With loading, therefore, frequency distortion is virtually eliminated. Mathematically, it may be shown that loading causes both the attenuation and the characteristic impedance to be substantially constant. With the characteristic impedance of the line constant, there is no variation in match between the receiver and the transmission line. Hence, the power transferred to the receiver is constant, and is maximum over a wider range of frequencies.

13-430. METHODS OF LOADING.

13-431. Loading may be accomplished in either of two ways. (Both methods must provide an evenly distributed inductance to give the effect of a transmission line with distributed inductance.) In the first method,

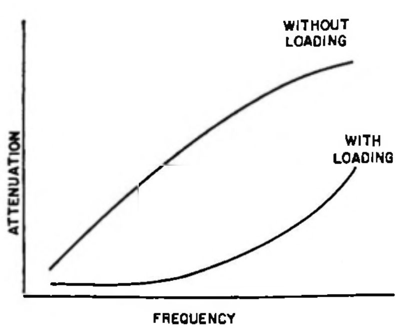


Figure 13-96. Graph Showing Effect of Loading on Attenuation

which provides uniform loading, the complete length of line is wrapped with a tape of ferrous material, such as iron or permalloy. This method is expensive. The other method uses loading coils spaced at equal intervals along the line.

13-432. LOADING COILS.

13-433. Loading coils are so spaced that the change produced in a signal over a given distance differs little from the change that is produced over the same distance by a uniformly loaded circuit. The spacing must be such that there will be several coils per wavelength of signal; otherwise, the attenuation will increase rather than decrease. The coils are wound on toroidal cores, which are made of powdered soft iron or powdered permalloy, with a binder to hold the powdered metal together. The air pockets in the powdered metal reduce the possibility of saturation of the core, which would distort the signal.

13-434. LOADING LIMITATIONS.

a. The number of loading coils cannot be increased indefinitely; this would cause an increase in the series resistance, which would cancel the beneficial effects of loading.

b. Distortion cannot be completely eliminated in a loaded circuit. The effective resistance of a loading coil varies with frequency, as a result of hysteresis and eddy current losses in the core of the loading coil.

c. The combination of distributed capacitance and lumped inductance acts as a filter. The curve for the attenuation of a loaded line (see figure 13-96) shows that this equivalent filter causes the attenuation to increase sharply when the frequency exceeds a certain value, referred to as the cutoff frequency. Therefore, loading coils are usually designed to cover the band of frequencies to be transmitted, with the cut-off frequency occurring beyond the upper end of this band.

13-435. INTERFERENCE IN LINES.

13-436. The theoretical discussion of transmission line characteristics up to this point has assumed that the lines under consideration are isolated, that is, that they are far removed from ground, other transmission lines, and any type of current-carrying circuit. Actually, isolated transmission lines are rare.

13-437. Two types of interference may occur between adjacent line: noise and crosstalk. Both affect practical transmission line operation to an important extent.

13-438. Most lines are near other transmission lines or near the surface of the earth, which itself is a conductor referred to as ground. Communication lines operate at low power levels; that is, the lines are called upon to transmit relatively small amounts of power. As a result, the lines are susceptible to noise interference from other electrical circuits, particularly power lines paralleling the communication lines. Most power lines carry current at a frequency of 60 cycles. The most troublesome interference arises from the harmonics of this 60-cycle frequency; these harmonics fall in the 250-cycle to 2700-cycle audio range.

13-439. Interference is also produced by coupling between adjacent communication lines. Such interference is called crosstalk. The presence of crosstalk on a transmission line must be prevented, if possible, or at least reduced to a minimum because of the effect on the intelligibility of a message. Crosstalk exists when electrical energy is transferred from one communication circuit to another, causing the conversation on one circuit to be faintly audible on the other.

13-440. SOURCES OF INTERFERENCE IN LINES.

13-441. Five common electrical conditions are responsible for interference:

a. An unbalanced condition between the transmission lines and ground is the cause of noise induced by power-line equipment.

b. An unbalanced condition between pairs of transmission lines results in crosstalk caused by electrostatic coupling between the lines.

c. Electromagnetic coupling between the lines may produce crosstalk.

d. Conductive coupling (which seldom occurs in well maintained lines) may cause crosstalk.

e. A resistive unbalance may be still another source of crosstalk.

13-442. ELECTRICAL CAUSES OF INTERFERENCE IN LINES.

13-443. UNBALANCE TO GROUND. Distributed capacitance exists between any two conductors of a transmission line. Similarly, there is distributed capacitance between ground and each conductor of a transmission line. This is shown in figure 13-97, in which C_1 and C_2 represent the capacitances between the conductors and ground.

13-444. A transmission line is said to be balanced to ground if C_1 and C_2 are equal, that is, if the distributed capacitances to ground of the two conductors of the line are equal. This is the case when the two conductors are at equal distances from ground. Voltages are induced across the two conductors, through the distributed capacitances, by current flow up from ground. If the line is balanced to ground, the voltages induced in the two conductors are equal and

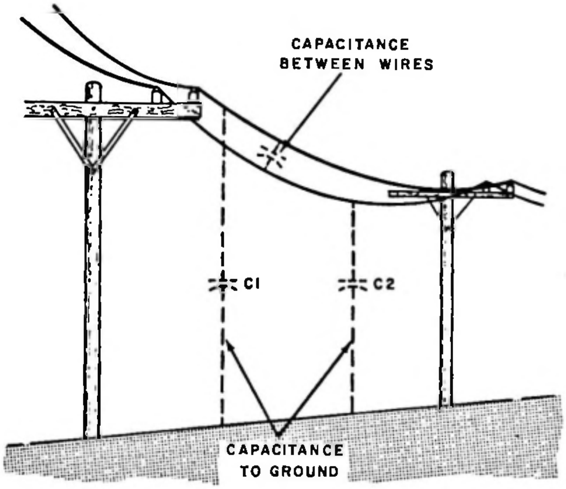


Figure 13-97. Distributed Capacitances Between Transmission Line and Ground

opposite, and the potential difference between the conductors caused by the induced voltage is thereby zero.

13-445. If the line is not balanced to ground, the induced voltages are not equal and there is a potential difference between the lines. This potential difference adds to the potential difference between the lines resulting from the transmitted signal. It constitutes a new and unwanted signal added to the first, the waveform of which is distorted. In other words, the unbalanced condition between the transmission line and ground is a source of interference. Since power-line equipments cause a 60-cycle current flow from the ground, an unbalanced condition between a communication line and ground is the source of a 60-cycle tone in the communication line.

13-446. UNBALANCE BETWEEN PAIRS. Interference can take place between adjacent transmission lines because of a capacitive unbalance between the lines themselves.

13-447. Figure 13-98, which is a diagram of two pairs of transmission lines, representing two circuits, illustrates this condition. C_{13} is the distributed capacitance between conductor 1 of circuit I and conductor 3 of circuit II. Similarly, C_{14}, C_{23}, and C_{24} represent the other distributed capacitances between the conductors of the two circuits. Energy may be transferred from one circuit to the other when the electric field of one circuit causes a voltage to be developed in the second circuit. For this reason, this type of unbalance is called electrostatic coupling, which results in crosstalk.

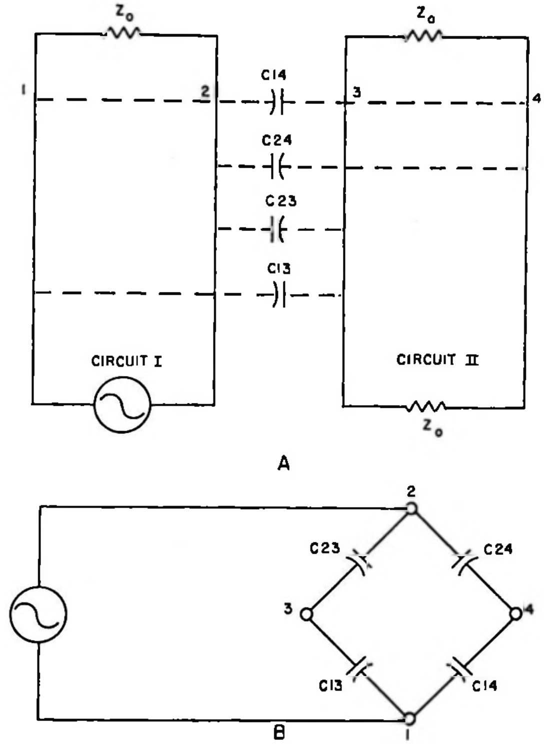


Figure 13-98. Distributed Capacitances Between Adjacent Conductors

13-448. The equivalent circuit of the two circuits, or pairs of conductors, so coupled in a bridge circuit is shown in part B of figure 13-98. If $C_{13} = C_{14}$ and $C_{23} = C_{24}$, the voltages on lines 1 and 2 divide in the same ratio across C_{13} and C_{23}, and across C_{14} and C_{24}. Thus, equal voltages appear across equal capacitances, and no potential difference exists between lines 3 and 4 because of the voltages on lines 1 and 2. The circuit is then said to be balanced. Note that for balance to exist in the opposite direction (that is, no voltage coupled from 3 and 4 into 1 and 2), it is also necessary that $C_{13} = C_{23}$ and $C_{14} = C_{24}$.

13-449. This balance is achieved by proper physical positioning of the lines, so that the capacitances have the equality desired. For example, an unbalance exists if the separation between conductors 2 and 3 is larger than the separation between conductors 1 and 4; under this condition, C_{23} is smaller than C_{14}, and a signal transmitted over circuit II will add to its part of any signal simultaneously being transmitted over circuit I as a result of the variation of the electric fields between the conductors.

13-450. ELECTROMAGNETIC COUPLING. Electromagnetic coupling exists between two pairs when the magnetic field of one pair induces a voltage in an adjacent pair. Figure 13-98 represents two pairs of transmission-line wires. The varying current in the wire pair of circuit I sets up a magnetic field about conductors 1 and 2. This magnetic field spreads out and cuts the magnetic field of conductors 3 and 4 of circuit II. If conductors 3 and 4 were at equal distances from conductor 2, they would be cut by an equal number of magnetic lines of force; the voltage induced in conductors 3 and 4 would thus be equal, and there would be no potential difference between them resulting from the current in conductor 2. However, conductor 3 is closer than conductor 4 to conductor 2. Therefore, both conductors in the pair of circuit II will be cut by different numbers of lines of force, so that different voltages, e_3 and e_4, will be induced in the two conductors. There is, then, a potential difference between conductors 3 and 4 as a result of the current flow in the pair of circuit I. This potential difference will be in addition to the potential difference between the conductors (3 and 4) caused by the transmitted signal. The combination of the two potential differences results in an altered signal and a garbled message at the receiver.

13-451. CONDUCTIVE COUPLING. Conductive coupling exists when there is leakage between two pairs of conductors as a result of faulty insulation. Such coupling usually does not occur on well kept transmission lines.

13-452. RESISTIVE UNBALANCE. A resistive unbalance exists when the resistances in the two wires of a transmission line are unequal; it is caused by the presence of different gages of wire, improper splices, or other types of poor connection. Referring to figure 13-99, it may be seen that if conductors 3 and 4 of pair II are at equal distances from conductor 2 in pair I, the induced voltages in conductors 3 and 4 are equal and no crosstalk results. If the resistances in conductors 3 and 4 are unequal, however, the induced currents in the conductors will be unequal. This difference in current will be added to the current of the transmitted signal, resulting in crosstalk. The obvious remedy for a resistive unbalance is to balance the circuit that has a wire containing additional moisture or to remove part of the excess resistance from the other line.

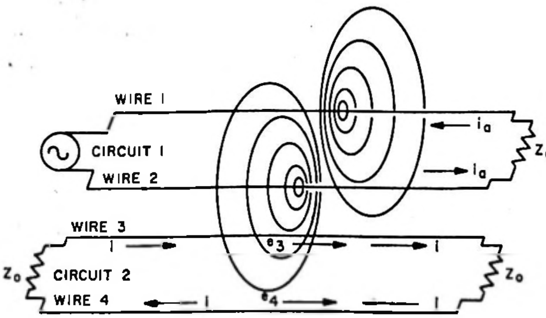


Figure 13-99. Effect of Magnetic Induction on Crosstalk

13-453. REDUCTION OF INTERFERENCE IN LINES.

13-454. CROSSTALK REDUCTION PRACTICES.

13-455. In considering practical methods for keeping the crosstalk in long lines at a reasonable minimum, it is desirable first to consider the effects of certain basic design features of long circuits with respect to crosstalk. In general, these will apply equally to both open wire lines and cable facilities, and at either voice or carrier frequencies. One such important feature is the effect of the location of telephone repeaters with respect to crosstalk. Thus, it is obvious that if two circuits are in close proximity at a point near a repeater station, and one circuit is carrying the high current levels approaching the input of the repeater, the tendency of the first circuit to interfere with the second circuit is very great. The very small percentage of the current in the first circuit which may be induced into the second circuit will be amplified by the repeater on that circuit along with, and to the same degree as, the normal transmission. The best practical remedy for this condition, of course, is to avoid such situations by keeping circuits carrying high-level energy away from low-level circuits as much as possible. Where such physical separation between the circuits is not feasible, differences in energy level between adjacent circuits can frequently be minimized by proper adjustment of repeater gains when the circuit is designed.

13-456. Another basic element of circuit design is that in most of the longer voice-frequency cable circuits, and in all carrier circuits, the effect of near-end crosstalk is minimized by the use of separate paths for transmission in the two directions. In cable circuits, the wires carrying the transmission in the two directions are physically separated as much as possible by placing them in different layers or segments of the cable, or, in the case of high-frequency carrier circuits, in different cables. A comparable separation is obtained in open-wire carrier circuits by using entirely different bands of frequencies for transmission in the two directions.

13-457. Furthermore, any near-end crosstalk occurring in spite of these physical separations is returned on the disturbed circuit to the output of an amplifier. Since the amplifier is a one-way device, the crosstalk can proceed no farther and does not reach the terminal of the circuit. Near-end crosstalk in such circuits is therefore of little importance, except insofar as it may be converted into far-end crosstalk by reflection from an impedance irregularity. To avoid this latter effect, it is essential that all circuit impedances be so matched as to eliminate important reflection possibilities.

13-458. Aside from the above-mentioned techniques for avoiding crosstalk through circuit design methods, practical procedures differ considerably, depending upon the type of facility. It is desirable, accordingly, to analyze the problem separately for both open-wire and cable facilities.

13-459. In the case of open-wire lines, crosstalk reduction depends upon three principal factors: wire configuration on the poles, transpositions, and resistance balance. Resistance balance is primarily a question of maintenance, and ordinarily presents no great difficulty. The use of high-frequency carrier facilities, with their much greater crosstalk possibilities, has led to the development of new configurations of open-wire lines in which the wires of individual pairs are spaced closer together and the pairs are spaced farther apart. One standard pole-head configuration of this kind is illustrated in figure 13-100, where it may be noted that the separation between the wires

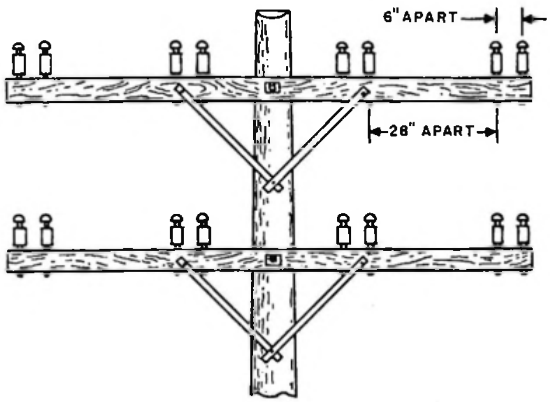


Figure 13-100. Spacing of Wires To Minimize Crosstalk

of each pair is 6 inches and the horizontal separation on the crossarm between any two pairs is at least 28 inches.

13-460. The basic principle of transpositions was outlined in the preceding paragraph. It was noted there that a large number of transpositions was needed in any long section of line to reduce crosstalk to the required minimum. In the entire discussion, moreover, only two pairs were considered. In actual practice, an open-wire line usually carries many more than two pairs of wires, and obviously there are crosstalk possibilities between any two pairs on such a line. These possibilities are greater between the pairs that are adjacent to each other, but all of the other possibilities are sufficiently large that they must be taken into consideration in designing a transposition arrangement for the line. A practical arrangement must also guard against crosstalk between side and phantom circuits and between the phantoms themselves, when such circuits are used.

13-461. There is still another extremely important factor which has not been considered up to this time. This is the possibility of crosstalk from one circuit to another by way of a third circuit. In a line carrying many circuits, there are a large number of these tertiary circuits by means of which crosstalk might be carried from any one pair to any other pair. Even the hypothetical line that we considered in the first place, carrying only four wires, has two such tertiary circuits. These are the phantom circuit, made up of the two wires of one pair transmitting in one direction and the two wires of the other pair transmitting in the opposite direction; and the "ghost" circuit, made up of the four wires acting as one side of a circuit, with a ground return. (Note that these circuits exist as tertiary crosstalk paths, regardless of whether a working phantom circuit is actually applied to the four wires.) Needless to say, the presence of these tertiary circuits in a line complicates the problem of designing an effective transposition arrangement. In fact, the problem can become so complicated that no attempt will be made here to analyze the problem in detail.

13-462. Transposition methods for open-wire lines are designed for unit lengths ranging from a few hundred feet to about 8 miles. The purpose of the design is to approach as closely as possible to a complete crosstalk balance in each such unit section. Any number of sections can then be connected in tandem. The nonuniformity in the length of sections is the result of discontinuities in the line, such as junctions with other lines, wires dropped off or added, etc. It is naturally desirable that such points of discontinuity coincide with junctions between transposition sections, where the crosstalk is balanced out.

13-463. Physically, there are two standard methods for effecting transpositions between wires on pole lines. These, which are known as point-type transposition and drop-bracket transposition, are shown respectively in figures 13-101 and 13-102. The former (point-type) is widely used in lines used for

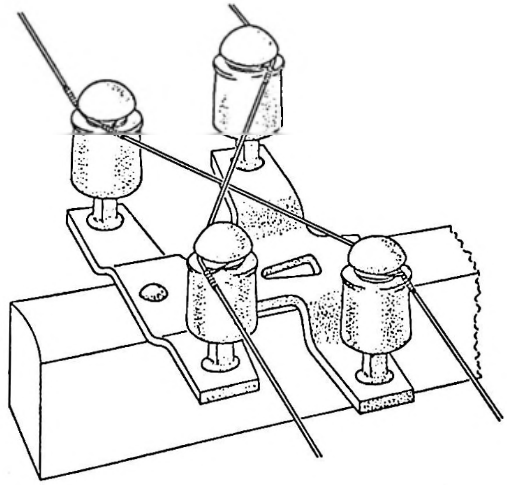


Figure 13-101. Point-Type Transposition

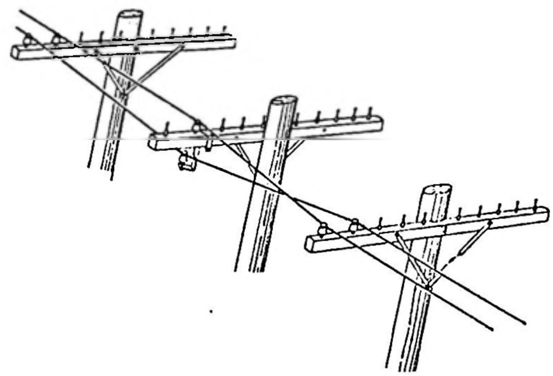


Figure 13-102. Drop-Bracket Transposition

carrier facilities because it does not change the configuration of the wires in the adjacent spans, as does the drop-bracket type. Where very high frequencies are used, this becomes extremely important. In fact, the sensitivity of these carrier equipments to crosstalk is so great that every possible effort must be made to avoid even slight deviations in the amount of sag of the wires in the spans between poles.

13-464. MINIMIZING CROSSTALK IN CABLE CIRCUITS.

13-465. The most striking feature of cable circuits, with respect to crosstalk, is that the conductors are crowded together. This is particularly true of the two wires of each circuit pair, which are separated by only thin coatings of paper insulation. As we have already seen, the close spacing of the two wires of a pair in which equal and opposite circuits are flowing tends to minimize the external effect of the electromagnetic field of the pair. Moreover, in the process of manufacture, the cable conductors are very thoroughly transposed by twisting the two wires of each pair together, by twisting the two pairs of each group of four wires together to form quads, and by spiraling the quads in opposite directions about the cable core. Also, cables are so manufactured and installed that their conductors are practically free from series resistance unbalances or insulation leakages. On the other hand, the close spacing of many circuits within the cable sheath, as well as their proximity to the sheath itself, offsets the above advantages to a considerable extent.

13-466. At voice frequencies, magnetic induction (inductive coupling) between circuits in a cable is normally so small that it is of relatively little importance in causing crosstalk. The same cannot be said of electric induction (capacitive coupling). Despite the most careful manufacturing methods, the capacitance unbalances between cable conductors usually remain large enough to cause objectionable crosstalk in long circuits. You can guard against this crosstalk by the use of additional balancing techniques when you install a long toll cable.

13-467. Voice-frequency crosstalk between circuits in different quads of a cable can be reduced to a satisfactory minimum by splicing the successive lengths of cable in a more or less random manner so that no two quads are adjacent to each other for more than a small part of their total length. This technique, of course, has no effect upon the crosstalk between circuits in the same quad. To reduce this circuit crosstalk, it is necessary to measure the capacitive unbalance of each quad at the time of installation and then to correct such unbalance if it is found to be large enough to cause serious crosstalk.

13-468. There are two principal methods of effecting this latter correction. One depends upon measuring the unbalances at several equally spaced splicing points within each loading section, and then splicing the quads together in such a way that a given unbalance in one section is counteracted by an equal and opposite unbalance in the adjacent section. This will perhaps be made clearer by referring to figure 13-103, where a cross section of the four wires of a quad is shown, with the capacitive values between the wires indicated by small capacitors. The wires marked 1 and 2 form one pair of the quad, and the wires marked 3 and 4 form the other pair. (The capacitances between the pairs themselves are not shown because they have no effect on crosstalk.) The ideal condition in such a quad is that the values of all four capacitances (A, B, C, and D) shall be equal, and that capacitance E shall equal capacitance F, and capacitance G shall equal capacitance H. In this case there is no unbalance within the quad and no crosstalk. However, if it is found, for example, that capacitance A in a certain quad of one section of the cable is too low, this quad can be spliced to a quad in an adjacent section of equal length in which capacitance A is too high by an approximately equal amount. The net unbalance of the connected quad over the two sections will thus be made to approach zero.

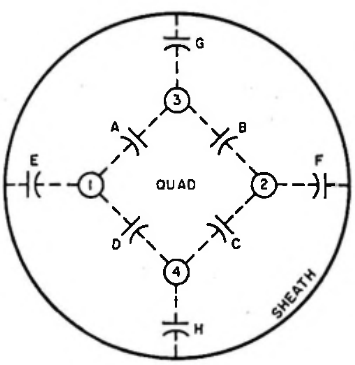


Figure 13-103. Balancing Adjacent Sections

13-469. In the second method, the unbalances are counteracted in part by connecting small balancing capacitors into the circuits at one or two points in each loading section. This, combined with a limited number of "test splices", accomplishes the net result desired with great accuracy, and reduces the number of capacitance unbalance tests that have to be made when a cable is installed. These balancing capacitors consist of short lengths of two parallel, insulated, fine-gage wire wound helically around a nonconducting core. Two terminals of this tiny capacitor are connected across the two line conductors whose capacitance is to be increased, and the other ends of the wires can be cut off at whatever point is necessary to give the capacitor the precise value required. A large number of these capacitors can be included within the sleeve at a splicing point. In certain cases where the cable conductors are to be used for 4-wire circuits, it is practicable to balance the capacitances for a whole repeater section by adding capacitors of this type at one end. The capacitance balancing methods outlined above have been found adequate in practice for keeping crosstalk to a tolerable minimum in voice-frequency cable circuits. When carrier systems are applied to cable circuits, the crosstalk problem becomes more severe. In this case, while capacitive coupling is still of consequence, inductive coupling becomes much more important as a cause of crosstalk. In fact, at the higher frequencies it predominates over capacitive coupling as a cause of crosstalk by a ratio of about 3 to 1. Accordingly, additional crosstalk reduction measures must be applied to cable conductors used for such carrier equipments.

13-470. The crosstalk possibilities at these high frequencies are so great, in fact, that a number of basic changes in circuit design are required. In the first place, the carrier pairs are used for carrier transmission only. Next, the transmitting paths in the two directions are kept entirely separated by using separate cables for transmission east to west, and west to east. The circuits in the two directions are likewise kept separated within the terminal offices and repeater stations, and shielded office wiring is used in all cases. This means that the energy levels of the carrier currents are approximately the same in all physically adjacent conductors, and that near-end crosstalk possibilities are completely eliminated (assuming that you have properly guarded against reflection effects).

13-471. Far-end crosstalk between carrier pairs is minimized by balancing out the capacitive and inductive couplings. In addition, special precautions are taken to prevent interaction crosstalk between carrier pairs by way of the voice-frequency pairs in the cable. The most effective means of accomplishing this is the complete transposition of the entire group of carrier pairs between the two cables at each repeater station. As may be seen from figure 13-104, such transposition automatically eliminates crosstalk by way of the voice-frequency pairs from the outputs of the amplifiers in the carrier pairs to the inputs of amplifiers in other carrier pairs. Carrier filters or noise suppression coils are also inserted in

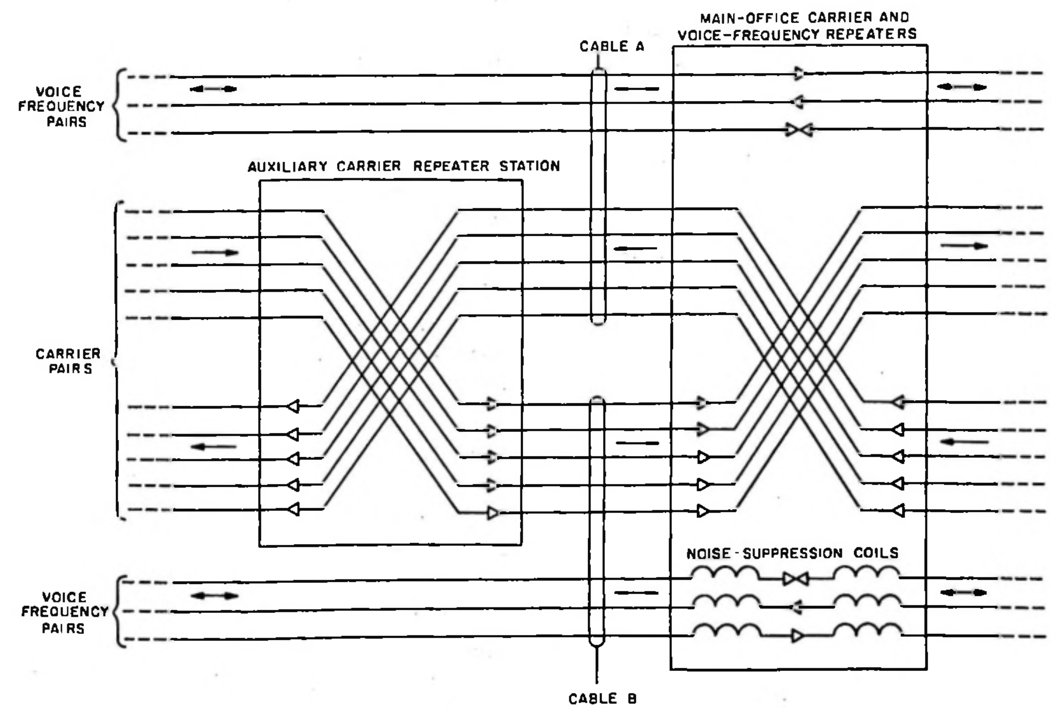


Figure 13-104. Arrangements for Reducing Noise and Crosstalk in High-Frequency Carrier Equipments

the voice-frequency pairs at voice-frequency repeater stations, and certain other points, to discourage the transmission of induced carrier-frequency currents over the voice-frequency conductors.

13-472. The methods of balancing out capacitive coupling between the carrier pairs themselves are essentially the same as were discussed above in connection with voice-frequency transmission. In balancing out crosstalk due to inductive coupling, different methods must be used. The fundamental problem involves the balancing of every carrier pair against every other carrier pair in the same cable, in each repeater section. The method used depends, in effect, upon counteracting the crosstalk currents with equal currents flowing in the opposite direction. Thus, if in a given disturbed circuit a crosstalk current is flowing in a clockwise direction, you will wish to set up an equal current in the circuit flowing in a counterclockwise direction.

13-473. An equal current flowing in a counterclockwise direction can be effected by means of tiny transformers connected between each carrier pair and every other carrier pair. However, since it is necessary to control the magnitude of the artificially induced currents and also cause them to flow in either direction, depending upon the direction of the crosstalk current, the transformers must be designed so that the coupling between circuits can be adjusted and so that they can be poled in either direction. The method used to obtain this result is indicated schematically in figure 13-105. Here it may be noted that there are really two separate transformers, one having a reversed winding in the disturbing circuit so that a current, I, flowing in the disturbing circuit will induce oppositely poled voltages in the disturbed circuit. If the cores of the two transformers are centered as shown in the drawing, the induced voltages will be exactly equal and the net effect on the disturbed circuit will be nil. By moving the two cores as a unit in either direction, however, one or the other of the induced voltages can be made to predominate. Thus, if the cores are moved to the left, voltage e_1 will be increased while voltage e_2 will be decreased by a like amount. The result will be a current flowing in a counterclockwise direction in the disturbed circuit. On the other hand, moving the cores to the right will cause a clockwise current in the disturbed circuit, the value of which will depend upon the extent of the movement of the cores.

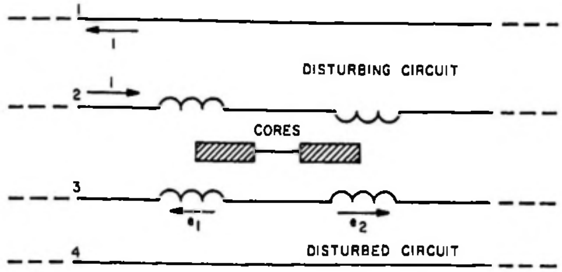


Figure 13-105. Principle of Crosstalk Balancing Coil

13-474. In practice, the balancing coils are designed to have a mutual inductance ranging from approximately +1.6 to —1.6 microhenrys for the two limiting positions of the cores. The coils are mounted in cylindrical containers arranged for rack mounting. The position of the coil cores is screw-controlled; the core can be moved through its maximum travel of 1/2 inch in about 16 complete screw turns.

13-475. In using these coils to balance out crosstalk, measurements of the inductive coupling between each pair of conductors must be made and each coil adjusted to counteract this coupling. In a cable containing a large number of carrier pairs, the number of coils required at each repeater station becomes rather large, since one coil is required for every possible combination of pairs. In practice, also, an additional coil is used for each quad to provide sufficient margin for balancing out side-to-side crosstalk. Thus, 20 pairs require a total of 200 coils, 40 pairs require 800 coils, and the maximum of 100 pairs requires 5000 coils. The coils are installed in unit panels arranged for balancing 20 pairs, and additional inter-group panels are added as successive 20-pair carrier groups are put into service. A special crisscross wiring arrangement, such as that indicated in figure 13-106, is employed. This is necessary in order that the currents in any two pairs shall flow through the same number of coils before reaching the coil that balances these two pairs, thus insuring that the phase shift up to the balancing coil will be approximately the same on both pairs.

13-476. APPLICATION OF RESONANT LINES.

13-477. GENERAL.

13-478. Rf lines are used for many purposes other than the transmission of power from one point to another. For example, rf lines are used as:

a. Line sections in parallel-resonant circuits.

b. Line sections in series-resonant circuits.

c. Reactances.

d. Impedance-matching devices.

e. Phase shifters and inverters.

f. Time delay devices.

g. Energy storage devices.

h. Switching devices.

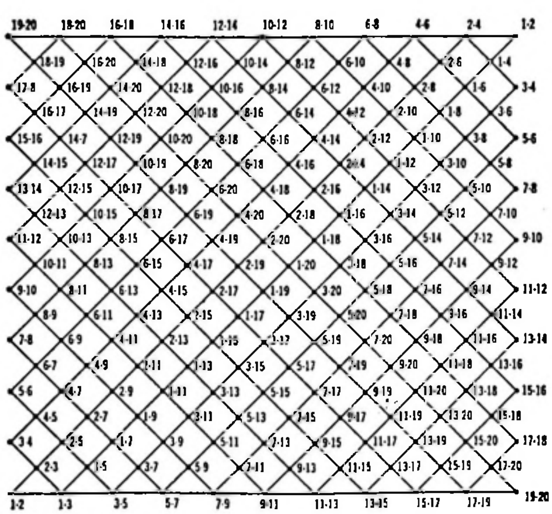


Figure 13-106. Method of Connecting Crosstalk Balancing Coils

13-479. LINE SECTIONS AS PARALLEL-RESONANT CIRCUITS.

13-480. Above 50 mc, the difficulty of constructing efficient tuned circuits with common coils and capacitors makes some other type of circuit desirable. Sections of transmission line can fill this need, and are used widely where space and weight considerations permit. Quarter-wave closed end and half-wave open-end resonant line sections both offer the characteristics of parallel resonance. They also have a high Q at frequencies where tank circuits with lumped property elements become inefficient or useless. Close to resonance, the impedance curve for a quarter-wave closed-end or half-wave open-end resonant line resembles the impedance curve for a conventional parallel-resonant circuit using lumped-property elements. The curves differ at points farther from resonance, because the reactance of the line section depends on the reflection effect which produces standing waves, and not on any lumped capacitance or inductance. Resonant-line sections may be used as tank circuits in either single-ended or push-pull arrangements, depending on the requirements of a particular application. A line section, any circuit, or any part of a circuit is said to be balanced when it is composed of two or more potential paths which operate similarly with respect to ground, as in a push-pull circuit. An unbalanced line section or circuit is one in which a single potential path operates at some value above or below ground. An example of this is the plate circuit of a single amplifier. A horizontal parallel-wire section is considered to be balanced. However, the concentric line and the two-wire line operated with one wire much closer to the ground than the other is considered to be unbalanced.

13-481. RESONANT LINE IN TANK CIRCUITS. Parallel-resonant line sections used as tank circuits in single-ended and push-pull arrangements are shown in figure 13-107. In a properly operated coaxial line, the outer skin of the outer conductor carries no rf current, and may be grounded. The outer skin of the inside conductor is above ground potential, and is at a relatively high impedance above ground. For this reason, the unbalanced coaxial line is usually found in single-ended circuits where it is desired to use line sections in place of lumped-property tank circuit elements. A common unbalanced arrangement is shown in part A of figure 13-107, together with a simplified schematic of a conventional tank circuit. In

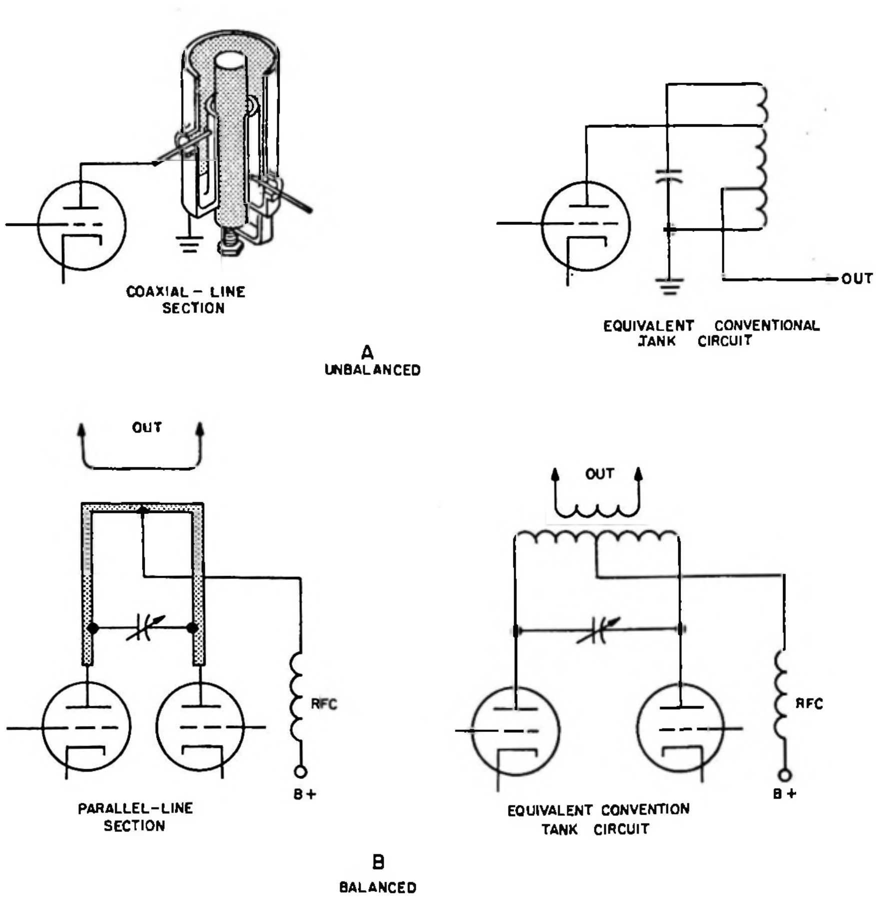


Figure 13-107. Resonant-Line Sections as Parallel-Resonant (Tank) Circuits

the parallel-wire line section, shown in part B of the figure, rf current flows in both conductors. Both are above ground potential, and they present equal impedances to ground; therefore, the line is considered balanced, and is particularly well adapted to the requirements of push-pull tank circuits. In both balanced and unbalanced circuits at the resonant frequency, the line section appears to the source as a high, pure resistance; energy is stored in the line, and very little power is required from the source to maintain this condition. The two-wire resonant-line section is generally used in the frequency range of 50 to 300 mc, because the tank circuit is somewhat more easily tuned. However, radiation losses can occur from two-wire sections, particularly at the higher frequencies. This reduces the effective Q and decreases the impedance at reasonance, resulting in a decrease in efficiency as the frequency is raised. Since coaxial lines are self-shielding, radiation losses from coaxial-line sections are extremely small, and there is little loss of Q from this source. Resonant-line sections made from coaxial line are more difficult to tune, however, because of their physical arrangement.

13-482. TUNING. Resonant-line sections used as parallel-resonant circuits differ from the conventional lumped-property components in their response to the harmonics of the fundamental resonant frequency. For example, a quarter-wave section of a closed line will act as a parallel-resonant circuit to the fundamental frequency. However, at twice the fundamental frequency (second harmonic), the quarter-wave section acts as a series-resonant circuit. The impedance curves of figures 13-79 and 13-80 show that a line section theoretically offers maximum or minimum impedance at every harmonic of the fundamental frequency. Actually, these curves are not a true representation because the end effect causes the open end of a resonant line operating at a harmonic to behave as if the section had been lengthened physically by a fraction of a wavelength. The section does not become resonant exactly at the second harmonic, but somewhat above it. When a lumped reactance is used across the line for tuning purposes, the end effect is increased. Where it is necessary to tune a line section, a variable capacitance (see part A of figure 13-108) is often used to avoid changing the actual physical length of the line. This capacitor shortens the effective wavelength of the resonant section, thus reducing the space taken up by the tank circuit. However, the presence of such a

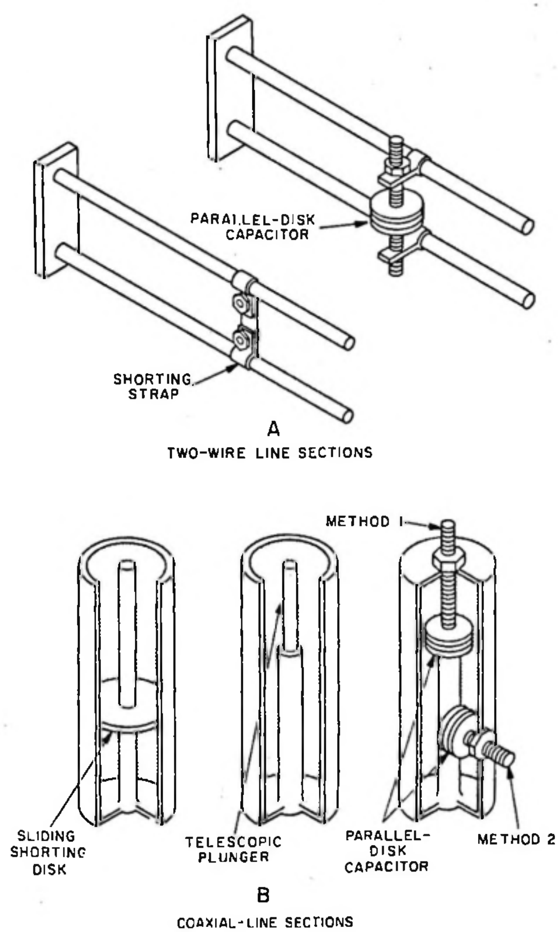


Figure 13-108. Common Tuning Methods for Two-Wire and Coaxial Resonant-Line Sections

lumped capacitance upsets the response of the section at the harmonic frequencies. Other methods of tuning resonant sections may be used; the most common forms are shown in part B of figure 13-108.

13-483. Two-wire sections should be spaced no farther apart than 1/10 wavelength at the resonant frequency, or radiation losses may become excessive. Spacing large-diameter conductors too closely also introduces losses, resulting from eddy currents, and adds to the danger of voltage breakdown and arcing. For this reason, parallel wires should not be closer than about twice the diameter of one conductor. When tuning two-wire line sections by means of a short-circuiting strap, as shown in part A of figure 13-108, the strap must make a very low-resistance contact with the conductors because any appreciable resistance will seriously reduce the Q of the tank circuit. When a shunt capacitance is used, the capacitor must have minimum distributed inductance and the lowest possible losses. Supporting the capacitor entirely on the line conductors, so that no solid dielectric is in the electric field, is the most practicable means of maintaining a high Q. A telescoping tube may be moved inside the line to change the effective length of the inner conductor. Coaxial-line sections may also be tuned by means of a shorting disk or a lumped capacitor. The lumped capacitor must have low loss and high Q. It may be connected at the open end of the line, which gives the greatest tuning effect per unit of capacitance, or by tapping down on the line, which has less effect on the circuit Q. If the shorting-disk method of tuning is used, the disk must make perfect electrical contact to avoid the introduction of additional contact resistance. Different methods of tuning coaxial lines are illustrated in part B of figure 13-108.

13-484. COUPLING TO RESONANT-LINE SECTIONS. As long as the total length of a line section is resonant, both the generator and the load may be tapped on at any desired impedance point. This is frequently done to make the impedances match, just as connections are tapped on a conventional tank circuit coil. No matter where the generator or the load is tapped, the tank circuit will appear resistive. When coupling inductively to the two-wire line section, a hairpin loop is used. Since the rf field about the two-wire resonant-line section is not confined, the hairpin loop may be placed at the necessary distance from the line section to give the desired degree of coupling. Inductive coupling to the coaxial-line section is mechanically more difficult because the field is confined almost entirely within the outer conductor. A small loop is inserted through an opening in the outer conductor, and provision is often made for rotating the loop to provide control of the degree of coupling. When the loop is at right angles to the field, maximum coupling is achieved; when it is parallel to the field, coupling drops to a minimum, and would go to zero except for the small capacitance coupling between the loop and the inner conductor. To reduce losses, the two leads from the coupling loop are frequently brought out in the form of a flexible coaxial cable.

13-485. IMPEDANCE AND Q. In coaxial-line sections, the inside surface (pertaining to diameter D) of the outer conductor and the outside surface (pertaining to diameter d) of the inner conductor, are called the effective surfaces. The ratio of their diameters, D/d, is a main factor in determining the unloaded Q of a line section using air as the dielectric. The highest unloaded Q is obtained when the ratio of diameters is 3.6 to 1, that is, when the inside diameter of the outer conductor is 3.6 times the outside diameter of the inner conductor. The actual unloaded Q of a line section in practical use is also influenced by the operating frequency. This is shown in part A of figure 13-109, with the Q increasing as the effective diameter of the outer conductor becomes a larger

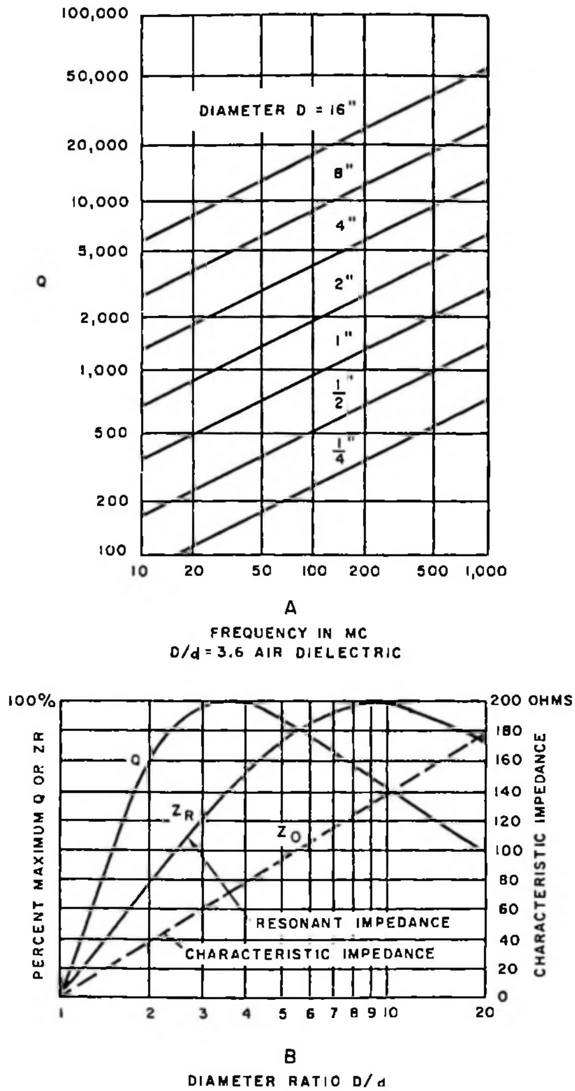


Figure 13-109. Impedance and Q of Coaxial-Line Sections

fraction of a wavelength. Obviously, this is limited by the available space in the equipment, although for laboratory purposes, where high Q is important and space is not, very large diameters may be used. The diameter of the inner conductor in each case is kept at the D/d ratio, 3.6 to 1. As shown in part B of figure 13-109, the diameter of the outer conductor is maintained constant, and the graph shows the effect of making the inner conductor smaller to change the D/d ratio. The maximum unloaded impedance at resonance occurs at a much higher ratio of D/d than the ratio which gives maximum Q. However, a tank circuit must usually be loaded if it is to be of any use, and loading always causes a reduction of the Q and the effective impedance. If a heavy load is connected to a line section designed maximum unloaded impedance, a D/d ratio of 9 to 1, the result may be an effective impedance nearly as low as the unloaded impedance of a line section designed for maximum Q, that is a D/d ratio of 3.6 to 1. This drop in effective impedance is not particularly important, since the resonance curve will remain sharp. In commercially available line sections, the D/d ratio will vary from 2:1 to as high as 10:1. Line sections that regulate the frequency in measuring devices require special construction methods; for example, the surfaces of the lines may be machined to exact tolerances, or they may be built up to the correct size by electrolytic deposit, to attain the desired D/d ratio for maximum Q. For two-wire line sections, the effects of loading on both the impedance and the Q are similar. The curves in part B of the figure are approximately true for two-wire sections if the ratio of center-to-center spacing to the radius of the conductors is substituted for the ratio of the diameters, D/d. The characteristic impedance of the two-wire line varies in the same straight-line manner, but the actual values will be double those shown for the coaxial line.

13-486. ADVANTAGES OF QUARTER-WAVE OVER LONGER RESONANT-LINE SECTIONS. It might seem that a half-wave section would have a higher Q than a quarter-wave line section, since you can store more energy in the half-wave section. However, the longer section also increases line losses, so that the Q remains about the same. Although the longer-line section suffers less reduction in Q than the quarter-wave section, you can use large conductors to reduce the effective rf resistance and thus improve the Q. You can tap the load on the tank at the desired impedance point. Harmonic response is controlled easily in a quarter-wave tank circuit, since it responds to odd harmonics only; the half-wave tank circuit, however, responds to both odd and even harmonics. You can practically eliminate the third harmonic by loading the tank with a small value of capacitive reactance to increase the end effect or by tapping the tank to the third-harmonic voltage node. Although quarter-wave and half-wave line sections are most commonly used, the statements made here about the basic quarter-wave sections are also true for sections composed of any odd multiple of a quarter-wave. Statements about half-wave sections hold true for any multiple of a half-wave.

13-487. TWO-WIRE VERSUS COAXIAL-LINE TANK CIRCUITS. The two-wire line is much more easily tuned and coupling is more convenient, but radiation from the parallel conductors is likely to be high if any unbalance is present. Unbalance may be caused by an unbalanced condition at either the source or the load, or by the fact that one conductor is either closer to ground or closer to a grounded object than the other conductor is. When radiation occurs, not only is power lost, but energy may be coupled back into the grid circuit, causing either regeneration or degeneration, and upsetting proper operation of the stage. Coaxial-line sections are more difficult to tune, but the self-shielding construction encloses all of the field except where an end is open; thus, the possibility of stray coupling is reduced considerably. This means that the coaxial tank circuit may be located closer to other circuit elements, resulting in a more compact physical arrangement with fewer losses.

13-488. PRACTICAL APPLICATIONS.

13-489. VACUUM-TUBE CIRCUIT IMPEDANCE. The characteristics of vacuum tubes are such that high impedances are often required in the grid, plate, and cathode circuits of amplifiers and oscillators. Resonant-line sections are used widely for these purposes at frequencies of 30 to 1000 mc, since high Q values are more conveniently obtained with these sections than with lumped-property tank-circuit elements. Figure 13-110 shows a low-power push-pull circuit using a half-wave resonant-line section, which is a combined oscillator and amplifier designed for operation at 400 mc. The half-wave resonant-line section used as the plate-load impedance offers a high impedance at the effective midpoint. The dc plate voltage for the tube is applied at the low-impedance midpoint, through the two 100-ohm resistors. These resistors help to damp out parasitic

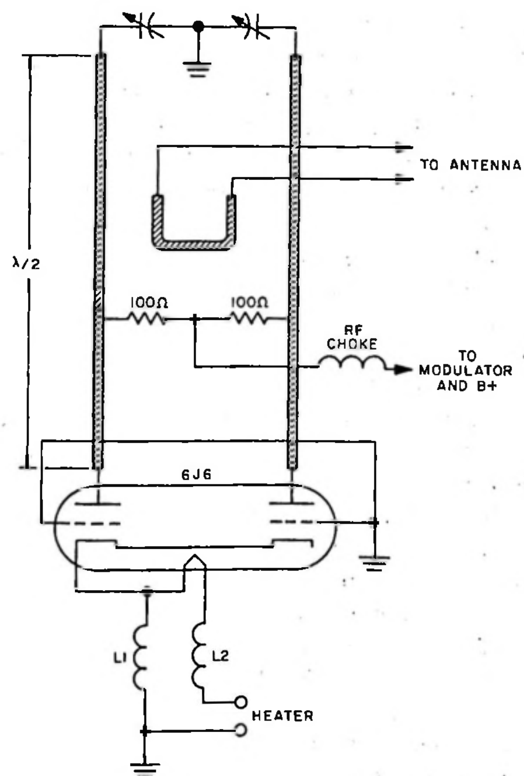


Figure 13-110. Two-Wire Line Section Used as Push-Pull Plate Load Impedance

oscillations. The plates of the tube feed into the desired high impedance at the input end of the line section. The line section is adjusted to exact resonance by the capacitor (see figure 13-110) at the output end. A hairpin loop is used to couple the antenna to the line. The stability and efficiency of the amplifier depend largely on the tank-circuit Q. By using the resonant-line section, the Q is increased approximately two to five times over that which you might expect from conventional coils and capacitors operated in the same circuit at the same frequency. Although the circuit shown here is balanced, the improvement in Q, stability, and efficiency can be obtained in single-tube amplifiers using unbalanced coaxial resonant-line sections as circuit impedances.

13-490. FREQUENCY-CONTROLLING ELEMENTS. Oscillators operating in the 30-mc to 1000-mc region require accurate frequency control. Circuits using crystal-controlled frequencies, overtone circuits, or frequency multipliers can be used; however, for many purposes, some other frequency-control element is desirable. Parallel-resonant line sections can be constructed to control frequency with an efficiency and frequency stability comparable to that of the crystal. A Hartley oscillator operating at 200 to 220 mc, with a coaxial-line section used as the tuned-grid circuit, is shown in part A of figure 13-111. The coaxial section uses a tubular inner conductor, but the outer conductor is a square shield instead of the more familiar cylindrical shield. As long as the field distribution is uniform, however, the section could offer a high Q and good frequency stability. You can tap the inner conductor at the desired impedance points and inductively couple the incoming signal to the line. You adjust the section to exact resonance by means of a capacitance which is paralleled by a temperature-compensating capacitor to avoid frequency changes during operating changes. A number of oscillator circuit arrangements

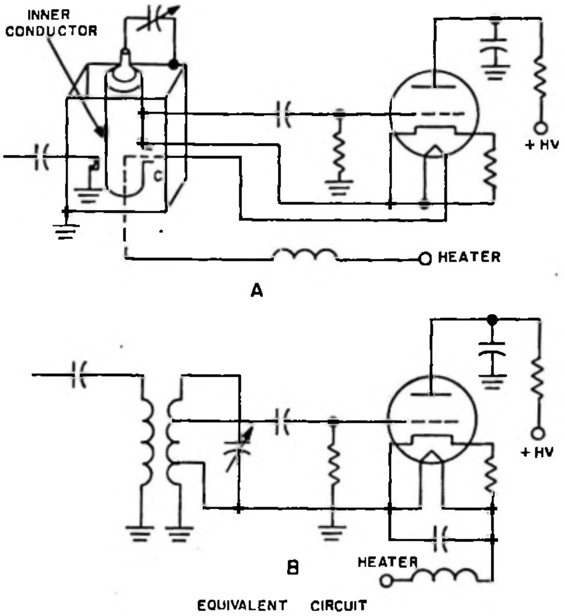


Figure 13-111. Coaxial-Line Section Used as Frequency-Controlling Element

are possible when you use the balanced line, the two-wire line, or the unbalanced line as resonant circuits. The main consideration is low loss and high-Q design to increase efficiency and minimize instability. The similarity of this circuit to one with lumped properties is shown in the equivalent circuit, part B of figure 13-111.

13-491. METALLIC INSULATORS. Since a quarter-wave closed-end resonant-line section offers a high impedance to currents at the resonant frequency, it may be used as a stand-off insulator (insulating stub) for a transmission line carrying that particular frequency. Part A of figure 13-112 shows a $\lambda/4$ stub used as an insulator in a two-wire line section, and part B of the figure shows the stub in a coaxial line. In order to offer the highest possible impedance to the transmission line, the Q must be very high. This requires that the insulating section be of low-loss construction and cut to an exact

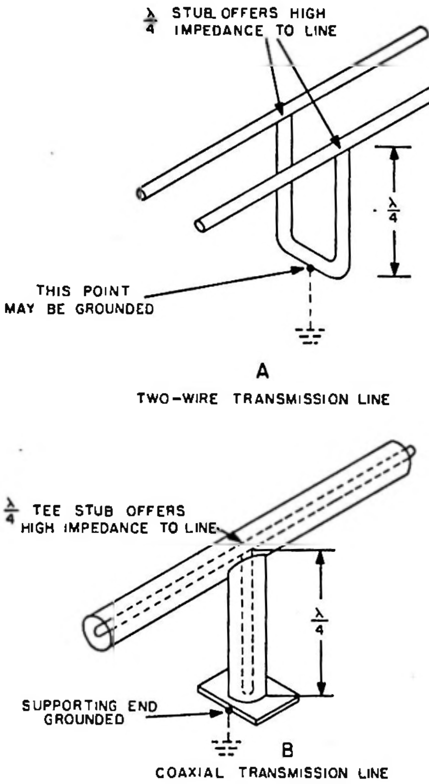


Figure 13-112. Resonant-Line Section Used as Metallic Insulator

electrical quarter-wavelength. For transmission lines operating at single frequencies, this arrangement is stable and mechanically strong and makes an efficient insulator, but it is highly sensitive to frequency. If the signal frequency is varied above or below the resonant frequency of the stub, the impedance of the stub is lowered and the stub will act as a capacitance or an inductance across the line. The sensitivity of such a line to small changes of signal frequency may be reduced somewhat by spacing stubs at odd quarter-wavelengths along the line. The additional stubs will minimize reflections and are effective on a relatively short line. A common method of broadbanding or reducing the frequency sensitivity is to incorporate, into the line, half-wave sections which have a lower characteristic impedance than the rest of the line. A stub is connected at the midpoint of each added half-wave section. This arrangement does not affect the characteristic impedance of the line, but cancels reactances and broadens the impedance curves of the stubs. As a result, the line frequency may vary up to 15 percent from the resonant frequency without serious loss of insulator efficiency.

13-492. IMPEDANCE TRANSFORMERS. The inverting action of a quarter-wave line section also makes such a section a convenient device for use as an impedance transformer. Impedances may be stepped up or stepped down, as desired, and the line section may be operated in a balanced or unbalanced condition. The balanced two-wire impedance transformer and its equivalent circuit are shown in part B of figure 13-113. The unbalanced coaxial impedance transformer and its equivalent circuit are shown in part C of the figure. The impedance curves in part A of the figure demonstrate the impedance inversion which takes place. A mathematical relationship exists which makes it easy to calculate the characteristic impedance the line section must have to provide the desired match. This relationship states that the square of the characteristic impedance equals the product of the input impedance and the load impedance, or:

$$Z_o^2 = Z_{in} \cdot Z_{load}$$

This frequently is written:

$$Z_o = \sqrt{Z_{in} \cdot Z_{load}}$$

The necessary characteristic impedance of a quarter-wave section required to match a 600-ohm transmission line to a 73-ohm antenna is calculated as follows:

$$Z_o = \sqrt{Z_{in} \cdot Z_{load}}$$

$$Z_O = \sqrt{600 \times 73}$$

$$= \sqrt{43,800}$$

$$= 209 \text{ ohms}$$

The quarter-wave line section must have an impedance of 209 ohms to provide the necessary match. Any two-line sections may be matched in this manner if the impedances are resistive. A line section used as a transformer is often called a Q section, and may be any odd number of quarter-wavelengths. To save space and make it less frequency-sensitive, a single quarter-wave section is normally used. The sensitivity to changes in frequency becomes greater as the ratio of input to load impedance increases. In the example mentioned above, the impedance transformation ratio is 600 divided by 73, or approximately 8 to 1. Under these conditions, the operating frequency might vary by several percent without seriously affecting the impedance match. At high ratios, however, a frequency deviation of as little as 1 percent will cause a mismatch and result in loss. Ordinarily, the Q section is made with a fixed characteristic impedance. For some applications, it may be necessary to adjust the output impedance because the load or source impedance is not known accurately. You can vary the spacing of two-wire line sections to achieve this, but in the coaxial line you must use a special type. This type has an elliptical outer shield and an inner conductor of the same shape, which may be rotated independently. Changing the position of the inner conductor with respect to the outer conductor varies the characteristic impedance.

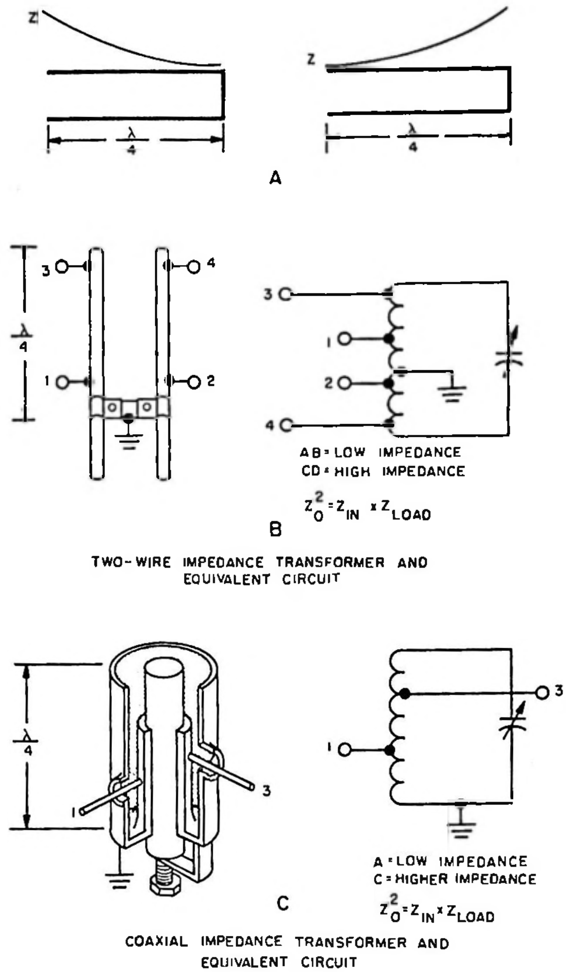


Figure 13-113. Resonant-Line Sections Used as Impedance Transformers

13-493. LINE SECTIONS AS SERIES-RESONANT CIRCUITS.

13-494. GENERAL PROPERTIES. The quarter-wave open-ended section and the half-wave closed-end section behave like series-resonant circuits at the resonant operating frequency. Part A of figure 13-114 illustrates the λ/4 section, part C the λ/2 section, and part B the equivalent conventional circuit using lumped property elements. The input impedance seen by the energy source at points 1 and 2 is low at resonance and is always a pure resistance. The Q would be infinite and the input impedance value would be zero, except for the losses in the line section. With low-loss lines, a high Q is obtained; therefore, the actual input impedance approaches zero. To the source, this looks almost like a short

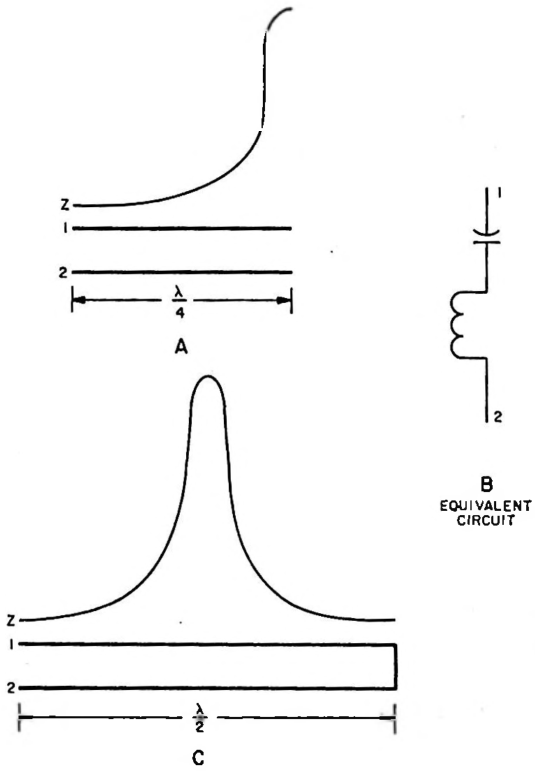


Figure 13-114. Resonant-Line Sections Used as Series-Resonant Circuits

circuit. The series-resonant effect is the same in either two-wire or coaxial-line sections. The series-resonant line differs from the parallel-resonant line only in that the quarter-wave section is open-ended and the half-wave section is closed-ended. The impedance, Q, the tuning, and other practical considerations apply equally to the use of the line sections as series or parallel-resonant circuits.

13-495. PRACTICAL USES. Resonant-line sections functioning as series-resonant circuits are used where it is desired to present a low, purely resistive impedance to a narrow band of frequencies. One of the most common applications is as a bandpass filter, used either alone in a transmission line to suppress harmonics, or in conjunction with other filter types for suppression of all the harmonic frequencies. Part A of figure 13-115 illustrates the manner in which even harmonics are practically eliminated from an antenna transmission line by inserting an open-end quarter-wave section in one side of the main line. This $\lambda/4$ section offers a low impedance, as shown in part B of figure 13-115, and does not prevent current flow at the fundamental frequency. At the second harmonic, however, the wavelength is halved, and the same section becomes a half-wave open-end section which acts as a parallel-resonant circuit. An extremely high series impedance, therefore, blocks the second harmonic. At the fourth harmonic, the filter becomes a full-wave section, and at every even harmonic, the sec-

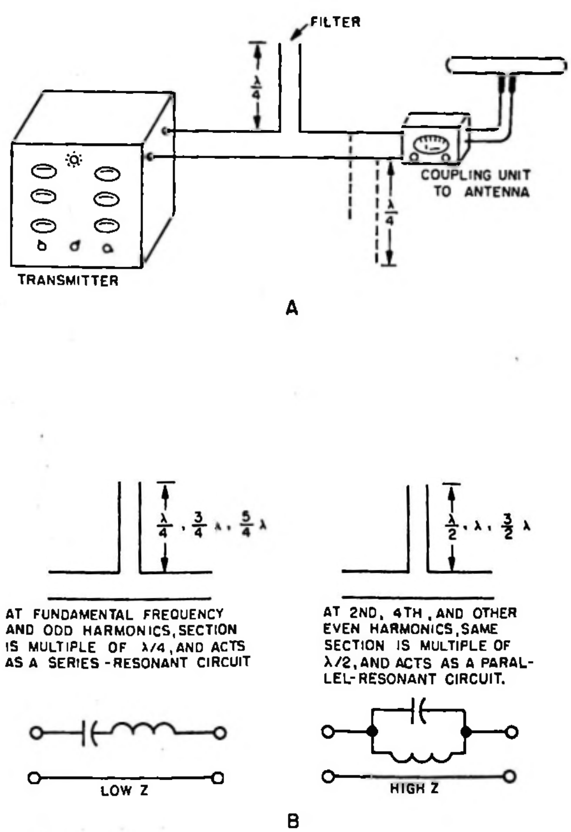


Figure 13-115. Resonant-Line Sections Used as Bandpass Filters

tion is a multiple of a half-wavelength and behaves as a half-wave section to block these frequencies. At odd harmonics, the same section becomes a multiple of the basic quarter-wave section and offers a low impedance that permits the current of the odd-harmonic frequencies to pass. If it is desired to eliminate these odd harmonics, another means must be used, since any attempt to use a resonant-line section for this purpose results in excessive loss at the fundamental frequency. Fortunately, you can eliminate the third harmonic, which is the most troublesome, in the resonant-line tank circuit of the final amplifier by captive-loading or tapping the tank.

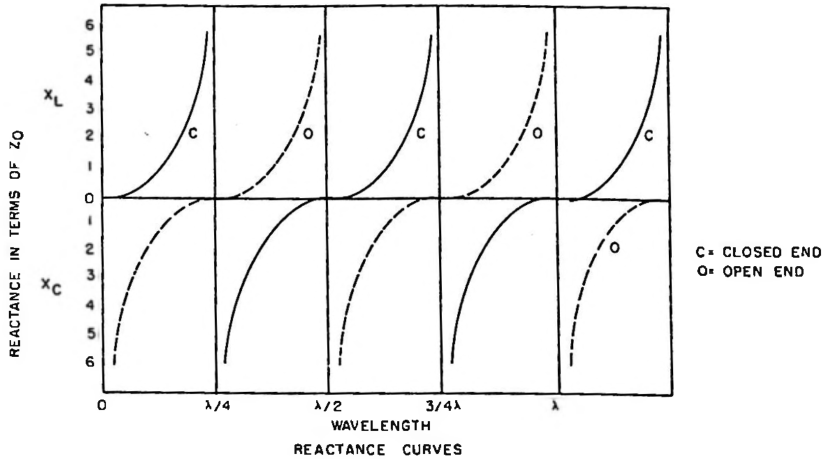
13-496. LINE SECTIONS AS REACTANCES.

13-497. GENERAL PROPERTIES. A line section other than a $\lambda/4$, or a multiple of a $\lambda/4$, functions as a reactance. The value of reactance may be high or low and it may be inductive or capacitive, depending on the electrical length and the line termination. An open-end line section less than $\lambda/4$ behaves as a capacitive reactance, and a closed-end section of the same length offers inductive reactance. For sections exactly $\lambda/8$ or odd multiples thereof, the theoretical reactance is numerically equal to the characteristic impedance of the line. Part A of figure 13-116 gives the reactance curves for line sections up to 1 wavelength. In a theoretical line with no losses, pure reactances with zero power factor would exist, but in any practical line a small resistance is present. This results in a resistive value in series with the reactance of the line section, as shown in part B of figure 13-116. At approximately a $\lambda/8$, or multiples thereof, the reactance offered by the line section is equal to the characteristic impedance. The reactance results in a definite power factor, but in a section of well designed line it will be very small and the section can function as an extremely low-loss reactance with much greater efficiency at higher frequencies than can be obtained with lumped-property reactances. The reactance of line sections, however, changes much more rapidly with frequency than that of lumped-property components.

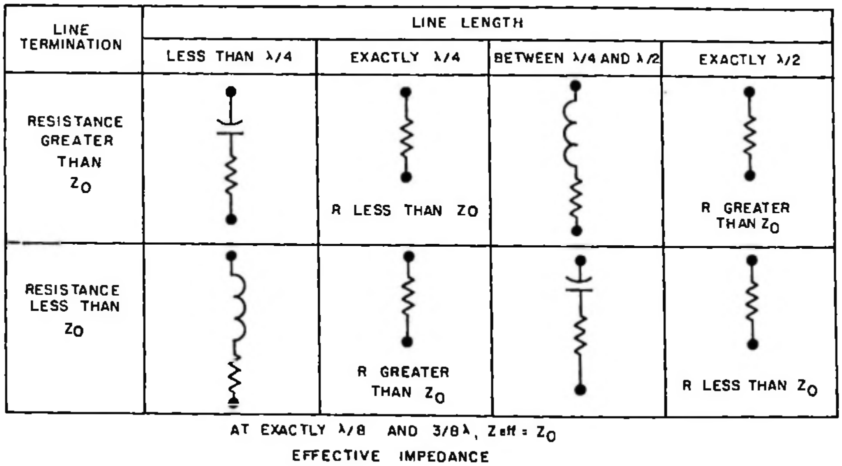
13-498. PRACTICAL USES. Line sections are often used as reactances to remove standing waves from transmission lines by tuning out or canceling reactive components in mismatched loads. Line sections also find applications as low-loss substitutes for lumped reactances in filters.

13-499. LINE-MATCHING STUBS. When a transmission line is used to feed a reactive load, such as some antenna arrays, standing waves are set up because of the load mismatch. To provide a proper impedance match and eliminate the standing waves, a stub is used. The stub is placed at a point on the line where the resistive component is equal to the characteristic impedance of the line. At this point there is also a reactive component (see part B of figure 13-116) and the length of the stub is adjusted until its reactance is equal and opposite that of the line. When this is done, the line is matched at that point and standing waves are eliminated from this point back to the input end, or source. For this reason, the stub is placed as near to the load as is practicable, even though there are four points on every electrical wavelength of line where the resistive component equals the characteristic impedance. The approximate location for the stub, as well as the stub length, can be found by using the chart of figure 13-117; however, because of variations in practical lines, some final adjustment is necessary.

13-500. A closed-end stub is generally used for convenience of adjustment. In practice, the standing-wave ratio must be determined by means of a probe and an indicating device Maximum and minimum voltage or current points are located, and the maximum voltage is divided by the minimum to give the swr.



A



B

Figure 13-116. Reactance Curves and Impedance of Line Sections

The stub is then connected to the line at the point determined from the chart. A closed-end stub always is placed between the last voltage maximum on the line and the input end—never between E_{max} and the load end. Usually, a simple final adjustment of stub length completes the matching procedure. If the swr of the line is at least 10 or more, the adjustment of the stub is critical. Figure 13-118 illustrates the effect of stub matching on a line with standing waves.

When the line is not stubbed, the source sees an impedance that may be any value, depending on the length of the line; standing waves appear along the line, as shown in part A of the figure. With the stub attached at the proper point, as in part B of the figure, the source sees an impedance which is equal to the characteristic impedance of the line and the resonant section of the line matches the load impedance, no matter what its value may be. An alternate method of

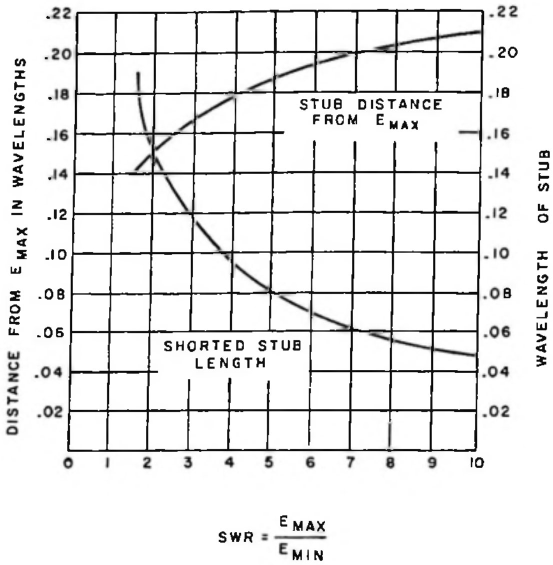


Figure 13-117. Stub Length and Position for Impedance Matching

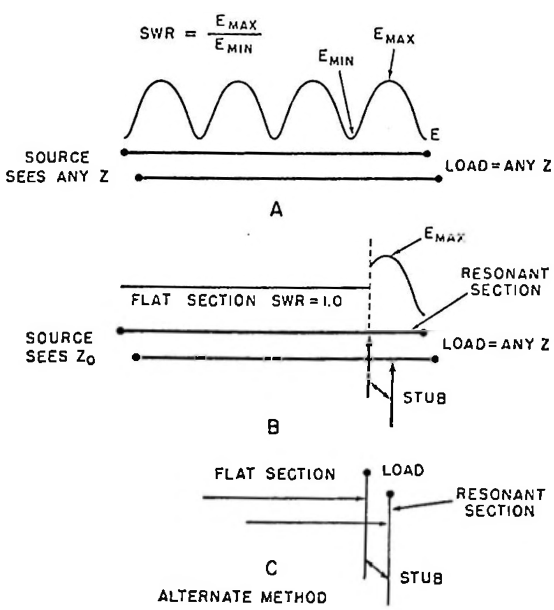


Figure 13-118. Effect of Stub Impedance Matching

connecting the stub is shown in part C of the figure. It is much easier to use single-stub matching with two-wire lines than with coaxial lines, because of the greater convenience of measuring the swr and moving the stub. The stub section should be identical, both physically and electrically, to the main line.

13-501. DOUBLE-STUB MATCHING. In coaxial lines, the difficulty of moving the stub along the line for final adjustment is eliminated by using two stubs, which are adjustable in length by means of shorting plungers. These stubs are located anywhere on the load end of the line (see figure 13-119), but the spacing between the stubs must be exactly $\lambda/8$ or an odd multiple thereof. The arrangement will not handle the variety of complex load impedances that the movable single stub will handle, but where the swr is not unusually high it is effective and relatively noncritical. The second stub functions

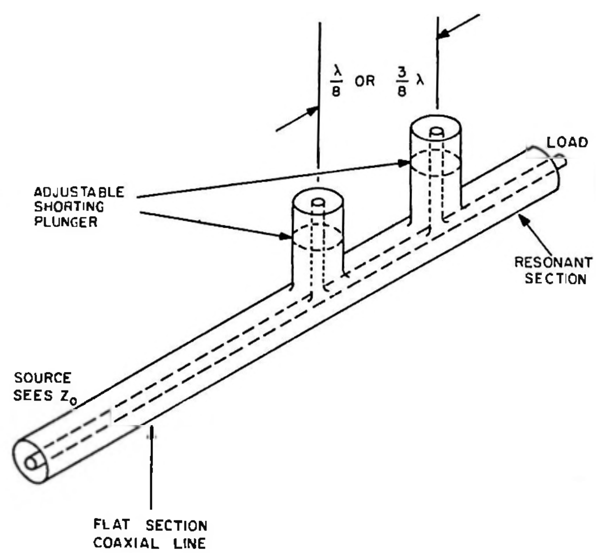


Figure 13-119. Double-Stub Impedance Matching

as a compensating adjustment, giving an electrical effect which moves the position of the first stub.

13-502. STUB AS IMPEDANCE TRANSFORMER. The matching stub, in combination with the short resonant section of line between it and the load, acts as an impedance transformer. To eliminate the standing waves, it transforms the complex load impedance into a resistive impedance equal to Z_O. Since this is true, the stub may also be used for transforming the line impedance to match the line to the source. The load-matching stub is located and adjusted as already described, and a second stub is placed at the input end to match the line to the source in exactly the same manner. Thus, the impedances shown in figure 13-120 are matched perfectly and the transmission line operates without standing waves.

13-503. MISCELLANEOUS USES OF LINE SECTIONS.

13-504. CONVERTING FROM BALANCED TO UNBALANCED IMPEDANCES. A balanced condition in a transmission line is defined as a condition in which equal or nearly equal amounts of positive and negative voltages appear above or below a reference point, which may be ground or some established voltage. In high-gain antenna systems it is usually essential to have fairly good balance to ground if the intended directive pattern is to be obtained; if such an antenna were connected directly to an unbalanced coaxial line, either the load balance or the line operation, or both, would be upset. The unbalance would shift the electrical feed point of the load away from the designated point, changing the ohmic value and introducing reactance into the effective load impedance. On the other hand, a balanced load would act as an interruption or discontinuity in the line. Any discontinuity in a transmission line can cause standing waves. For example, if you put rf current on the outside braid of the coaxial line, causing unwanted radiation, and couple the load reactance back into the source, the discontinuity will cause standing waves, if the impedance of the antenna is purely resistive and matches the line impedance. Obviously, you must use some means of converting from an unbalanced to a balanced condition.

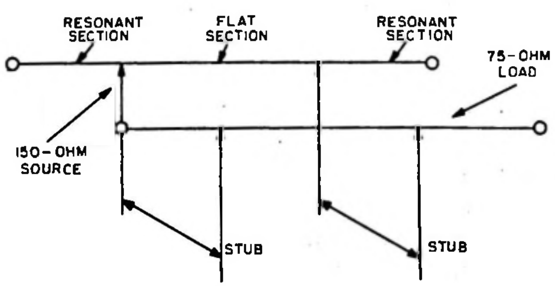


Figure 13-120. Stubs Used as Impedance Transformers

13-505. When the impedances of the devices to be connected already match and no impedance transformation is desired, a bazooka type of line-balance converter, called a balun, is widely used. This is a quarter-wave shield which is placed around the end of the coaxial line. Part A of figure 13-121 shows a closed-end quarter-wave coaxial section between the detuning sleeve and the outer braid of the unbalanced line. This causes a high impedance to exist between points 1 and 2. The inner conductor, labeled 3, is already a high impedance with respect to ground. Part B of the figure shows that point 2, which was formerly at ground potential, is now free, and its impedance to ground will depend on the load to which it is connected. If points 2 and 3 are connected to a balanced line, they will assume equal impedances, and point 1 will be at ground potential. The equivalent circuit, shown in part C of the figure, demonstrates the 1:1 transformation in terms of lumped-property components. The bazooka converter gives excellent performance as long

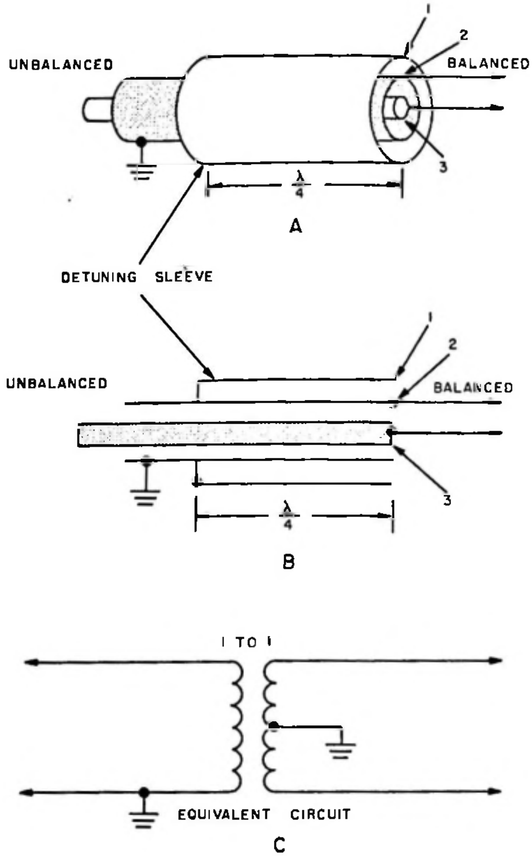


Figure 13-121. Bazooka Line-Balance Converter

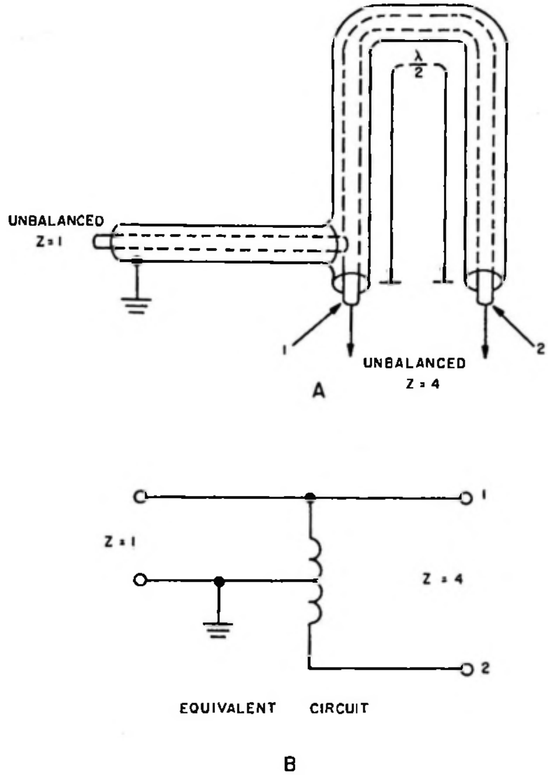


Figure 13-122. Phase Inverter Used as Line-Balance Converter

as the operating frequency does not vary more than a small percent. It may also be operated in the reverse manner, to convert from a balanced to an unbalanced condition.

13-506. Another type of balun or line-balance converter (see figure 13-122) is the half-wave phase inverter, which acts as an impedance transformer with a 4:1 ratio. Since the phase inverter is $\lambda/2$, a negative peak appears at 2 every time a positive peak appears at 1, and since both peaks appear on the inner conductor, they have a high impedance with respect to ground. If each peak has a value of 50 volts, as measured to ground, the voltage across 1 and 2 will be 100 volts. This is a voltage ratio of 2:1, which gives an impedance ratio of 4:1. Like the bazooka converter, the phase inverter may be operated in either direction and is highly efficient only within a narrow range of frequencies.

13-507. HALF-WAVE LINE SECTION AS 1:1 TRANSFORMER. Because a $\lambda/2$ line section or any multiple thereof repeats the input impedance at the output end of the line, it can be used as a 1:1 transformer. If a source having an impedance of 600 ohms is connected to the input end of the line, an impedance of 600 ohms will appear at the output end, no matter what the characteristic impedance of the line may be. The same is true of the load impedance, and its actual value appears unchanged at the input end. Therefore, if a source and a load are matched, and thereby have impedance values that are approximately equal, they may be connected by a half-wave line section. This is convenient where the source and load are approximately matched, but are separated physically by some distance. By using a line that is a multiple of a half-wavelength, the two may be connected without a mismatch. Because no actual matching is done, it is now possible to connect a source and a load whose impedances differ greatly. Coaxial or two-wire line may be used.

13-508. LINE SECTIONS USED FOR TIME DELAY. In electronic apparatus, it may be desirable to have a difference of a fraction of a second between the time an electrical impulse arrives at one point and the time the same impulse arrives at another point. Because of the velocity factor, an electrical impulse is slowed down in a transmission line and requires a definite period of time to travel through a given length (see part A of figure 13-123). Therefore, a time delay can be obtained by using a suitable length of line to connect the two points. You must use the velocity factor to calculate the correct length. Another means for providing the desired time delay is shown in part B of the figure. In this method, the impulse from the source travels a short length of line to arrive at point 1, but has farther to go to reach point 2. This extra distance provides the time delay between points 1 and 2.

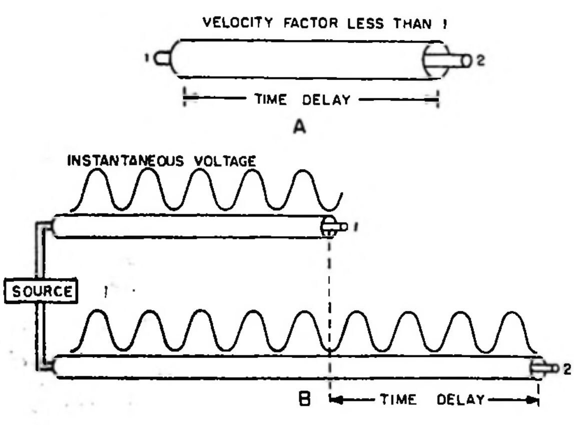


Figure 13-123. Time Delay Obtained with Line Sections

13-509. LINE SECTIONS AS PHASE SHIFTERS. When a time delay between two points is required, it may or may not be necessary to maintain an exact phase relationship between them. When it becomes necessary to produce a definite change of phase, you can use a line section of the appropriate length between the two points, or you can cut the lines to the proper effective difference in the two lengths connecting the source. For example, if a phase shift of 90 degrees is required, the line should be $\lambda/4$ or any number of full wavelengths plus $\lambda/4$. This is important, because it is possible to use any number of full wavelengths of line to introduce a desired time delay, but the phase shift is determined by the fraction of a wavelength left over at the end of the line of full wavelengths.

13-510. COMBINED PHASE SHIFTING AND IMPEDANCE TRANSFORMATION. A half-wave section of two-wire line shorted at both ends, as shown in figure 13-124, is frequently used as a phase shifter and impedance transformer; it is referred to as a half-wave frame. If an rf source is connected at points 1 and 2 of part A of figure 13-124, or any desired impedance points, the standing waves of voltage and current are demonstrated by curves E and I, which, in turn, result in the impedance curve Z. The load is tapped on at whatever location offers the proper impedance match, and the phase shift is determined by the distance in electrical degrees between points 1 and 3 or 2 and 4. The lowest impedance points are at the shorting bars, and the highest are at points 5 and 6, $\lambda/4$ away. This makes possible a wide range of transformation ratios. Beyond 5 and 6 the impedance decreases and

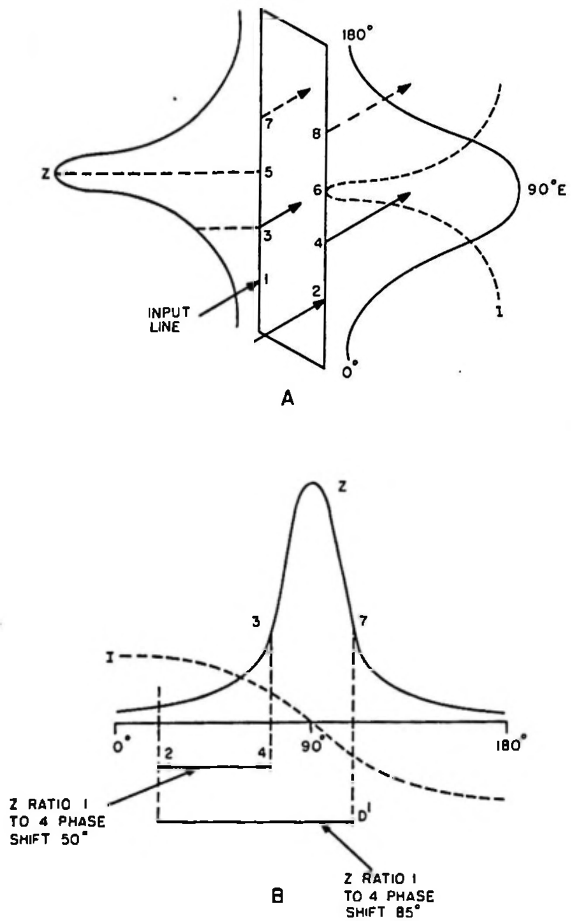


Figure 13-124. Half-Wave Frame Used as Phase Shifter and Impedance Transformer

the current again rises, but in opposite phase. Therefore, there are two points on each conductor which offer the same impedance but different phase relationships. If you move the load to 7 and 8, the impedance is the same, but the phase is not. This is true because the half-wave frame is actually composed of two quarter-wave sections connected in series. Part B of figure 13-124 shows the phase relationships.

13-511. LINE SECTIONS FOR ENERGY STORAGE. A transmission line or a line section may also be used to store pulses of energy in the form of standing waves and deliver them at a desired rate. Delays of approximately 1 to 2 microseconds can be obtained in this manner. Longer times, however, are impracticable, because of the length of line that would be necessary. When longer time delays are necessary, it is possible to construct artificial delay lines, as has been previously discussed, composed of lumped-property capacitance, inductance, and resistance to achieve the same purpose.

13-512. LINE SECTIONS FOR SWITCHING FUNCTIONS. In installations where a transmitter and a receiver use the same antenna, a means must be provided for switching from one to the other. This keeps damaging amounts of power out of the receiver circuits and prevents the loss of feeble incoming signals. Mechanical switching is satisfactory at the low frequencies, but at the higher frequencies it causes losses by introducing discontinuities which set up standing waves in the transmission line. Special gas-discharge tubes are used in conjunction with quarter-wave and half-wave resonant-line sections (see part A of figure 13-125) to solve this problem. At point X, which is close to the transmitter, the line to the receiver is tapped off. On the receiver line, $\lambda/4$ from point X, a tube is connected which short-circuits the line when excited by sufficient rf energy. Another special tube is connected a short distance from X toward the transmitter, at the end of a half-wave line section inserted in one side of the main line. Part B of figure 13-125 shows the action of the switch in the transmitting position. The tube in the $\lambda/2$ section causes a short circuit which is reflected back to complete the main line. The energy from the transmitter then enters the receiver line and causes the tube there to discharge, but this creates a $\lambda/4$ closed-end line section between the tube and point X, which offers a high impedance at the input end. Only enough energy is admitted to keep the gas tube discharging, and none can get through to damage the receiver. As soon as the transmitted energy is interrupted or stopped, both tubes become open circuits, as shown in part C of the figure. Incoming rf signals, which might otherwise be lost in the transmitting circuit, find an open circuit toward the transmitter. Since received signals are too weak to energize either tube, there is no danger that the conditions will be reversed.

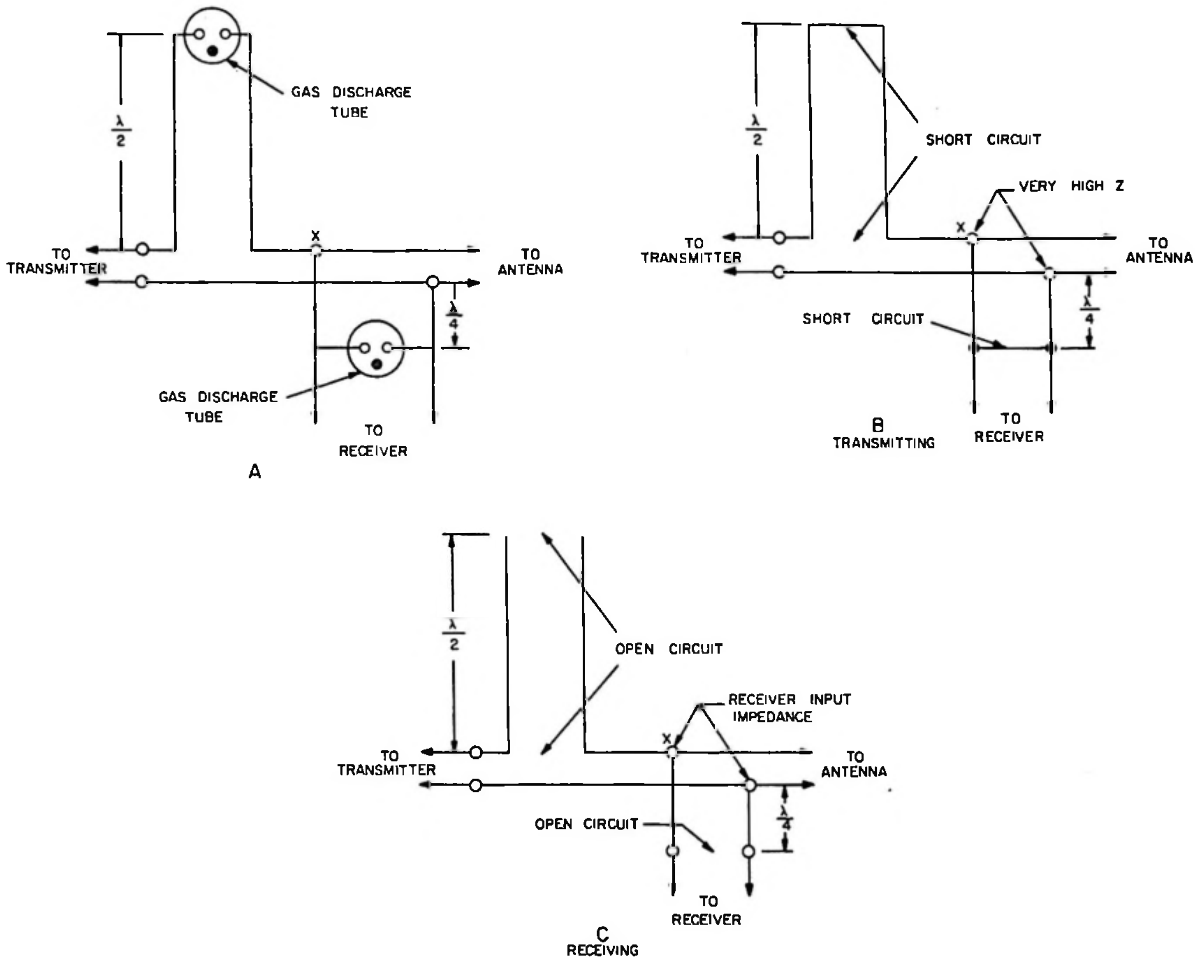


Figure 13-125. Line Sections for Switching Functions

SECTION VIII

ARTIFICIAL TRANSMISSION LINES

13-513. GENERAL.

13-514. An artificial transmission line can be thought of as an electrical network composed of lumped inductance and capacitance, and in certain applications lumped resistance as well. The artificial line has characteristics similar to those of an actual transmission line. Although the primary purpose of a transmission line is to guide the transfer of energy from one place to another, the transmission line has characteristics which may be useful for other applications. For example, if you apply a voltage to the input terminals of a line, a definite amount of time passes before the voltage appears at the output terminals. In other words, the line has the ability to delay voltage and current and the longer the line the greater the delay. If a voltage must be applied to the grid of a tube 1 or 2 microseconds after it has been applied to another part of the circuit, a length of transmission line can be placed ahead of the grid to provide the necessary time lag.

13-515. CIRCUITS.

13-516. Although the delay characteristics and charging action of rf lines make them useful in some radar applications, they cannot be used in others, chiefly because they would have to be excessively long. An actual rf line would have to be several hundred feet long for some uses in radar. Even if a line of such length could be coiled, it would occupy more space than is available in an aircraft. Thus, so far as airborne equipment is concerned, actual transmission lines are out of the question. For this reason artificial lines are used instead of actual lines.

13-517. Artificial lines possess all the characteristics of actual lines, but do not have the bulk and physical length of actual lines. They are constructed by first determining the capacitance and inductance of the desired line, and then lumping these variables into equivalent inductors and capacitors. The result is an electrically equivalent, but physically different, line.

13-518. Part A of figure 13-126 shows an actual two-wire line. Part B of the figure shows the equivalent circuit for each section as it is when the section is split into two parts and its characteristics measured. Since both sections have identical characteristics, the two can be joined together (see part C of figure 13-126). The final circuit, shown in part D of the figure, lumps the inductive and capacitive characteristics of each section and simulates the artificial line. As you can see, the artificial line performs just as an actual line would perform, but it occupies much less space.

13-519. In any line, time is required for a voltage change to travel the length of the line. This time characteristic makes it possible to slow down the transfer of a voltage change in its travel from one circuit to another.

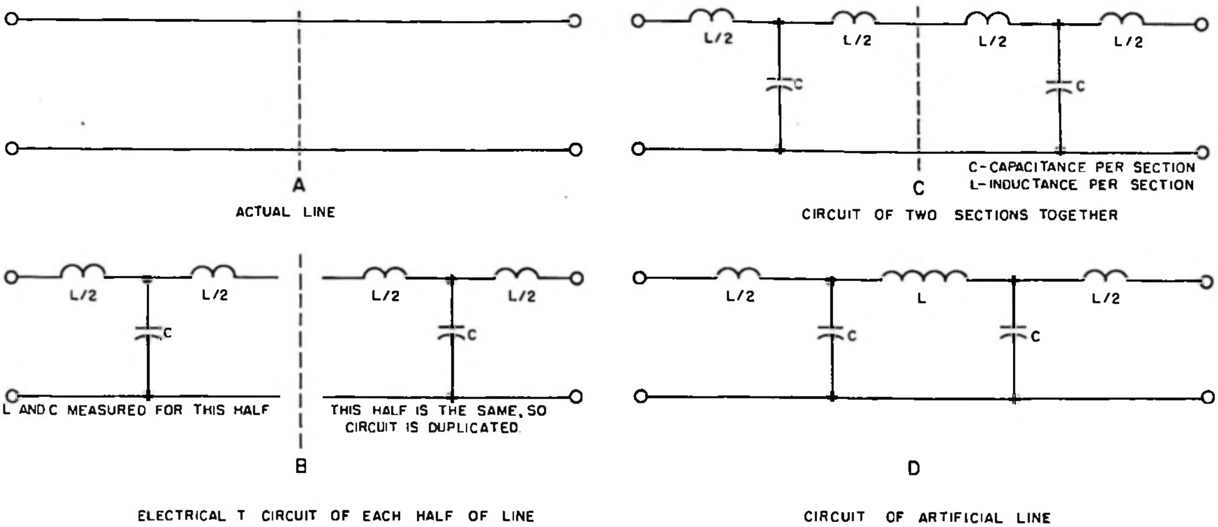


Figure 13-126. Design of Artificial RF Line

13-520. The artificial line shown in figure 13-127 provides a means of delaying a voltage square-wave pulse 10 microseconds. This line is constructed with an inductance of .25 millihenry in conjunction with one half of the lumped 0.5-millihenry inductance, to create a total inductance of 0.5 millihenry working with a capacitance of 0.05 microfarad in each of the two sections. To compute the time for a voltage change to move through one section, use the equation $T = \sqrt{LC}$. The equation for time delay for the entire line is:

$$T_d = N\sqrt{LC}$$

where:

T_d = time delay in seconds

N = number of sections

L = inductance per section in henrys

C = capacitance per section in farads

To calculate the time delay for the entire artificial line shown in figure 13-127:

$$T_d = N\sqrt{LC} = 2 \cdot \sqrt{0.5 \cdot 10^{-3} \cdot 0.05 \cdot 10^{-6}}$$

$$= 2 \cdot \sqrt{5 \cdot 10^{-4} \cdot 5 \cdot 10^{-8}}$$

$$= 2 \cdot \sqrt{25 \cdot 10^{-12}}$$

$$= 2 \cdot 5 \cdot 10^{-6}$$

$$= 10 \cdot 10^{-6} \text{ seconds}$$

$$= 10 \text{ microseconds}$$

13-521. In one radar set the time base is generated in such a way that it does not start until after the first radar echo returns to the receiver. For these signals to be displayed at their correct range mark, they are sent through an artificial rf delay line, which delays the signals just enough that the very first returning signal does not arrive at the cathode-ray tube until the time base has begun to form. Although a large number of video frequencies are present in the signal going through the line, all of them are delayed the same amount.

13-522. Another phenomenon of rf lines is the manner in which the lines charge and

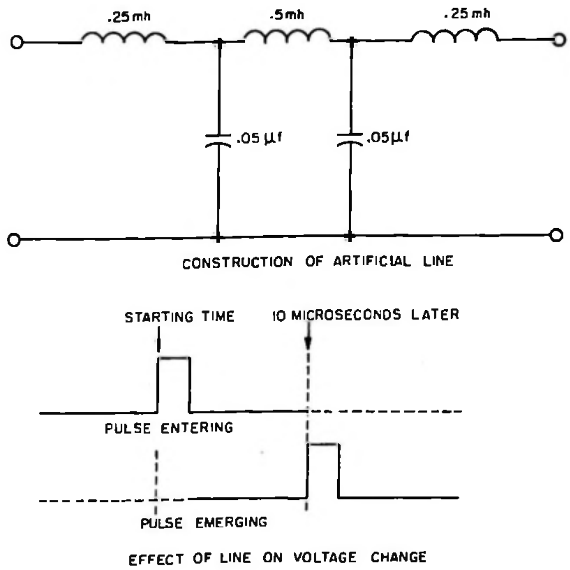


Figure 13-127. Artificial Delay Line

discharge when a dc voltage is applied to them. When dc is applied to a line in which the source impedance is matched to the line impedance, half the source voltage appears across the line impedance at the time the battery is connected to the line. A voltage change is thereby produced across the line. Figure 13-128, which shows the oscilloscope picture at the sending end, illustrates this condition. This voltage change travels down the line, charging the line as it goes. When the voltage change reaches the open end, it is reflected and starts back along the line.

13-523. All capacitors charge to half the battery voltage when the voltage change is going down, and to full battery voltage when the voltage rise is coming back. The oscilloscope picture in figure 13-128, at the center of the line, illustrates this condition. This change in voltage is commonly called a step voltage. When all capacitors have been charged to a value equal to the battery voltage, current stops flowing from the battery. By observing the oscilloscope picture which illustrates the end of the line, you will note that the end of the line has a continuous rise in voltage which equals the applied voltage.

13-524. When this line is discharged into a resistance equal to its characteristic impedance, the result is a square wave that has a constant amplitude and a duration equal to twice the time for the pulse to travel the length of the line. The amplitude is equal to half the charge voltage because the line discharges through its own impedance and the load impedance, Z_R, which are connected in series. This cuts the load voltage by half and produces a drop in voltage that immediately travels down the line and back to the starting place, as shown in figure 13-129. As the end of the line is open, the reflection there is in phase. This produces a voltage across the load impedance, Z_R, equal to half the charge voltage, and which lasts for twice the time of one-way travel on the line.

13-525. Usually, an artificial line is constructed to generate this square wave. When the L and C for each section are known, the time duration of the square pulse, t_p, is:

$$t_p = 2N\sqrt{LC}$$

The following are the calculations for the time duration, both forward and rearward, of the pulse generated in the artificial line shown in figure 13-127:

L = 0.5 mh per section

N = .05 µf per section

N = 2 sections

$$t_p = 2N\sqrt{LC} = 2 \cdot 2 \cdot \sqrt{0.5 \cdot 10^{-3} \cdot .05 \cdot 10^{-6}}$$

$$= 4 \cdot \sqrt{25 \cdot 10^{-12}}$$

$$= 4 \cdot 5 \cdot 10^{-6}$$

$$= 20 \cdot 10^{-6}$$

= 20 microseconds

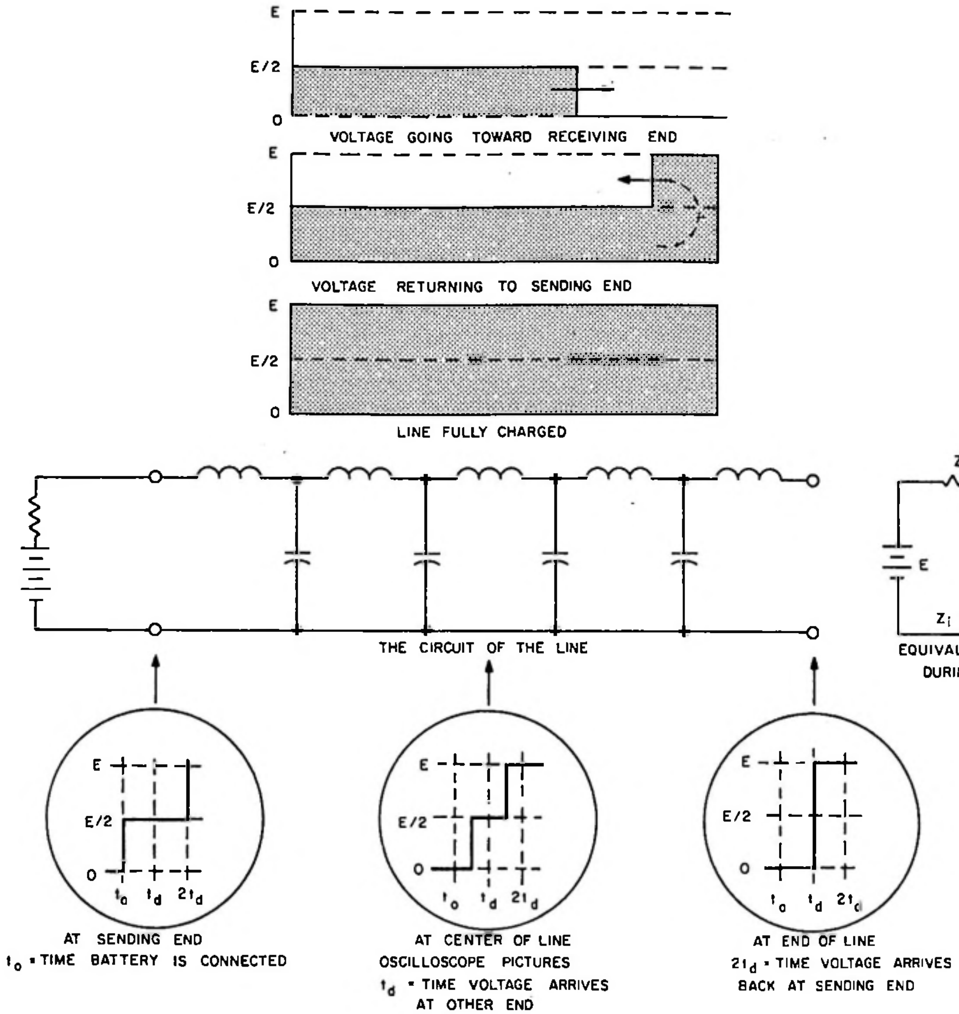


Figure 13-128. Mechanics of Charging Artificial Line

When an artificial line is not terminated in its characteristic impedance, Z_o, several trips (reflections) are required for complete discharge. This means that discharge cannot be completed in a single round trip.

13-526. In figure 13-129 there are three oscilloscope pictures that show the discharge action of an artificial line across Z_R. When Z_R is large, the pulse is tall. But this is followed by several smaller steps, called tails. These tails exist because the line does not completely discharge during the first trip, and more trips are necessary for complete discharge. When Z_R is small, the change that is reflected not only cancels the voltage left on the line, but also recharges the line to a smaller voltage in the opposite direction, producing an oscillatory output waveshape. To calculate the Z_o value of the line, use the following formula:

$$Z_o = \sqrt{\frac{L}{C}}$$

The calculations for finding the Z_o of the artificial delay line shown in figure 13-127 are as follows:

$$Z_O = \sqrt{\frac{L}{C}} = \sqrt{\frac{0.5 \cdot 10^{-3}}{0.05 \cdot 10^{-6}}}$$

$$= \sqrt{\frac{5 \cdot 10^{-4}}{5 \cdot 10^{-8}}}$$

The characteristic impedance is as follows:

$$Z_O = \sqrt{10^4} = 10^2 = 100 \text{ ohms}$$

13-527. An artificial pulse line is capable of producing a nearly perfect square-wave pulse which has an amplitude of several thousand volts. For purposes of comparison, a vacuum-tube multivibrator and amplifier circuit that could provide a similar output would weigh 100 pounds and occupy a space of several cubic feet. On the other hand, a pulse-forming line of equal capacity would consist of a few inductors and capacitors and occupy a space of only about four cubic inches.

13-528. Figure 13-130 shows a simplified radar transmitter which uses a pulse line. In it a conventional power supply charges the pulse line through a diode tube to 4000 volts. The diode prevents the line from discharging through the power supply. When you close the switch, a 1-microsecond, 2000-volt pulse is generated. The step-up transformer increases this 2000 volts several times. This voltage serves as the cathode voltage for the transmitter; the plate of the transmitter tube is at ground potential. The transmitter oscillates during the time cathode voltage is applied, a period of 1 microsecond. Thus, a radio-frequency signal of 1-microsecond duration (or a radar pulse) is radiated by the antenna.

13-529. A pulse can also be formed by a short-circuited line. The operation of the line depends upon the energy stored in the magnetic fields around the inductors in the

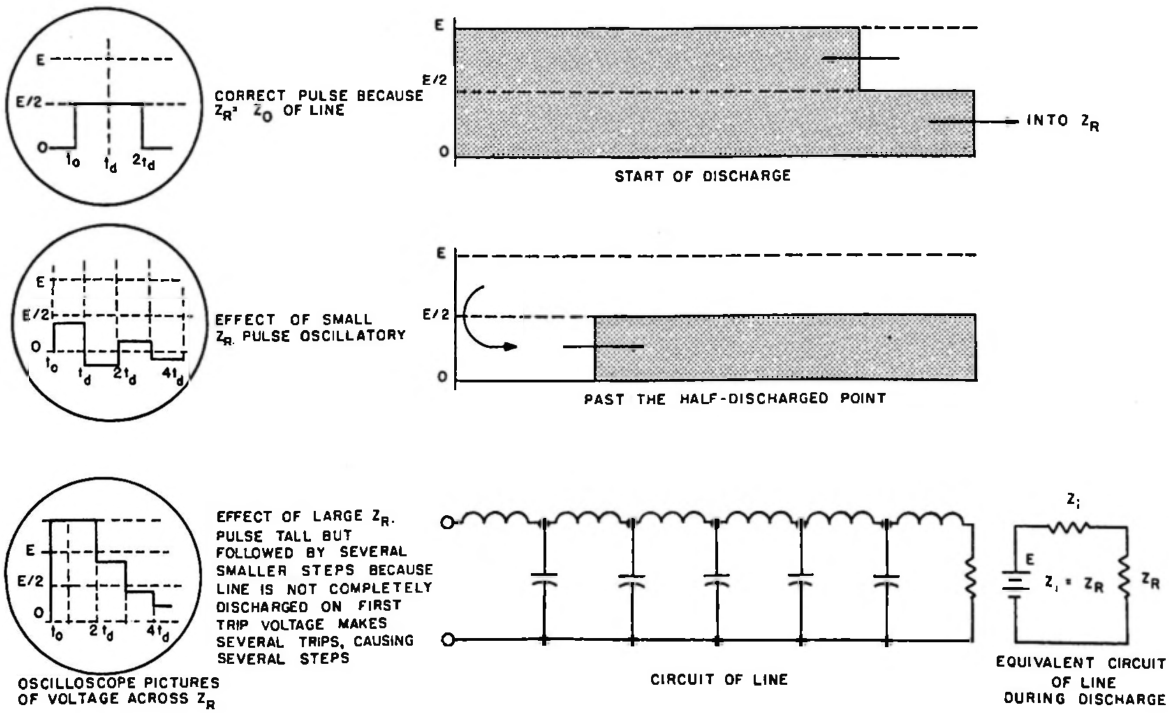


Figure 13-129. Discharging Artificial Line Through Z_R

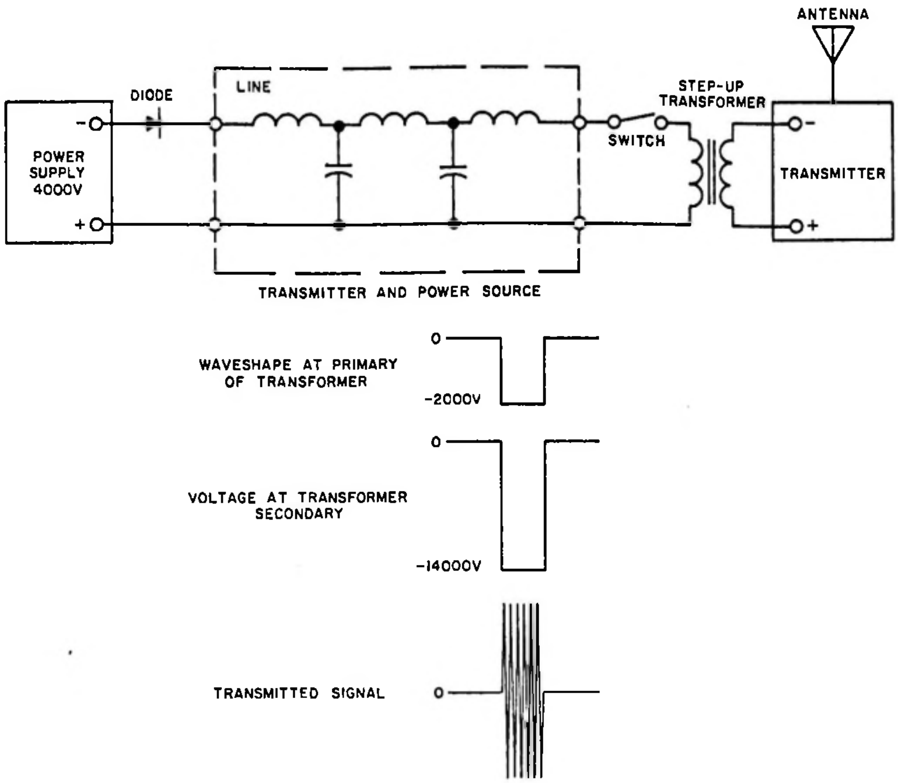


Figure 13-130. Simple Circuit Showing Pulse Line Used in Radar Transmitter Circuit

line. The line is charged from a constant-current source, and produces a voltage across the load, R_L. Part A of figure 13-131 shows a short-circuited line during charge. At the instant the constant-current source is connected to the line (see T1 in part B of figure 13-131), a current, I_o, flows. If R_L is equal to the characteristic impedance of the line, Z_o, the current is divided equally between the line and R_L. A wave of voltage whose amplitude is one-half I_oZ_o travels down the line, and is reflected back up the line with a polarity which is opposite that at the short-circuited end. This reflected wave travels back to the input end.

13-530. The voltage pulse generated at the load during the charging period has an am-

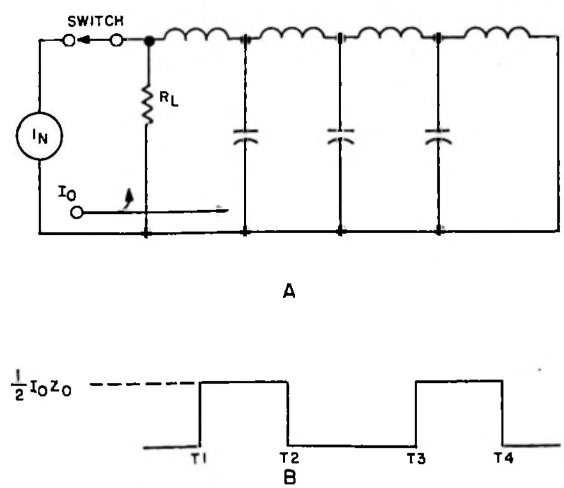


Figure 13-131. Delay Network Details

plitude of $\frac{1}{2}$ IoRL. It lasts the length of time it takes the pulse to travel to the end of the line and back. This is the pulse shown from T1 to T2 in part B of figure 13-131.

13-531. The voltage across R_L is now zero. (The line is effectively shorting the load.) A constant current, I_o, is flowing through the inductances of the line, and the line is fully charged with magnetic energy. The current flowing down the line during charge is reflected in-phase, and the incident current and reflected current add, doubling the current on the line.

13-532. At time T3 the constant-current source is disconnected from the transmission line. A voltage pulse that is exactly similar to the charging voltage pulse is developed across R_L as the magnetic energy stored on the line is discharged. The duration of the pulse is the time it takes the current to travel down the line and back (T3 to T4). At time T4 the line is discharged completely. As the magnetic fields about the inductances collapse, half the current ($\frac{1}{2}I_o$) flows through the load and half charges the capacitors. As the current tends to decrease through the inductors, the capacitors discharge, maintaining the current constant to the end of the pulse.

13-533. SUPERSONIC DELAY LINE.

13-534. The supersonic delay line has two generator purposes: (1) to delay an electrical signal, such as a trigger or a range mark, and (2) to delay a pulse carrier without introducing distortion. In present-day radar methods, comparison of video signals with each preceding signal is a necessity for efficient Moving Target Indicator (MTI) operation. Recent advancements in the development of electronic equipment have indicated an increasing need for triggers whose timing cannot vary more than 0.5 microsecond.

13-535. The principal elements of a supersonic delay line are the transmission media and the electromechanical transducers. In the supersonic delay line, electrical impulses are changed into sound impulses; after they have traversed a transmission medium, they are converted back into electrical signals. Since sound waves travel more slowly than electrical waves, delay is introduced. For frequencies in the megacycle range, quartz crystals are the most satisfactory transducers. Liquids (water and mercury) and solids (fused quartz) are normally used as the transmitting medium.

13-536. LIQUID MEDIA.

13-537. Figure 13-132 illustrates a typical mercury delay line. The signal is applied to a quartz crystal, which is the electromechanical transducer. (A transducer is a device that converts electrical energy into mechanical energy, or vice versa.) The crystal undergoes changes in thickness due to the piezoelectric effect; thus, electrical energy is converted into mechanical energy (vibrations). The vibrations of the crystal are communicated to the mercury as supersonic, or ultrasonic, waves. The waves travel down the tank with the mercury acting as a transmission medium.

13-538. On reaching the end of the tank, the waves are reflected by a reflecting plate. The reflectors consist of circular surfaces arranged on the two opposing ends of the tank. They are formed in the walls of the tank by grooves cut around these areas. If the reflecting surfaces make good contact with the mercury, they will tend to absorb the mechanical vibrations. For maximum reflection capabilities, a rough surface is used, with air trapped in the pockets of the rough surface area. This provides a mismatch to the mercury that causes almost

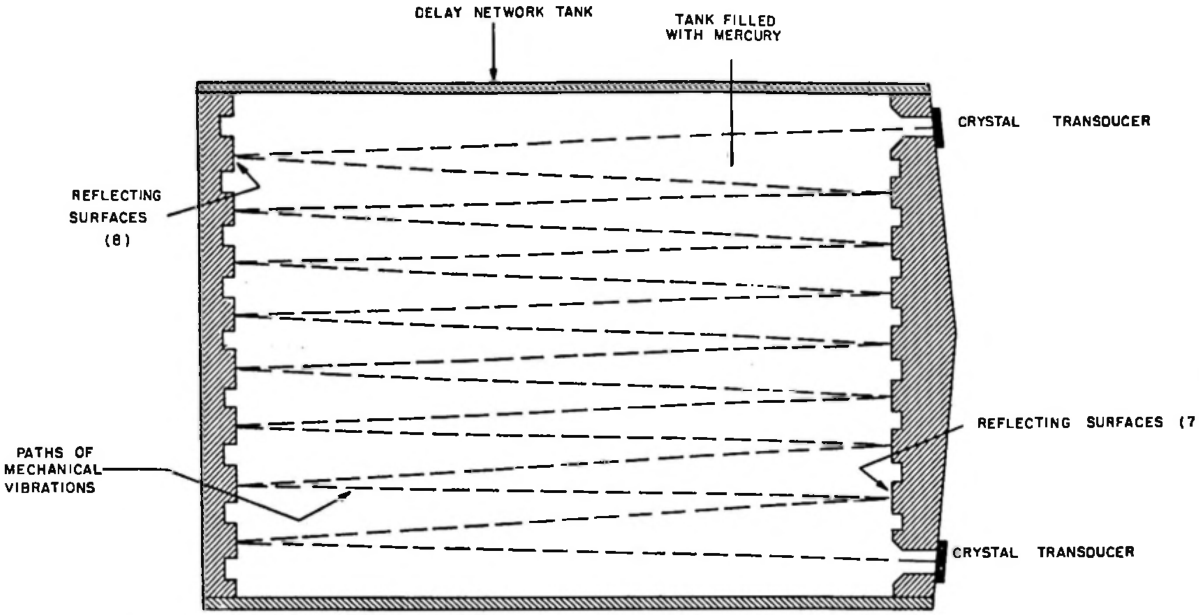


Figure 13-132. Mercury Delay Line

perfect reflection. This process is repeated over several paths. The waves finally strike a crystal at the opposite end of the tank and produce electrical impulses. The piezoelectric effect here changes the mechanical energy back to electrical energy.

13-539. In mounting the crystal, you must be careful to support and protect it so that it will not bend or break. The crystal must be aligned accurately with the tube axis. Provision must be made to attach electrodes to the crystal surface so that there will be no interference with the crystal's sonic operation. Elimination of unwanted multiple echoes is a requirement for lines that must reproduce the pulse shape accurately. This is accomplished by reducing the time the signal is applied to the delay line so that the remaining wave can be damped out. The thickness vibration of the crystal is usually at the crystal's lowest resonant frequency. Some crystals are backed by steel on the sides that are not in contact with the mercury. In other crystal mountings the crystal is backed by mercury, which makes plating unnecessary on that side. The large acoustic impedance of the mercury likewise loads the crystal and widens its bandpass characteristic. In the mercury delay line shown in figure 13-132, direct mercury-to-crystal contact is used at both the input and output of the delay line. The mercury acts as the ground side of the electrical input and output circuits. Insulated plates on the opposite sides of the crystals provide the other electrical connections for crystal operation.

13-540. SOLID MEDIA.

13-541. Fused-quartz delay lines are mechanical delay circuits that depend upon the time it takes mechanical ultrasonic waves to pass through the fused quartz. Normally, each quartz delay line is contained in a hermetically sealed shell. Figure 13-133 shows the multiple-symmetry type of fused-quartz delay line. Transducer crystals with a resonant frequency near the frequency of the carrier signal convert electrical input waves to supersonic mechanical waves.

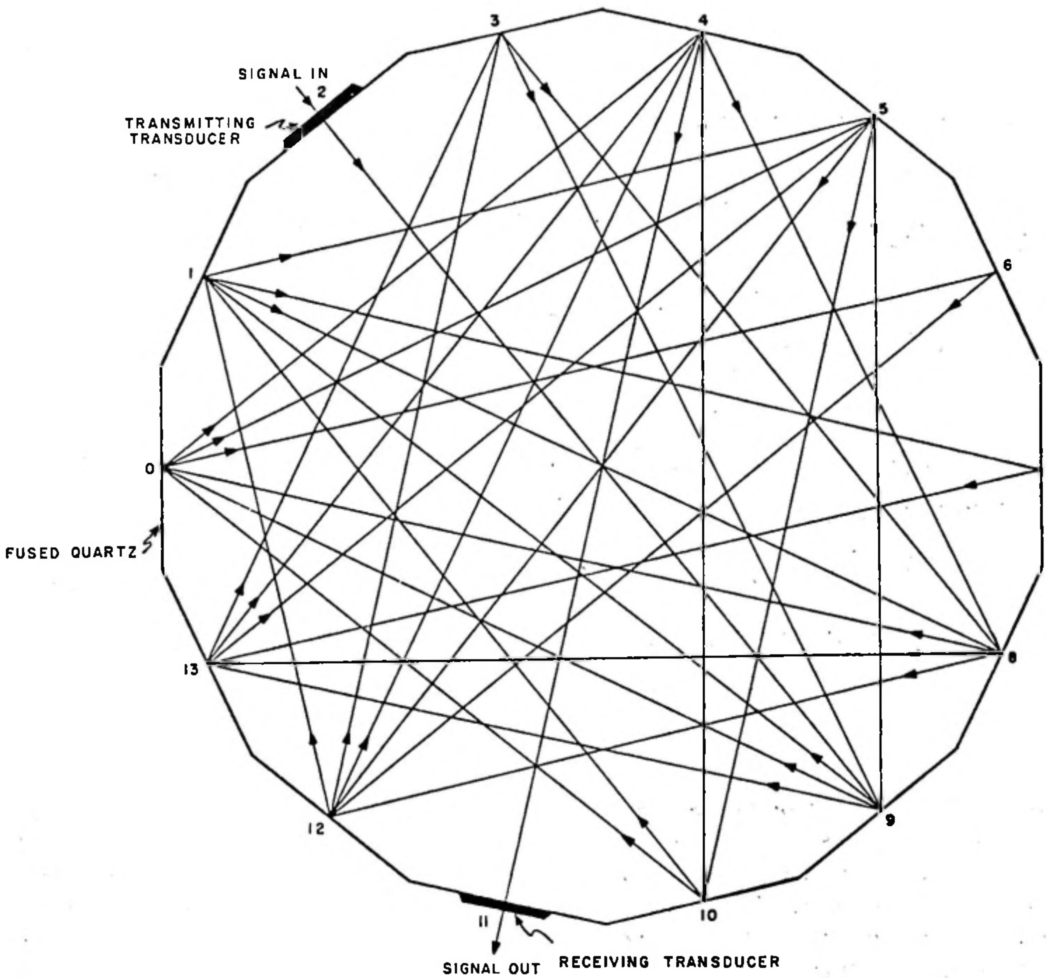


Figure 13-133. Quartz Delay Line

13-542. A polygon-shaped quartz block is used to provide surfaces from which the signals are reflected. The facets, or sides, of this polygon are precision-ground at the necessary angles to obtain the desired signal reflection path. Figure 13-133 illustrates the theoretical polygon-shaped quartz crystal. The sides and angles of an actual fused-quartz delay line of 14 sides would vary. In figure 13-133, the signal is fed in at side 2 and fed out at side 11. Some of the facets of the multiple-symmetry quartz delay line reflect the signal as many as three times. The figure shows only one of the possible paths which the signal may use.

13-543. Fused-quartz delay lines of the multiple-symmetry type and mercury delay lines are used because of their ability to delay signals faithfully by much longer times than electrical delay lines. In the propagation of acoustical waves, there is much less phase and amplitude distortion than in the electromagnetic waves. Such delay lines can also be constructed to withstand severe conditions of temperature and pressure, and may be stored indefinitely without any harmful effects. Fused-quartz delay lines may be mounted in any position for operating. One factor which limits the operation of fused-quartz delay lines is that spurious secondary signals are produced in the line because the signal travels paths other than that corresponding to the desired delay time; these signals must be attenuated to a level at which they can be ignored. Another important factor concerning the use of the fused-quartz delay line is the amount of signal attenuation in the line. Some radar systems have a requirement of less than 60-db signal attenuation.

13-544. MAGNETOSTRICTION DELAY LINE.

13-545. The basic material used in the construction of a magnetostriction delay line is a nickel ribbon or tube. Kinking should be avoided, because serious reflections occur at kinks or dents. The driver coil is approximately the same as the pickup coil, the pickup coil having more turns than the driver coil. A nonmetallic, nonmagnetic material is used as a coil mount, with the coil centered very accurately on the coil form. The output coil requires a polarizing magnetic field for proper operation. You should prevent undesired end reflections by some means of mechanical dampening (usually a rubber mount).

13-546. The main theory of ferromagnetism is based on the assumption that a ferromagnetic crystal is composed of a large number of regions, each of which is spontaneously magnetized to an intensity that is a function of the temperature alone. When the net over-all magnetization is zero, these regions are aligned in equal numbers along the various directions of easy magnetization of the crystal. The effect of a magnetizing field is to alter this uniform distribution so that most of the regions lie in that direction of easy magnetization which most nearly coincides with the field.

13-547. The application of a magnetic field increases the stability of the atoms aligned with the field, and decreases the stability of other orientations. As the field intensity increases, the atoms with unstable orientation shift suddenly. The corresponding sudden changes in the interatomic force of the face-centered cubes of nickel crystals set up elastic vibrations that act upon adjacent atoms and are thereby transmitted. The sum of a large number of these sudden local shifts in orientation results in the observable magnetostriction effect, which is used in present-day electronic equipment.

13-548. The magnetostriction effect may be described as the change in dimensions of a ferromagnetic material when it is subjected to a longitudinal magnetic field. The maximum variations occur in the length of the material, with smaller changes occurring in the volume of the material.

13-549. In figure 13-134, V1 is a twin triode whose grids are biased beyond cutoff. A positive pulse drives both grids of V1 positive, causing a short pulse of current to flow in each plate circuit. Each plate circuit of the twin triode contains three coils, with a set of relay contacts in parallel with each coil. The driver coils are wound around the magnetostriction line of nickel tubing. The relays are remotely controlled. If any relay contact is opened, the plate current pulse flows through the associated coil and creates a magnetic field along the longitudinal axis of the nickel tubing. This field sets up a shock wave in the tubing because of the magnetostriction effect.

13-550. A mechanical stress is felt along the tubing in the vicinity of the driver coils. The shock wave travels from its source toward the end of the line with a velocity of 0.187 inch per microsecond. When this shock wave reaches the end of the line, most of the energy is absorbed in the rubber tubing that is clamped around the end of the line. The driver coils may be spaced so that the time required for the shock wave to travel from one coil to another meets the time requirements of the equipment used.

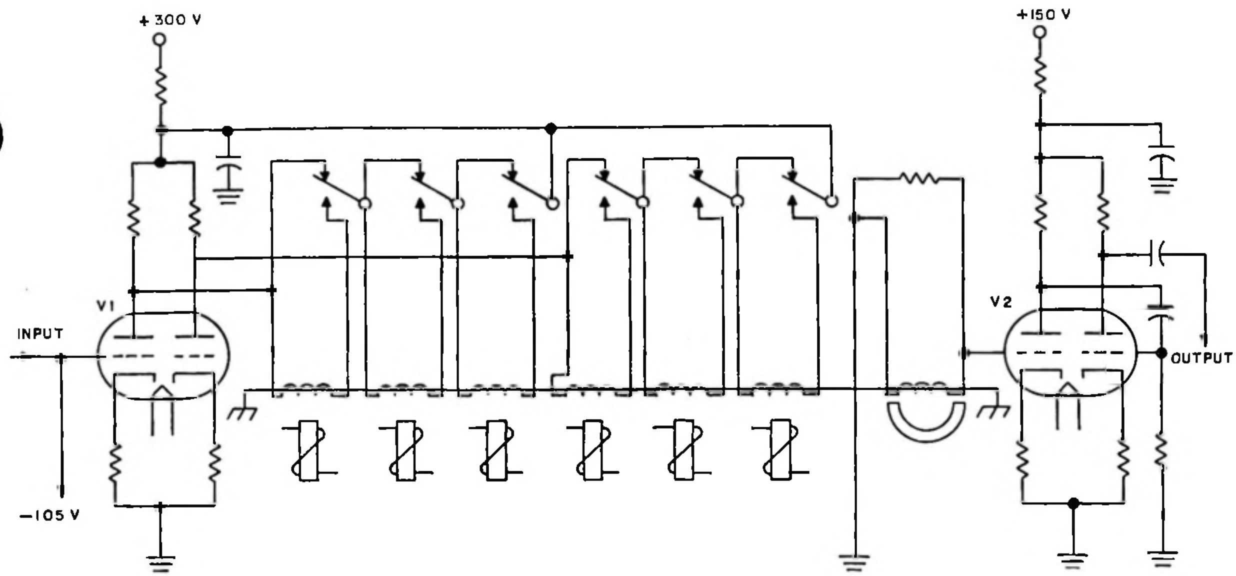


Figure 13-134. Magnetostriction Delay Line

13-551. A permanent horseshoe magnet mounted near the tubing provides a magnetic field in the vicinity of the pickup coil. The shock wave, which actually compresses the nickel tubing, travels down the tubing and shifts the lines of flux created by the permanent magnet. As the lines of flux cut the pickup coil, a voltage is induced. Because of the short duration of the shock wave and the direction of the magnetic field, a short negative pulse is applied to the grid of V2. The magnetostriction delay line has been satisfactorily developed as a flexible device for delaying pulses, for the mixing of pulses, and for generating codes.

SECTION IX

TRANSMISSION LINE MEASUREMENTS

13-552. NEED FOR TRANSMISSION-PERFORMANCE MEASUREMENTS.

13-553. A transmission line may cause deterioration of a transmitted signal in many ways—by frequency versus attenuation characteristics of a transmission line; by attenuation of its power; by distortion of its waveshape; and by permitting interference. You must devise methods to minimize these types of deterioration.

13-554. Suppose that a repeater is to be installed somewhere on the line to make up for the power loss. The amount of amplification to be provided by the repeater should be determined by the amount of power loss. It would be inefficient to keep rebuilding a repeater until it provided the necessary amount of amplification; however, if the amount of power loss could be ascertained, the required amount of amplification would be known, and it would be easy to build the repeater to provide the amount of amplification necessary to make up the loss. To do this, there must be a way to measure and express numerically the power loss, distortion, and crosstalk of a transmission line, the power gain of an amplifier, and the performance characteristics of components such as filter and equalizer networks that might be used with them.

13-555. UNITS OF TRANSMISSION MEASUREMENTS.

13-556. BEL. The ratio of the input and output powers was formerly expressed in terms of a unit called the bel, a name derived from Alexander Graham Bell, the inventor of the telephone. With this unit, the output power level is said to be 1 bel below the input level when the input and output powers are in the ratio of 10 to 1. In this case, the bel is used as a unit of attenuation. If the output power level is 10 times the input power level, it is said to be 1 bel above the input. In this case, the power ratio of input to output is 1 to 10, and the bel unit is a gain.

13-557. DECIBEL.

13-558. The bel proved to be an inconveniently large unit. It was found desirable to use a unit which more closely expresses commonly encountered (smaller) power ratios without using fractional parts of the unit. For this reason, the decibel has been universally adopted. As its name implies, a decibel (db) is one-tenth of a bel. Today, power ratio is expressed as a definite number of decibels. Because the decibel is a unit one-tenth as large as the bel, the number expressing a power level in decibels (abbreviated N_{db}) is 10 times as large as the number of bels. The formula for determining power level in decibels is: Number of decibels, $N_{db} = 10 \log_{10} \frac{P_2}{P_1}$ or P_1 over P_2 for special reasons discussed later. Applying this formula, a power ratio of 100 to 1 is expressed as 20 decibels, or 20 db, instead of 2 bels; a power ratio of 2 to 1 is expressed as 3.01 decibels, instead of 0.3010 bel.

13-559. It should be noted that although the decibel is one-tenth of a bel, the power ratio expressed by 1 decibel is not one-tenth of the power ratio expressed by 1 bel. One decibel expresses a power ratio of 1.26, because the logarithm of 1.26 is 0.1, and the number of decibels is 10 times this logarithm.

13-560. MEANING OF DB. The decibel is a unit of power ratio—not a unit of power. When a transmission line causes an attenuation of 10 db, the output power is only one-tenth of the input power. The attenuation figure does not state the amount by which the power has changed. If the input power is 100 watts and the output power is 10 watts, the line has absorbed 90 watts of power. If the input power is 1 watt and the output power is 0.1 watt, the line has absorbed 0.9 watt of power. In both cases, the power ratio is 10 to 1, and the attenuation is 10 db.

13-561. USE OF DECIBEL.

13-562. CONVERSION OF POWER RATIO TO DECIBELS. The decibel, like the bel, may be used as a unit of either attenuation or gain, expressing the decrease in power caused by a transmission line or a piece of equipment, or the increase in power caused by an amplifier. In order to distinguish whether a given number of decibels of power change refers to gain or loss, the number of decibels is customarily preceded by a plus sign or a minus sign. Thus, a power change of +6 db refers to a gain, and a power change of —6 db refers to a loss.

13-563. In converting a power ratio into its decibel expression, the following two rules should be followed. Rule 1: In the formula $N_{db} = 10 \log \frac{P_1}{P_2}$, choose the larger power as P_1 and the smaller power as P_2, regardless of whether the larger power is at the input or the output. The power ratio will then always be greater than 1, and its logarithm will be a positive number. Rule 2: Prefix to the answer a plus sign if the power change is a gain, or a minus sign if the power change is a loss.

13-564. For example, assume that the power input to a transmission line is 10 milliwatts, and the power output is 2.46 milliwatts. By rule 1, the larger power (the input) is P_1, and the smaller power is P_2. The number of decibels of power change is then expressed as follows:

$$N_{db} = 10 \log \frac{P_1}{P_2} = 10 \log \frac{10.00}{2.46} = 10 \log 4.06$$

(Note that the subscript 10, denoting the base of the common system of logarithms, is omitted. This is general practice when no misinterpretation is possible, and is followed throughout). Reference to a table of common logarithms shows that the logarithm of 4.06 is 0.6085. Hence:

$$N_{db} = 10 \log 4.06 = 10 \cdot 0.6085 = 6.085$$

Since the input power is greater than the output power, the power change is clearly one of attenuation. By rule 2, the answer is prefixed by a minus sign. Thus:

$$N_{db} = -6.085 \text{ db}$$

13-565. As an example of a power gain, consider a repeater amplifier having an input power of 2 milliwatts and an output power of 400 watts. By rule 1, the larger power (the output) is P_1, and the smaller power (the input) is P_2. Both powers must be expressed in the same units. Hence, the input power of 2 milliwatts is expressed as 0.002 watt. The power change is then calculated as follows:

$$N_{db} = +10 \log \frac{P_1}{P_2} = +10 \log \frac{400}{0.002}$$
$$= +10 \log 200{,}000$$
$$= +10 \cdot 5.3010 = +53.01 \text{ db}$$

In this example, as a short cut, the plus sign (indicating that the power change is a gain) is inserted at the start, and carried through to the final answer.

13-566. It should be noted that whether a power change is a loss or a gain, it is necessary to know the power ratio. In the case of an amplifier, the power gain is the power ratio required. The voltage gain of an amplifier is not the same as the power gain.

13-567. CONVERSION OF DECIBELS TO POWER RATIO. If it is required to find the power ratio of a circuit for which the gain or loss in decibels is known, a reversal of the procedures described for the preceding paragraphs can be used. The following rule applies. Rule 3: If the number of decibels is positive, the circuit has a power gain, and the output power is greater than the input power; if the number of decibels is negative, the circuit has a power loss, and the output power is less than the input power.

13-568. As one example of how to determine the ratio of output and input powers when the ratio is expressed in decibels, assume that there is a circuit known to have a power change of +12 db. Inserting this value in the formula, and dividing both sides by 10:

$$12 \text{ db} = 10 \log \frac{P_1}{P_2}$$

$$1.2 = \log \frac{P_1}{P_2}$$

A table of logarithms shows that 15.85 is the number which has the logarithm 1.2. Therefore:

$$\frac{P_1}{P_2} = 15.85$$

Since the number of decibels is given as positive, it is known from rule 3 that the circuit has a gain, and that its output power is 15.85 times its input power.

13-569. As a second example, consider a power change of —25 db. The minus sign is disregarded temporarily, and the power ratios are determined in the manner already explained, as follows:

$$25 \text{ db} = 10 \log \frac{P_1}{P_2}$$

$$2.5 = \log \frac{P_1}{P_2}$$

$$\frac{P_1}{P_2} = 316.2$$

Since the number of decibels has a minus sign, the circuit attenuates the power, and its output power is less than its input power by a ratio of 1 to 316.2.

13-570. ADDITION OF POWER CHANGES. The total number of decibels of attenuation or gain of a circuit is the algebraic sum of the attenuations and gains of its stages or sections, each expressed in decibels. For example, consider the transmission-line network of figure 13-135. There is a power loss of —10 db in the first section of the line, a power loss of —20 db in the equipment between the sections of the line, a power loss of —10 db in the second section of line, and a power gain of +30 db in the repeater. The total power change is:

$$N_{db} = -10 \text{ db} - 20 \text{ db} - 10 \text{ db} + 30 \text{ db} = -10 \text{ db}$$

The total power change is —10 db, indicating a net power loss in a ratio of 10 at the input to 1 at the output.

13-571. Reference to table 13-3 provides an explanation of this result. Each —10-db loss represents a power-loss ratio of 10 to 1. The —20-db loss represents a power

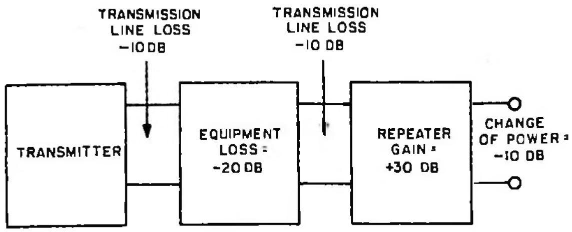


Figure 13-135. Power Changes in Transmission-Line Network

loss ratio of 100 to 1. The +30-db gain represents a power gain ratio of 1 to 1000. The over-all loss is the product of these ratios:

$$\text{Over-all loss} = \frac{10}{1} \times \frac{100}{1} \times \frac{10}{1} \times \frac{1}{1000} = \frac{10}{1}$$

An over-all loss in a ratio of 10 to 1 is expressed as —10 db, the same result obtained by algebraic addition of the primarily calculated decibel losses and gains of the circuit sections.

13-572. TABLE OF DECIBELS.

13-573. In order to eliminate the need for extensive calculations and reference to a table of logarithms each time a power ratio must be expressed in terms of decibels, reference may be made to a table, such as table 13-3, which gives power ratios directly from decibels, and vice versa.

13-574. The manner in which table 13-3 can be used is illustrated by a consideration of all the figures on one line.

Table 13-3. Power Ratios

NUMBER OF DECIBELS	LOSSES		GAINS	
	POWER-RATIO FACTOR	% POWER LOSS	POWER-RATIO FACTOR	% POWER GAIN
1	0.79	21	1.26	26
2	.63	37	1.6	60
3	.5	50	2.0	100
4	.4	60	2.5	150
5	.32	68	3.2	220
6	.25	75	4.0	300
7	.2	80	5.0	400
8	.16	84	6.0	500
9	.126	87.4	7.9	690
10	.1	90	10.0	900
20	.01	99	100.0	9,900
30	.001	99.9	1,000.0	99,900
40	.0001	99.99	10,000.0	999,900

a. Consider the case for which $N_{db} = 3$:

$$3 = 10 \log \frac{P_1}{P_2}$$

$$.3 = \log \frac{P_1}{P_2}$$

.3 is the logarithm of the number 2; consequently:

$$\frac{P_1}{P_2} = 2$$

The power-ratio factor (table 13-3) is the ratio of the output and input power. In the example above, if there is a power gain, the power-ratio factor is 2; the output power P_1 is greater than the input power P_2 in a 2-to-1 ratio. If there is a power loss, the power-ratio factor is 0.5, because the output power must be less than the input power in a 1-to-2 ratio. The table quickly gives the power-ratio factor for both gains and losses.

b. Both the percentage-power loss and the percentage-power gain are the ratio of the difference between the input and output powers, and the input power, expressed as a percentage. Example of percentage-power loss: The input power to a transmission line is 10 milliwatts, and the output at a receiving switchboard is 5 milliwatts. The percentage-power loss is:

$$\frac{\text{Input power - output power}}{\text{Input power}} \times 100$$

Substituting:

$$\frac{\text{10 milliwatts - 5 milliwatts}}{\text{10 milliwatts}} \times 100 =$$

50 percent

The percentage-power loss is 50 percent. (The power-ratio factor is 0.5, and there is a 3-db power loss.) Example of percentage-power gain: The input power to a transmission line is 7 milliwatts, and the output is 14 milliwatts. The percentage-power gain is:

$$\frac{\text{Output power - input power}}{\text{Input power}} \times 100$$

Substituting:

$$\frac{\text{14 milliwatts - 7 milliwatts}}{\text{7 milliwatts}} \times 100 =$$

100 percent

The percentage-power gain is 100 percent. (The power-ratio factor is 2, and there is a 3-db power gain.)

13-575. Table 13-3 can also be used for figures that do not appear in it. For example, consider a power change of 24 db. A power change of 24 db is equal to the sum of 20 db and 4 db. According to the table, 20 db corresponds to a power change in the ratio 100 to 1, and 4 db corresponds to a power change in the ratio 2.5 to 1. As a result, 24 db corresponds to 100 · 2.5 to 1, or 250 to 1. For gains, the power-ratio factor is 250 to 1; for losses, the power-ratio factor is 0.01 · 0.4 to 1, or 0.004 to 1, or 1 to 250. For gains, the percentage-power gain is (250-1) · 100 = 24,900 percent; for losses, the percentage-power loss is (1 - 0.004) · 100 = 99.6 percent.

13-576. OTHER USES OF DECIBELS.

13-577. Up to this point, the decibel has been considered only as a measure of power losses caused by attenuation and power gains resulting from amplification. The decibel is also used in the measurement of three other types of power loss that can occur as the result of the insertion of equipment into, or across, the transmission line; these are insertion loss, mismatch or reflection loss, and bridging loss.

13-578. INSERTION LOSS. Insertion loss is a measure of the power lost because of the insertion of some device, such as a filter network, into a transmission arrangement. In part A of figure 13-136, the power transmitted directly to the receiving end without network insertion is P_r. In part B, the network has been inserted into the line and causes a reduction in the received power. The reduced power is identified as P_l. Using the ratio P_r over P_l as the power ratio (to avoid negative logarithms), insertion loss can be conveniently expressed in decibels, as follows:

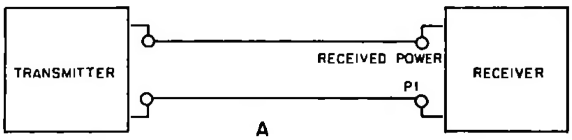
$$N_{db} = -10 \log \frac{P_r}{P_l}$$

Because it is always apparent whether there is a gain or a loss in decibels, it simplifies calculations to always place the larger power in the numerator of the power ratio and, after completing the problem, affix the proper sign to indicate a gain or a loss.

13-579. MISMATCH OR REFLECTION LOSS. Mismatch or reflection loss is the ratio of the power measured when there is a mismatch at the receiving end of the transmission line (or elsewhere on the line), to the power measured at the receiving end when the receiver is matched to the line (and there is no mismatch elsewhere on the line). In part A of figure 13-137, the receiver is matched to the line; consequently, the impedance at every point on the line is equal to the characteristic impedance of the line, and there is maximum transfer of power to the receiver. This amount of power is P_r. In part B of the figure, a cable of different characteristic impedance has been inserted into the line, so that a mismatch exists at both junctions of the cable and the line. As a result of the mismatch, a portion of the transmitted power is reflected, thus reducing the power at the receiver. This reduced power is P_l. As was done for insertion loss in the preceding discussion, if the power ratio is taken as P_r over P_l, the power loss due to reflection or mismatch, expressed in decibels, is:

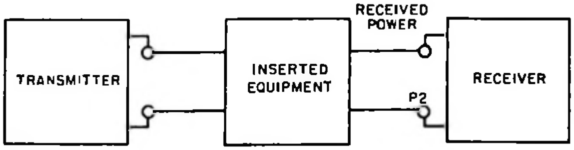
$$N_{db} = -10 \log \frac{P_r}{P_l}$$

A reflection loss can be caused by a mismatch at the junction of the transmission line and a cable or a piece of equipment



A

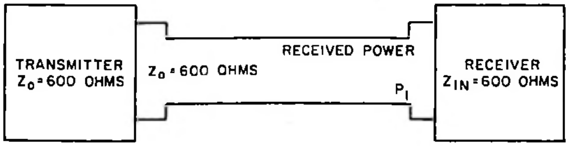
TRANSMISSION WITHOUT INSERTION LOSS



B

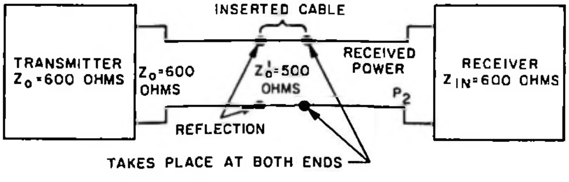
TRANSMISSION WITH INSERTION LOSS

Figure 13-136. Insertion Loss



A

MATCHED TRANSMISSION-LINE NETWORK



B

MISMATCH ON TRANSMISSION-LINE NETWORK

Figure 13-137. Mismatch or Reflection Loss inserted into the line, by a change in the properties of the line, or by a mismatch between the receiver and the line.

13-580. BRIDGING LOSS. Bridging loss is the loss in power in a transmission line between a transmitter and receiver when a test set (such as a transmission measuring set) is bridged across the line. The calculation of bridging loss involves the ratio of two powers. Let P_r represent the power in the transmission line when the measuring set is connected, and P_l the power when the set is not connected across the line. See figure 13-138. The power ratio used in the calculation of the bridging loss, then, is:

$$\frac{P_l}{P_r} = \frac{\text{power without test set}}{\text{power with test set}}$$

Using this ratio, the bridging loss in decibels is found as follows:

$$N_{db} = -10 \log \frac{P_l}{P_r}$$

It is usually desirable that the bridging loss resulting from the connection of a measuring set across a transmission line be kept to a minimum. This means that the measuring set must draw minimum power from the line.

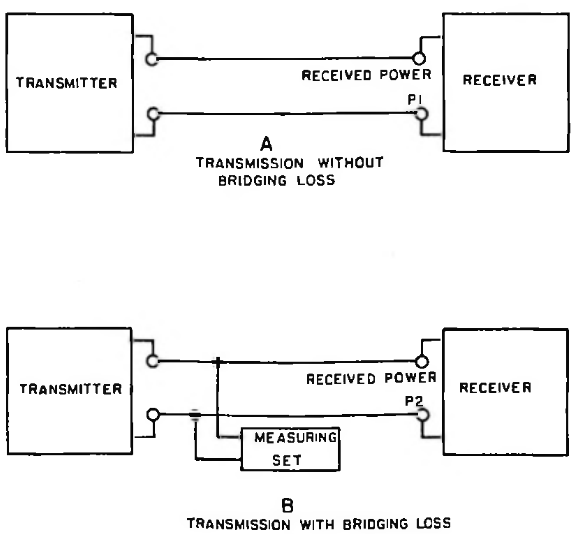


Figure 13-138. Bridging Loss

13-581. USE OF STANDARD TESTING POWER.

13-582. The decibel is fundamentally a unit of power ratio, rather than power. This results in a disadvantage in its use. For example, 10 db indicates that there has been a change of power by a factor of 10, but there is no indication of what the amount of change is. The values of the initial and final powers are not given by the decibel as a unit of attenuation or gain. It would be advantageous to be able to use decibels not only as the measure of a power ratio, but also as the measure of a specific amount of power. This is done in the following example. Consider the following equation:

$$N_{db} = 10 \log \frac{P_l}{P_r}$$

If $P_r = 1$, then:

$$N_{db} = 10 \log P_l$$

Consequently, if the input power has a value of unity, and if P_l is the output power, then N_{db} is 10 times the logarithm of the output power. When the number of decibels is given the output power can be ascertained quickly, since the number of decibels refers to the ratio of the output power to 1. On a transmission line, the power transmitted is very small, being measured in milliwatts. Therefore, the reference or input power level has been chosen arbitrarily as 1 milliwatt. Then, $N_{db} = 10$ indicates that there has been a gain of 10, and the output power is 10 milliwatts; $N_{db} = 20$ indicates that there has been a gain of 100, and the output power is 100 milliwatts. But there must be some way of indicating the difference between N_{db} used as an expression of a power ratio, and N_{db} used as an expression of an absolute amount of power. To show this difference, db is used only as the expression of a power ratio, and dbm is used as the measure of absolute power as compared with an arbitrary reference level of 1 milliwatt; 1 milliwatt of power is equal to 0 dbm. $N_{db} = 10$, therefore, represents a change of power by a factor 10, and $N_{dbm} = 10$ represents 10 milliwatts of power.

13-583. In low-frequency transmission, it has been found that the elimination of all frequencies below 200 cycles and above 2700 cycles per second causes very little loss in the intelligibility of the received signal. Most voice-frequency equipment is designed to transmit effectively this 200- to 2700-cycle band, known as the voice-frequency range. A 1000-cycle sine-wave generator is commonly used as a standard for checking and calibrating all types of equipment. This calibrating frequency is near the middle of the voice-frequency band. In order to measure power loss or gain in a speech circuit, a 1000-cycle sine-wave tone with 1 milliwatt of power is used as a standard testing signal.

13-584. It must be emphasized that the measurement of power in dbm is independent of a standard frequency. Although the testing signal for voice-frequency circuits is a 1000 cycle tone, the dbm can be used for measuring power at any frequency in any type of circuit as long as the reference power is understood to be 1 milliwatt (0 dbm).

13-585. ZERO-LEVEL POINT.

13-586. The circuit in part A of figure 13-139 consists of a transmitter and a receiver, each connected to the end of a transmission

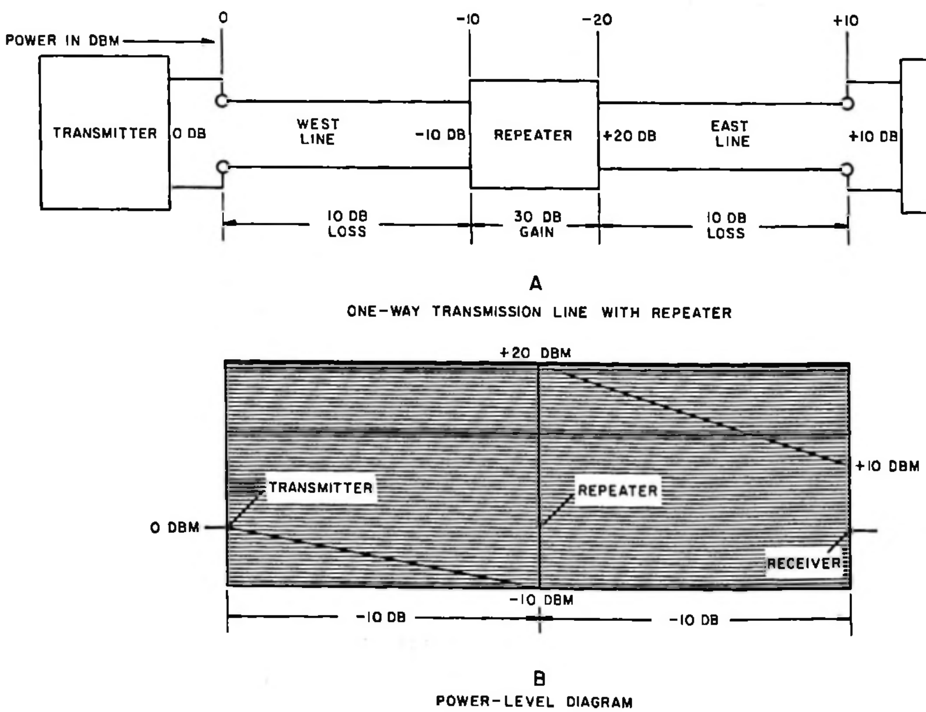


Figure 13-139. One-Way Transmission Line with Power-Level Diagram

line. A repeater is used in the line to make up for the power loss caused by attenuation. When a 1000-cycle test tone with 1 milliwatt of power is used to measure the power loss or gain in the speech circuit, the point at which the test-tone signal is applied is the reference point. In this case the transmitter shown in part A of the figure is used as the reference point; consequently, there is 0-dbm absolute power. This point is referred to as the 0-level point.

13-587. The west section of the transmission line shown in part A of the figure causes a 10-db power loss, the repeater causes a 30-db power gain, and the east section of the transmission line causes a 10-db power loss. These power changes are expressed in db (not dbm) when they are with respect to themselves—that is, along the line between the transmitter and receiver; the test tone powers are expressed in dbm (not db) when they are with respect to the fixed 1-milliwatt level. The total power change in db (the algebraic sum of all the power changes in db) is +10 db. This total power change is with respect to the 1-milliwatt level at the transmitter, so that at the input of the receiver the test tone power is +10 dbm above the zero-level point at the transmitter.

13-588. To summarize the discussion of the zero level point the following factors should be remembered: Zero dbm indicates a power level of 1 milliwatt. A positive number of dbm indicates a power level of greater than 1 milliwatt. A negative number of dbm indicates a power level of less than 1 milliwatt. It is desirable that in all communication equipment and facilities the zero-level point be the same; in this case it is taken to be the transmitter.

13-589. POWER-LEVEL DIAGRAMS.

13-590. The mathematical results obtained in earlier paragraphs can be represented as a graph called a power-level diagram (also called energy-level diagram). Such a diagram graphically shows the absolute powers throughout a circuit, and is used in both the design and maintenance of long lines incorporating repeaters.

13-591. Part B of figure 13-139 represents the power-level diagram for the one-way transmission arrangements shown in part A of the figure. The horizontal base line indicates a level of 0 dbm. Each horizontal line above and below the base line represents a value of 1 dbm. Positive values are located above the base line, and negative values below it. At the transmitter there is a 0-dbm level. Over the distance between the transmitter and the repeater, the power decreases 10 db below the 0-dbm level. The repeater increases the power 30 db. This increase is represented by a vertical line because it does not take place over a distance, as does the power loss between transmitter and the repeater. Between the repeater and the receiver, there is again a decrease of 10 db in the power level. A line representing this loss is drawn downward 10 horizontal lines from the +20-dbm level at the output of the repeater to the +10-dbm level at the input of the receiver. The final result indicates that between transmitter and receiver the test signal power has increased to +10 dbm, so that the test signal input to the receiver is 10 milliwatts.

13-592. A two-way transmission line is shown in part A of figure 13-140. The four repeaters in the line pass the signal in both directions, as indicated by the arrows. The values above and below the arrows show the gains of the repeaters in both directions.

13-593. Part B shows the power levels in transmission from west to east. At transmitter A, the power is at 0 level (1 milliwatt). Power gains caused by the four repeaters are represented by vertical lines proportional in length to the gains. Thus, at repeater No. 2 the input power is —10

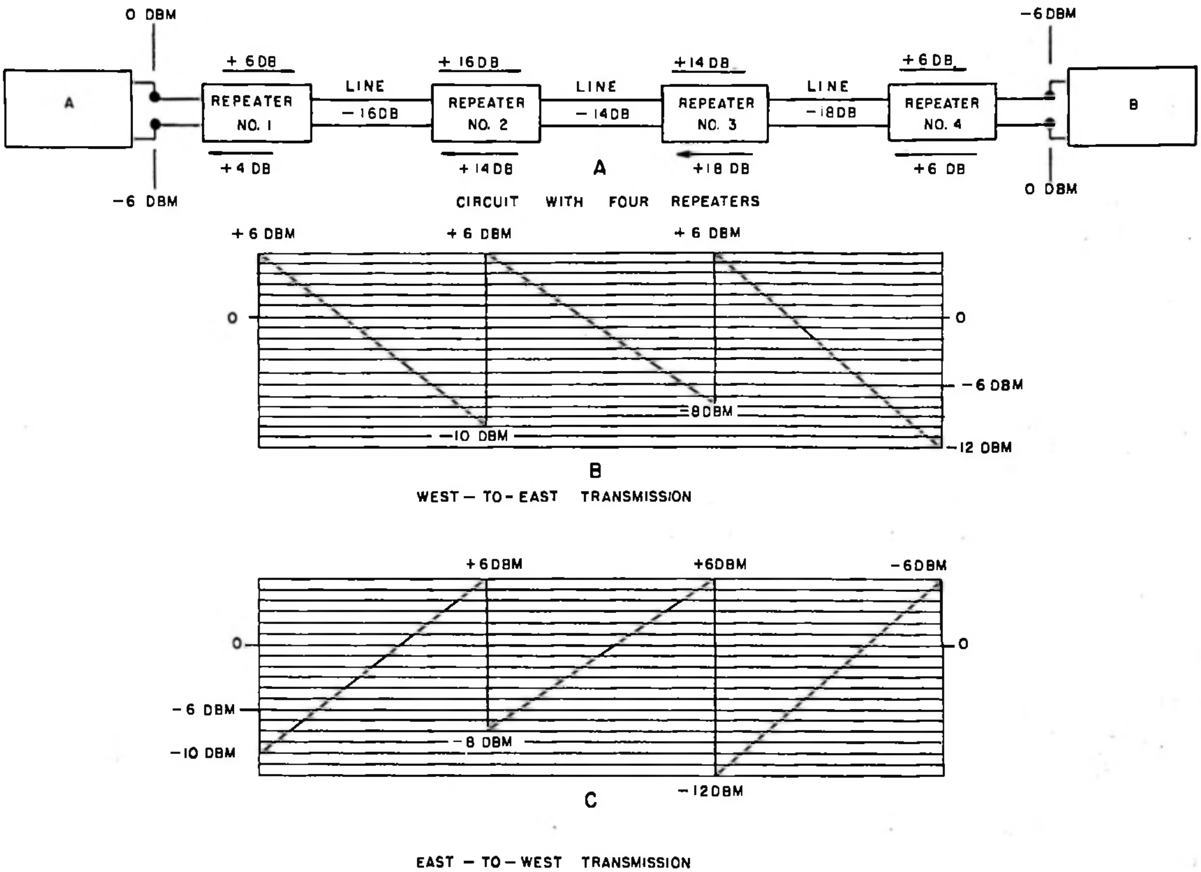


Figure 13-140. Two-Way Transmission Line with Power-Level Diagram

dbm (10 db down from the 0-level point). The line indicating the lower gain caused by the repeater goes from —10 dbm up through 16 horizontal lines to a power of +6 dbm, because the gain of the repeater is +16 db. Power losses resulting from attenuation along the transmission lines are represented by lines drawn diagonally downward. The vertical components of the lines indicate a lowering of the power; the horizontal component indicates that the power losses take place over the distances between the repeaters. Thus, the power loss between repeaters No. 2 and 3 is represented by a line extending from the +6-dbm level at the output of the second repeater downward through 14 horizontal lines to a power of —8 dbm, because the transmission line has caused a power loss of 14 db. At the input to repeater No. 3 the power is —8 dbm.

13-594. Part C of figure 13-140 shows the power levels in transmission from east to west. The diagram is drawn in the same manner as in part B; since transmitter/receiver B is now the transmitting end, it is the 0-level point, and the diagram is read from right to left.

13-595. The transmitter/receivers represented in parts B and C use a test tone input power of 1 milliwatt. In voice transmission, the actual power will vary. A diagram showing the speech power range in dbm of transmitted signals at various points is shown in figure 13-141. For example, note that the talker input can vary from +6 to —25 dbm. It is standard practice to assume that the noise signal power should be at least 20 db below the weakest speech signal.

13-596. In lower frequency communications, there is frequently erroneous usage of the two terms commonly known as level and power. This difficulty can be cleared up when db or dbm is specified. The present tendency is to use the word power to mean absolute power in dbm. The term db remains the ratio of the two powers. A new term, relative transmission level, is coming into use. Relative transmission level in db is defined as a ratio of sine wave test tone power measured at any point in the equipment to the power at some other point in the equipment chosen as the reference point. Figure 13-142 is a relative transmission diagram. Part A represents the power losses and gains on a line from Atlanta to Louisville, and part B represents the same line from Louisville to Atlanta. The diagram enables the operators of the repeater stations to know at a glance the proper output power and repeater gains for transmission in either direction, in db. It should be noted in this relative transmission level diagram that, when the input power at the 0-level reference point is 1 milliwatt, all the values can also be expressed in dbm. That is, any value of db equals the same value in dbm.

13-597. RF TRANSMISSION LINE MEASUREMENTS.

13-598. CHARACTERISTIC IMPEDANCE AND WAVELENGTH.

13-599. At radio frequencies the characteristic impedance, Z_O, of a well-designed transmission line may be considered as

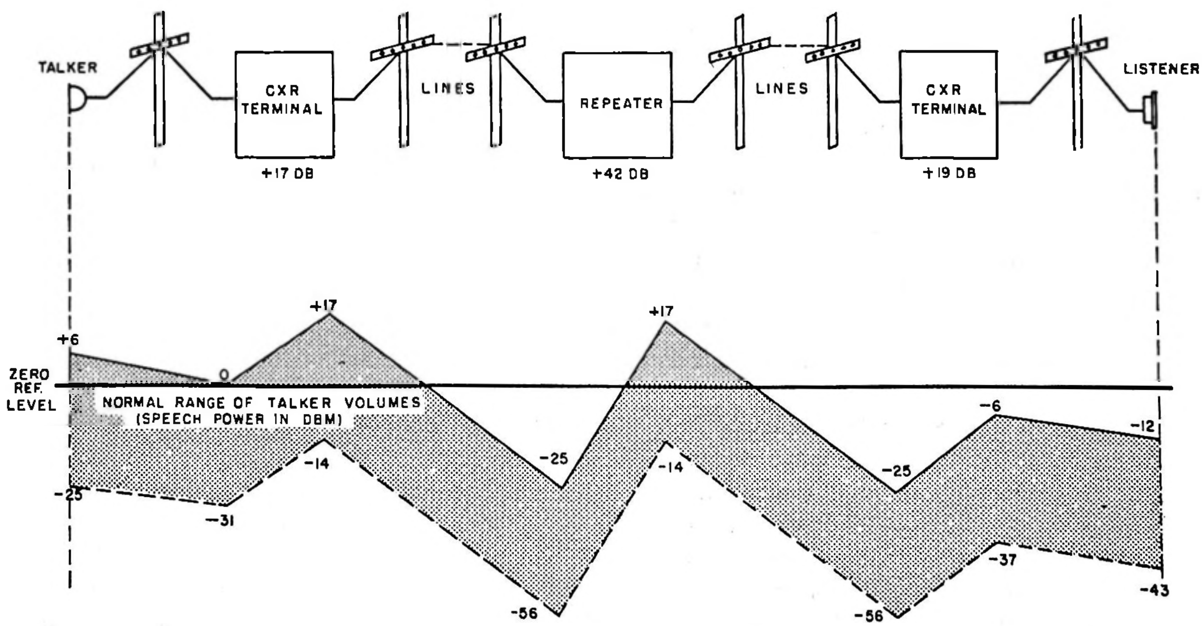


Figure 13-141. Normal Range of Talker Volumes

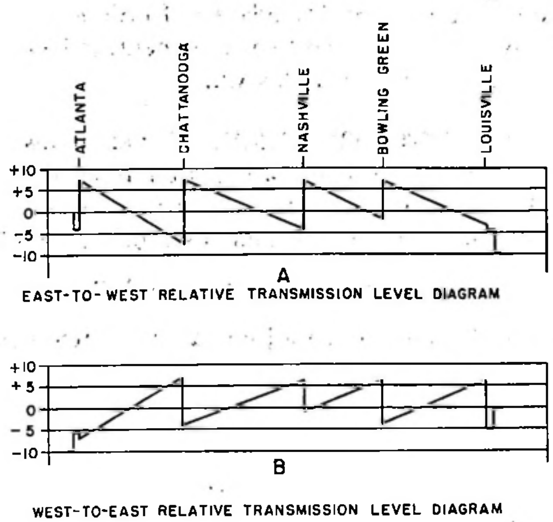


Figure 13-142. Relative Transmission Level

approximately equal to the square root of L/C. A two-wire line having almost any desired characteristic impedance can be constructed for practical applications.

13-600. The approximate values obtained when using the formula $Z_o = \sqrt{L/C}$ indicates that the characteristic impedance of an rf line depends principally on the values of L and C in the line. An increase in the separation of the wires increases the inductance, L, and decreases the capacitance, C. This effect takes place because the effective inductance is very low if two wires are closely spaced and carrying currents in opposite directions, and because the capacitance is low if the wires (plates of a capacitor) are widely separated. The effect of increasing the spacing is to increase the characteristic impedance, since increasing L and reducing C increases the L/C ratio. A reduction in the diameter of the wires also increases the characteristic impedance. This reduction affects the capacitance more than the inductance, for it is equivalent to decreasing the size of the plate in a capacitor in order to obtain lower capacitance. Any change in the dielectric material between the two wires also changes the characteristic impedance. Thus, if a change in dielectric material increases the capacitance between the wires, the characteristic impedance is reduced.

13-601. The approximate characteristic impedance of a two-wire line with air as a dielectric may be obtained from the following formula:

$$Z_o = 276 \log_{10} b/a$$

where b is the spacing between the centers of the conductors, and a is the radius of each conductor. This formula is sufficiently accurate at high frequencies, where the characteristic impedance is practically a pure resistance.

13-602. Example: If two wires 1/4 inch in diameter are spaced 2 inches apart, what is the characteristic impedance?

$$Z_o = 276 \log_{10} b/a$$

$$b = 2 \text{ inches}$$

$$a = 1/2 \cdot 1/4 = 1/8 = .125 \text{ inch}$$

$$Z_o = 276 \log_{10} 2/.125 = 276 \log_{10} 16$$

From a table of logarithms:

$$\log_{10} 16 = 1.204$$

Therefore:

$$Z_o = 276 \cdot 1.204 = 332.3 \text{ ohms}$$

13-603. CONCENTRIC LINE.

13-604. The characteristic impedance of a concentric line also varies with L and C; however, because of the difference in construction, L and C vary in a slightly different manner. The following formula must be used to calculate the characteristic impedance of a concentric line:

$$Z_0 = 138 \log_{10} b/a$$

where b is the inner diameter of the outer conductor and a is the outer diameter of the inner conductor.

13-605. IMPEDANCE GRAPHS.

13-606. Two convenient graphs (figure 13-143) make it possible to determine quickly the characteristic impedance of either a two-wire parallel line or a concentric line when the ratio b/a is known.

13-607. TWO-WIRE LINE. For example, a two-wire line made of copper tubing of 1/4-inch diameter and with a center-to-center spacing of 3 inches has a b/a ratio of $\frac{3}{1/8} = 24$. In part A of figure 13-143, the value of 24 on the baseline can be carried up

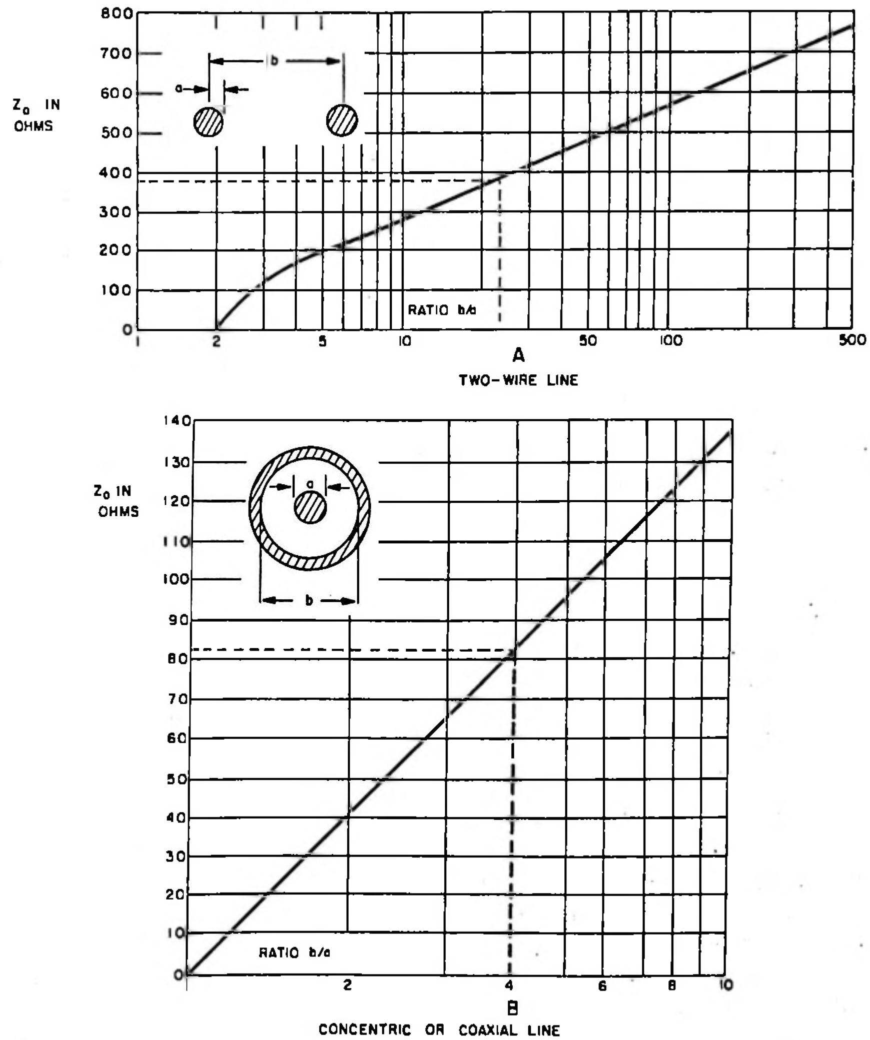


Figure 13-143. Characteristic Impedance Graphs

the slant line and thence over to the left to the vertical scale (see dotted lines). A value of approximately 375 ohms is obtained. A more exact value of 381 ohms can be calculated from the equation $Z_o = 276 \log_{10} b/a$.

13-608. CONCENTRIC LINE. A concentric line having an outside inner-conductor diameter of 1/16 inch and an inside outer-conductor diameter of 1/4 inch has a b/a ratio of $\frac{1/4}{1/16} = 4$. In part B of figure 13-143, a value of approximately 82 ohms can be calculated.

13-609. ELIMINATION OF STANDING WAVES.

13-610. The difficulties of making high-frequency measurements are so great that it is usually more practicable to calculate the characteristic impedance, Z_o, by means of the approximation formulas. The characteristic impedance of a line may be measured simply, and with sufficient accuracy for practical purposes, by the following method. Terminate the line with a non-inductive, calibrated variable resistor. Adjust the resistance so that the standing waves are eliminated or reduced to the minimum possible magnitude. A suitable rf indicator must be used to measure the standing waves. The resistance read on the calibrated scale, which reduces the standing waves to zero or to the minimum possible magnitude, is approximately equal to the characteristic impedance of the line. In most practical work it is not usually necessary to determine the characteristic impedance, since the only requirement is to eliminate the standing waves. To do this, vary the load at the output end of the line until the standing waves are reduced to the minimum possible magnitude.

13-611. WAVELENGTH MEASUREMENTS.

13-612. Reflections always begin at the output end of a line. The distance from the output end to any maximum (or minimum) point is therefore a function of frequency and line termination. It is convenient to measure these distances in terms of wavelength. The distance between successive maximum (or minimum) points of voltage (or current) is equal to one-half the wavelength (see figure 13-144). One wavelength is twice the distance between successive maximum (or minimum) points. The distance corresponding to one wavelength of a wave in free space is given by the following equation:

$$\lambda = V/f = 300,000,000/f$$

where λ is the wavelength in meters, V is the velocity in meters, and f is the frequency in cycles.

13-613. The velocity, V, of an electromagnetic wave varies with the medium through which the wave is passing. In mediums other than free space, the velocity is always less than the velocity in free space (300,000,000 meters/second). Referring to the previously used equation for wavelength, you can see that if the velocity is reduced, the length of the wave must also be reduced. This means that in any real transmission line, the velocity is always less than 300,000,000 meters/second, and, therefore, a wavelength on the line is shorter than the corresponding wavelength in free space. The electrical quarter-wavelength for various types of rf line may be calculated from the following equation:

$$\frac{\lambda}{4} = 246\ k/f$$

where $\lambda/4$ is the quarter-wavelength (measured in feet), f is the frequency in megacycles, and k is a multiplying factor. The factor k expresses the ratio of the actual velocity of the waves on a particular line to the velocity of the same waves in free space.

13-614. LECHER LINES.

13-615. Lecher line is a term applied to a length of parallel two-wire transmission

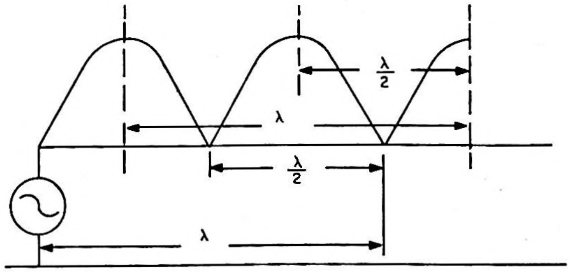


Figure 13-144. Standing Waves as a Measure of Wavelength

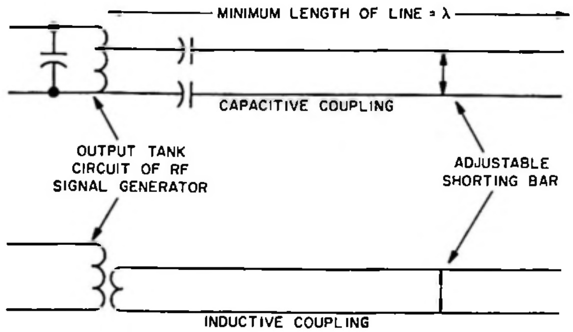


Figure 13-145. Methods of Coupling Lecher Lines

line which is used as a tuned-circuit element, or a resonant line, to measure wavelength. A Lecher line is usually from one-quarter to five wavelengths, and has a shorting bar (or bridge) which can be placed across the line and moved to any point along its length. In figure 13-145 the generator and the Lecher line are capacitively coupled through capacitors of very low capacitance (high reactance), so that the line will be electrically open at its sending end. To provide practically short-circuit conditions across the input terminals, the line is inductively coupled to the generator. The inductance should consist of no more than a single turn of wire. At higher frequencies a straight shorting bar across the input terminals will provide sufficient coupling to the generator and, at the same time, establish essentially short-circuit conditions across the input terminals of the line.

13-616. Consider the capacitively coupled Lecher line shown in figure 13-146. If the shorting bar is placed at the quarter-wave point, as shown in part A of figure 13-146, there will be a reflection of current from the shorting bar, and the standing waves of voltage and current will be as shown. An ammeter, A, in the shorting bar indicates a high current. A plot of current readings in the wires at successive points from the bar toward the generator gives the current curve, I. A similar plot of voltage readings gives the voltage curve, E. You know that the impedance at any point in the line is given by the equation $Z = E/I$. At the input terminals,

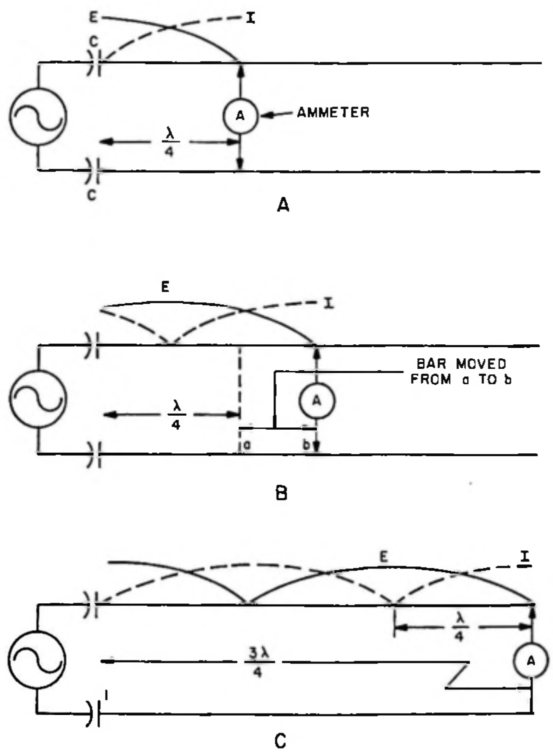


Figure 13-146. Variation of Standing Wave as Shorting Bar is Moved

points CC in part A of figure 13-146, I is zero and E is equal to the generator voltage; therefore, assuming that the resistance of the line is zero, the input impedance is:

$$Z = E/I = E/O = \text{infinity}$$

This indicates that the shorted quarter-wave line acts as a parallel-resonant circuit. In a real line, of course, there is a certain amount of resistance; thus the actual input impedance is less than infinity. However, if the line resistance is reduced to the minimum possible value, the input impedance will be very high.

13-617. If the shorting bar is moved from the quarter-wave position, there will be little change in the current through it. However, there will be a change in the standing waves, as shown in part B of figure 13-146, and the line will no longer act as a parallel-resonant circuit. Increased current from the generator into the line indicates that the input impedance is lower.

13-618. If the shorting bar is placed at the three-quarter-wave point, as shown in part C of figure 13-146, the voltage and current at the input terminals will be identical to the voltage and current at the input terminals of the quarter-wave line, shown in part A of the figure. This assumes that there is no resistance in the line. With resistance in the line the current into the 3 $\lambda/4$ line will be somewhat higher than the current into the $\lambda/4$ line. This shows that the maximum possible input impedance is obtainable with a $\lambda/4$ line shorted at the receiving end. Lecher lines may be used to measure wavelengths by detecting the positions of the voltage (or current) maximum (or minimum) points along the line.

13-619. Current values can be measured roughly by the use of a small loop or coil of wire coupled inductively to the line. A meter across the loop will read a maximum value when the loop is adjacent to a current maximum point. One wavelength is twice the distance between successive maximum (or minimum) points on the line. Parts A and B of figure 13-147 show simple rf current indicators which may be used to determine the maximum and minimum points. In part B, the diode rectifies rf currents to provide direct current for the dc meter. You should note here that a point of maximum current corresponds to a point of minimum voltage, and a point of minimum current corresponds to a point of maximum voltage.

13-620. It may be more convenient to use an rf voltage indicator to locate the points of maximum and minimum voltage. Parts C and D of figure 13-147 show two simple types of rf voltage indicators. The indicator shown in part C of the figure consists of a neon bulb and a resistor connected directly across the line. In part D the neon bulb is replaced

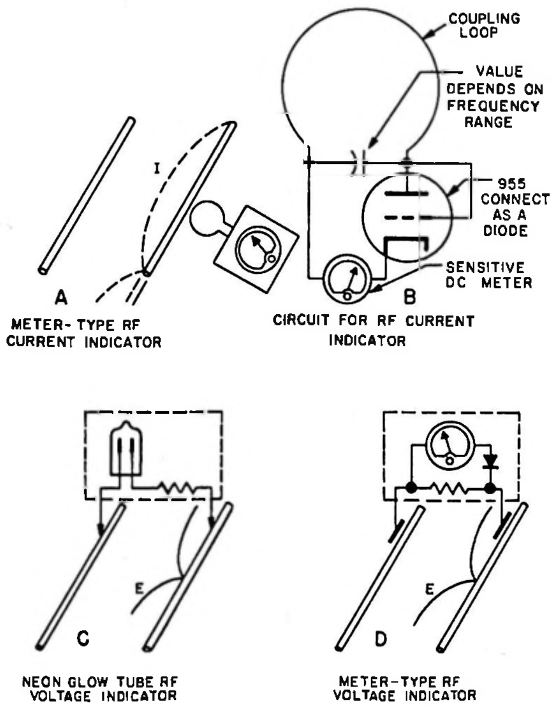


Figure 13-147. Simple RF Indicators

by a meter and a crystal or tube-type rectifier. The device is capacitively coupled to the line. The following method may be used to obtain very rough indications of the standing waves along a line. Hold an ordinary tubular-type, fluorescent light bulb by one end and bring the other end close to the line. If the line has standing waves, the bulb will glow with varying brightness as it is moved along the line. Points of maximum brightness are points of maximum voltage on the line. For correct results, the bulb must be held at a uniform distance from the line.

13-621. The devices discussed in the previous paragraphs are simple devices and may be constructed easily. For more accurate work, well-designed indicators are available. To obtain information on the theory and operation of any of these indicators, refer to the instruction book for the particular item.

SECTION X

WAVEGUIDES

13-622. GENERAL.

13-623. A waveguide is the equivalent of a coaxial cable with the central conductor and supporting insulation spacing removed so that all that remains is a hollow pipe. In practice, waveguides are hollow pipes, normally rectangular or cylindrical, and fabricated out of material having good electrical conductivity.

13-624. Waveguides become increasingly necessary at higher and higher frequencies; they are indispensable for the transmission and propagation of radio-frequency energy when the wavelength dimension is reduced to inches or even less, as in the case of super high frequencies. This is particularly true when the distance between the central conductor and the inner sheath of a coaxial cable, for example, approaches the dimensions of a quarter-wavelength. At that limit, a coaxial cable becomes either impossible or unpredictable in actual practice. The radio-frequency energy under such a condition takes a complex or unpredictable path in addition to the longitudinal path provided by the central conductor. It may, for example, try to take lateral or semi-lateral paths as if the central conductor were the earth and the conducting inner sheath wall of the coaxial cable were an ionospheric reflecting layer. While it is conceivable that a set of conditions involving extremely careful dimensions and uniformity of the cable can permit the use of coaxial cable at super high frequencies, it is much more feasible and efficient to use simple waveguide pipe.

13-625. WAVEGUIDE THEORY.

13-626. An exact mathematical analysis of the way in which fields exist in a waveguide is beyond the scope of this book. However, it is possible to obtain an understanding of many of the properties of waveguide propagation by using the following simple analogy, which shows how the fields are able to exist in a waveguide and how you can handle them.

13-627. To understand the action of a waveguide, visualize the waveguide as having the form of a two-wire line. In this condition there must be some means of supporting the two wires. Furthermore, the support must be a nonconductor, so that no power will be lost by radiation leakage. An efficient method of both insulating and supporting the two-wire is shown in part A of figure 13-148. This line is spaced, insulated, and supported by porcelain stand-off insulators. At communication frequencies, the absorption of power by the dielectric material (insulation) causes the insulators to act as a combined low resistance and capacitance.

13-628. The equivalent electrical circuit at higher frequencies is shown in part B of figure 13-148. For frequencies of 3000 mc and higher, you must use a better insulator than nonconducting porcelain. A superior high-frequency insulator for this purpose is a quarter-wave section of rf line, called a metallic insulator. Such an insulator is shown in part C of figure 13-148. Since there are no dielectric losses in a quarter-wave section of rf line, the impedance at

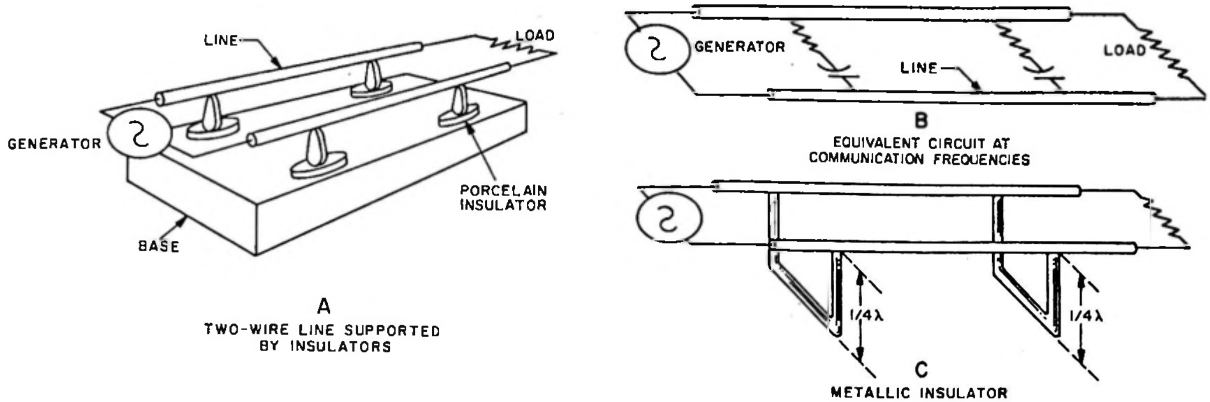


Figure 13-148. Insulating the Two-Wire Line

the open end (the junction of the two-wire line) is very high.

13-629. A metallic insulator can be placed anywhere along a two-wire line; part A of figure 13-149 shows several on each side of the line. A point which you should note in this type of line is that the supports are a quarter-wavelength at only one frequency. This limits the high efficiency of the two-wire line to one frequency only.

13-630. The use of several insulators results in improved conductivity of a two-wire line when the insulator sections are connected together. Part B of figure 13-149 shows a switch connection between two adjacent sections. When the switch is open, both quarter-wave sections are excited by the main line. In this condition, there will be standing waves on the quarter-wave sections.

13-631. When the switch is connected to the same place on each section, the relative phase relationship of the voltages at the connections will be the same for each section. In this condition the No. 1 section will be excited first by the generator. When the switch is closed, the No. 2 section will be partly excited by the No. 1 section through

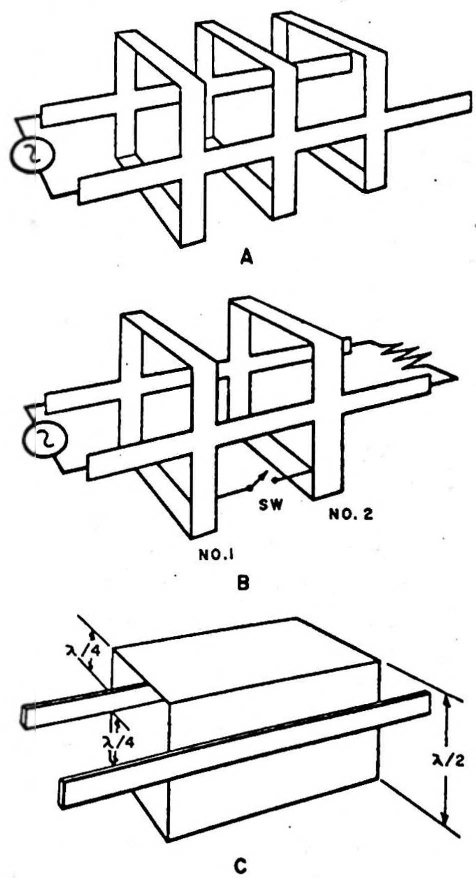


Figure 13-149. Development of Waveguide by Adding Quarter-Wave Sections

the switch connection. The parallel paths shown cause less resistance to exist along a given length of line, and energy is transferred with less dc resistance loss.

13-632. When more and more quarter-wave sections are added to the line until each section makes contact with the next, the result is a rectangular box in which the line is at the center, as shown in part C of figure 13-149. The line itself is actually part of the wall of the box. The rectangular box thus formed is a waveguide.

13-633. TYPES OF WAVEGUIDES.

13-634. Although any transmission arrangement serves as a guiding structure for electromagnetic waves, the term waveguide is reserved for the type that consists of a single hollow conductor. Although a waveguide may have almost any shape, the two types in common use are those having rectangular and circular cross sections. The characteristics of the rectangular waveguide are determined by the height and the width. These dimensions are always the inside measurements, the smaller of the two customarily being taken as the height, even though it may lie in a horizontal plane. The characteristics of the circular waveguide are determined by the diameter, which is always the inside measurement.

13-635. TRANSMISSION IN A RECTANGULAR WAVEGUIDE.

13-636. Although a rectangular waveguide can carry energy in many ways, only one way, the simplest, is usual in practice. In this case, vertically polarized waves travel down the waveguide along paths that are parallel to the top and bottom of the guide.

13-637. TERMINOLOGY OF RECTANGULAR WAVEGUIDE.

13-638. As discussed in the previous paragraphs, the electrical characteristics of a rectangular waveguide are determined by the inside dimensions, as shown in figure 13-150. Whatever the orientation of the waveguide, it is customary to call the shorter dimension the height, and the longer one the width. Thus, the longer walls of the waveguide are the top and bottom, and the shorter walls are the sides.

13-639. Vertical polarization as applied to an electromagnetic wave in a rectangular waveguide, therefore, means that the electric field is parallel to the shorter dimension of the guide. The small letters a and b are customarily used to denote the height and width, respectively, when referring to rectangular waveguide dimensions.

13-640. PATHS OF WAVES.

13-641. The paths followed by the electromagnetic waves are shown in figure 13-151. Part A is a side view, with the wave paths parallel to the top and bottom walls of the waveguide. Part B is a top view of the paths. All of the paths make the same angle, B, with the sides of the waveguide.

13-642. DETERMINATION OF ANGLE B.

13-643. Angle B cannot be chosen arbitrarily. Instead, it depends on b, the width of

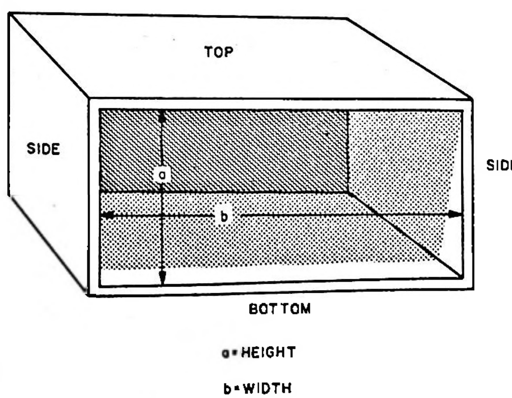


Figure 13-150. Rectangular Waveguide Dimensions

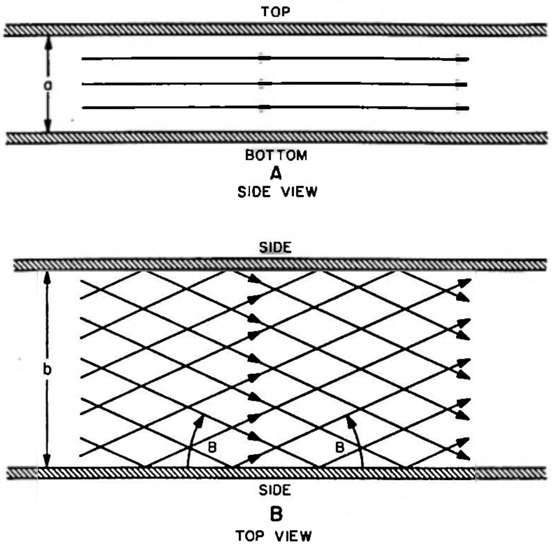


Figure 13-151. Paths of Waves in a Rectangular Waveguide

the waveguide, and on λ, the wavelength of the signals. The relationship is given by the following equation:

$$\sin B = \lambda/2b$$

This equation results from the fact that propagation is impossible unless all signals traveling in a given direction reinforce one another. The equation is derived easily from consideration of the diagram shown in figure 13-152. Consider a signal that passes through point C moving to the left. It is reflected from the left-hand wall at point D, and from the right-hand wall at point E; it finally moves in direction EF, which is parallel to the original direction, CD. A wave front is always perpendicular to the path of a wave, and since DH is perpendicular to CD, DH is a wave front for signals moving to the left. The phase of a signal is the same at all points on a wave front. Therefore, a signal moving to the left through point H is in phase with the incident signal at point D. If energy is to be propagated down the waveguide, the reflected signal at point E must have the same phase as the signal moving to the left through point I, since both are on a wave front perpendicular to point EF. Otherwise, the signals will cancel each other after traveling a short distance through the waveguide. Since the electric field of the signals is parallel to the reflecting wall, there is a phase reversal at point D and another one at point E.

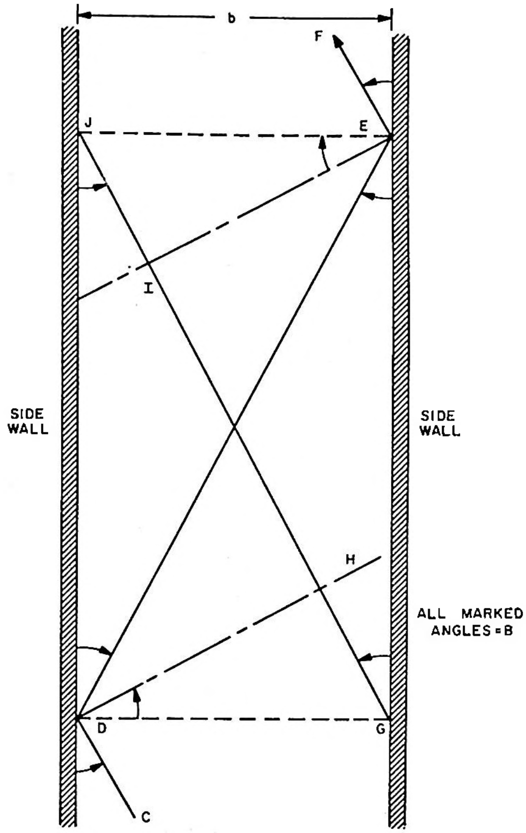


Figure 13-152. Geometry of Equation for Determination of Angle B

13-644. The signal behavior on reflection is the same as that of a horizontally polarized signal reflected from the surface of the sea. However, it does not matter whether this is the case or not, since two phase reversals have a net effect of zero. To have the desired phase relations, angle B must have a value which makes path DE an integral (whole) number of wavelengths longer than path HI. Path GJ is equal to path DE, since G lies opposite D, and J lies opposite E. The requirement for the propagation of energy is that HI and GJ differ by an integral number of wavelengths. The difference between GJ and HI is the sum of GH and IJ, and each of these distances is equal to b sin B. The condition for propagation, then, is as follows:

$$2b \sin B = n\lambda$$

where n is an integer (1, 2, 3, etc). In the simplest case, which is the one under discussion, n is equal to 1; thus, when 1 is substituted for n, the equation given above immediately reduces to the following:

$$\sin B = \lambda/2b$$

It must be emphasized that the waveguide will carry energy only for certain values of angle B, because the signals cancel for other values of the angle. This has nothing to do with attenuation or losses; the phenomenon is similar to the directive effect of an antenna that does not radiate in directions for which signals from various portions of the antenna cancel each other.

13-645. FIELD PATTERNS IN A RECTANGULAR WAVEGUIDE.

13-646. The patterns of the electric and magnetic fields in the rectangular waveguide may be found by adding the fields belonging to the two sets of electromagnetic waves—those moving to the left and those moving to the right. This may be done separately for the electric field and the magnetic field.

13-647. SPACING OF WAVE FRONTS.

13-648. To construct the field patterns, it is necessary to know the spacing in the waveguide between two wave fronts having a phase difference of one half-cycle. This is developed in the diagrams shown in figure 13-153. Here CD is a line drawn across the waveguide, and CE is a wave front of the signal that is moving to the left. The discussion in paragraphs 13-642 has shown that DF (which corresponds to GH in figure 13-152) must be one half-wavelength in the simple case (for which n is 1). Wave front DG in figure 13-152 is therefore one half-cycle different in phase from wave front CE in figure 13-152. Part A of figure 13-152 shows a fairly wide waveguide, and part B shows a narrow waveguide.

13-649. PATTERN OF ELECTRIC FIELDS.

13-650. Figure 13-154 illustrates the way in which the electric fields combine. Part A shows the electric fields of the waves moving to the left, and part B shows those of the waves moving to the right. In each illustration, the direction of the electric field is indicated by the notes beside the figure, and the strength of the field is shown by the darkness of the shadowing. Part C shows the

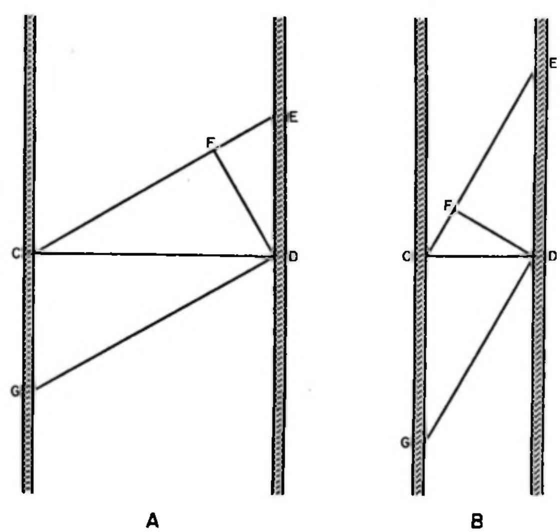


Figure 13-153. Spacing of Wave Fronts

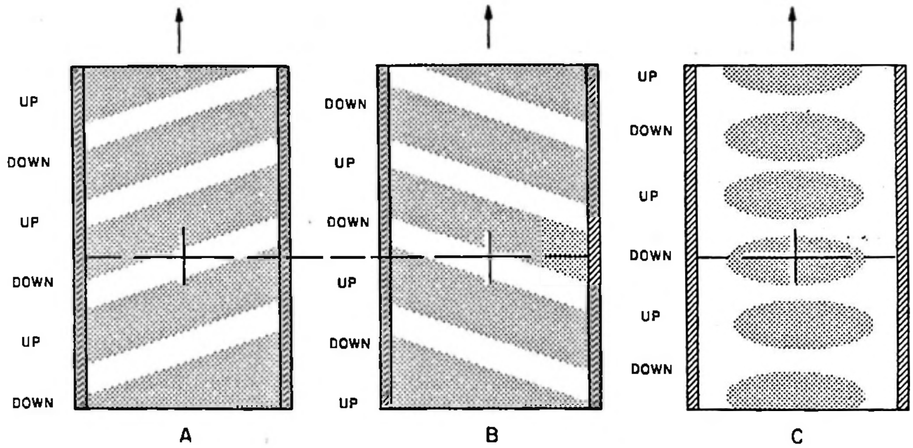


Figure 13-154. Pattern of Electric Fields

combined fields that are obtained by simply adding the fields of the two sets of waves.

13-651. PATTERN OF MAGNETIC FIELDS.

13-652. Figure 13-155 illustrates the way in which the magnetic fields combine. When adding these fields, it is necessary to consider the relative directions of the individual fields, in the same way that, when adding two ac voltages, it is necessary to consider the relative phase. The direction of the magnetic field is shown, in parts A and B, by small arrows next to the waveguide, and in part C, by arrows superimposed on the shading. The strength of the fields is indicated by the darkness of the shading.

13-653. COMBINED PATTERN OF ELECTRIC AND MAGNETIC FIELDS.

13-654. The combined pattern of the electric and magnetic fields is shown in figure 13-156. Here, the strength of the field is indicated by the darkness of the shading, and the direction is given as a note beside the waveguide. The character of the magnetic field is indicated by the closed loops. The arrowheads on these loops show the direction of the magnetic field, and the spacing of the loops indicates the strength of the magnetic field (closely spaced loops indicate a strong field). The magnetic field pattern is shown more precisely in part C of figure 13-155. At both side walls, the electric field vanishes and the magnetic field is longitudinal (parallel to the axis of the waveguide). Midway between the side walls, the field is the same as in unobstructed space, both electric and magnetic fields being transverse (perpendicular to the waveguide axis). At intermediate points, fields of both sorts are present with reduced strength, the longitudinal and transverse magnetic fields adding to produce a combined magnetic field that is neither transverse nor longitudinal.

13-655. DETAILS OF COMBINED PATTERN.

13-656. At any point inside the waveguide, sinusoidally oscillating fields will be observed as the field pattern travels down the guide past the point of observation. The character of the observed fields depends only on the distance between the point of observation and the side wall of the waveguide. At a point midway between the side walls of the waveguide, the electric field, part A, and the transverse magnetic field, part B, vary as shown in figure 13-157. The two

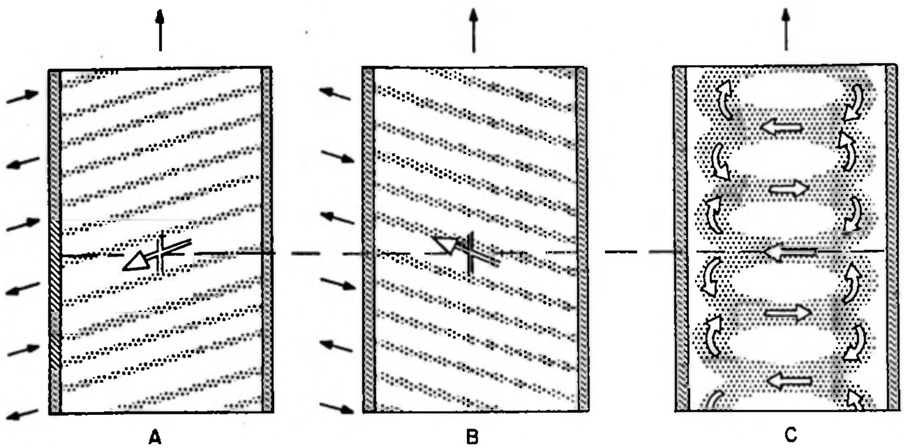


Figure 13-155. Pattern of Magnetic Fields

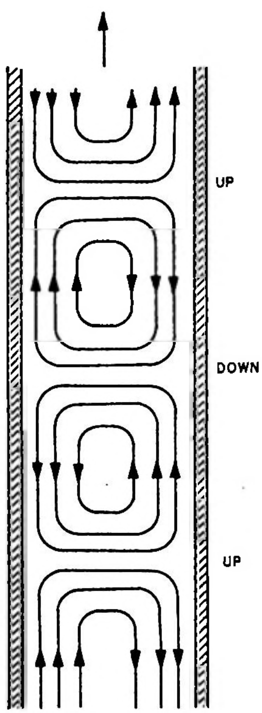


Figure 13-156. Combined Pattern of Electric and Magnetic Fields

fields are in phase with each other, the peak intensity of the electric field occurring at the same instant as the peak intensity of the transverse magnetic field. No longitudinal magnetic field is observed at a point midway between the side walls. At a point somewhere between the center of the guide and one of the side walls, the fields vary, as in parts A and B of figure 13-158. This figure shows the fields for a point between the center of the guide and the right-hand side wall. For a point between the center of the guide and the left-hand side wall, the observed fields are the same as those in figure 13-158 except that the phase of the longitudinal magnetic field, shown in part C, is reversed. At both side walls the longitudinal magnetic field is present, but both the transverse magnetic field and the electric field vanish. If the amplitude of the oscillating field is plotted against distance from the side wall of the waveguide, the resulting curve is half of a sine wave (see figure 13-159). For this reason, the pattern is sometimes called a <u>half-sine pattern</u>. The electric field and the transverse mag-

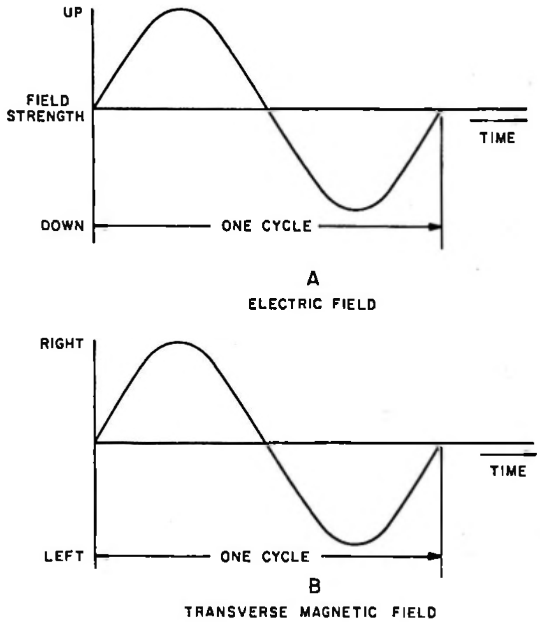


Figure 13-157. Fields Observed Midway Between Side Walls

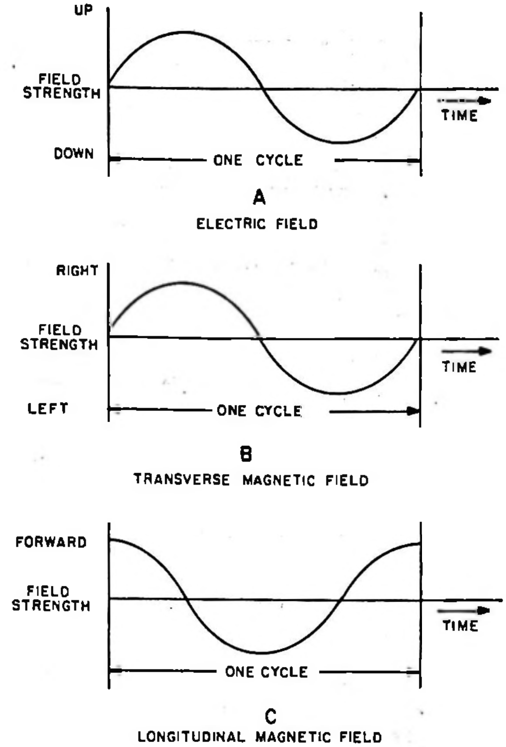


Figure 13-158. Fields Observed at Point Between Center of Guide and Right Hand Wall

netic field vanish at both walls, as in part A. The longitudinal magnetic field vanishes at the center of the guide, as in part B, and it has opposite phases in the right and left halves of the guide, as shown by the negative value of the amplitude in the figure.

13-657. CUTOFF WAVELENGTH.

13-658. As explained in previous paragraphs, for the simplest wave pattern that will carry energy down the waveguide, angle B is determined by the following relationship:

$$\sin B = \lambda/2b$$

This equation shows that B must equal 90 degrees when the wavelength is equal to 2b. Here the two sets of waves merely travel back and forth across the waveguide, and no energy is carried in the desired direction (along the guide). If the wavelength is greater than 2b, there is no value of B that satisfies the relation. The wavelength at which B equals 90 degrees is called the cutoff wavelength (λ_c), and the corresponding frequency is called the cutoff frequency (f_c). For the wave pattern described above, λ_c is equal to 2b. To avoid confusion, the wavelength of the signal is referred to as the wavelength in air (λ_a). Assuming a signal with a frequency of 3000 mc per second, for example, λ_a is equal to 10 centimeters. Waveguide transmission is sometimes explained in terms of a two-wire transmission line with an infinite number of quarter-wave shorted stubs, each of which has an infinite impedance and therefore looks like an open circuit. Figure 13-160 shows a two-wire

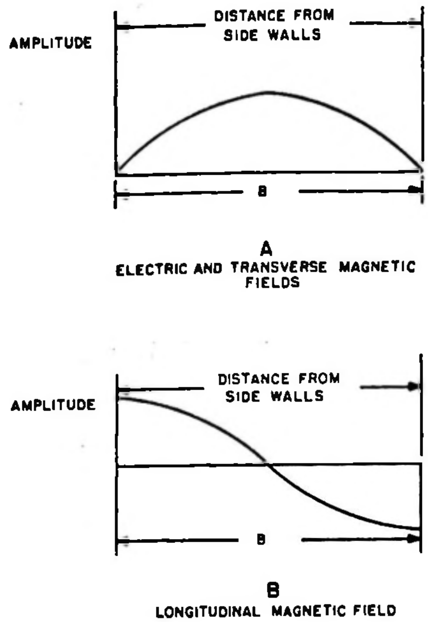


Figure 13-159. Field Amplitude

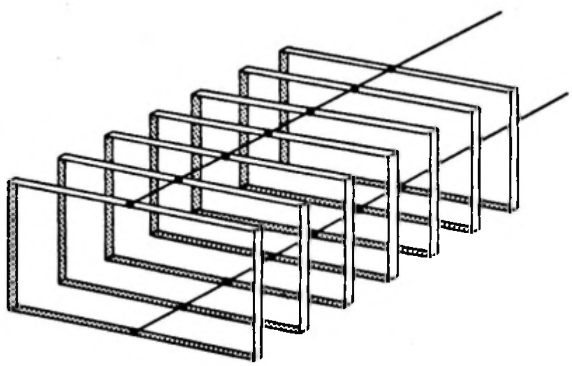


Figure 13-160. Two-Wire Transmission Line with Quarter-Wave Shorted Stubs

line with a large number of shorted stubs; the similarity to a waveguide is immediately apparent. This explanation does not cover the field pattern nor the group and phase velocities. It implies that transmission is possible only if the width of the guide is one half-wavelength, which is not correct since transmission is possible only when the width of the guide exceeds one half-wavelength. It may be useful to remember, however, that a waveguide at the cutoff frequency (where it will not carry energy) looks like a two-wire line with an infinite number of quarter-wave shorted stubs.

13-659. GUIDE WAVELENGTH.

13-660. In addition to λ_a and λ_c, it is customary to refer to the guide wavelength (λg). This is the wavelength of the pattern in the waveguide, and is equal to $\lambda_a/\cos B$. The relationship can be derived from the diagram in figure 13-161. Here, CD is a wave front of the signal that moves to the left; point E is on the left-hand wall, directly opposite point D, and EF is another wave front of the same signal. Distance DG is equal to b sin B and, from previous discussions, it is clear that this must be $\lambda_a/2$. The two wave fronts, CD and EF, are,

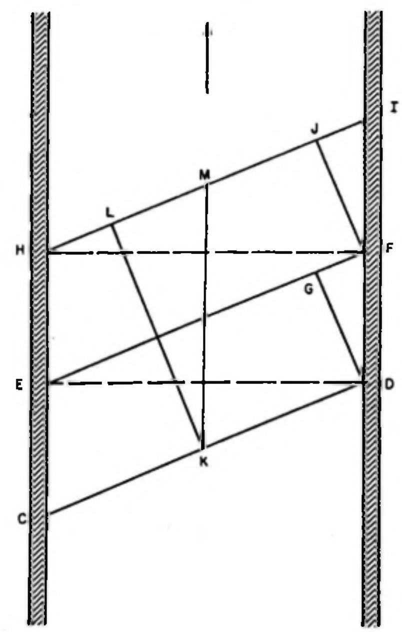


Figure 13-161. Geometry of Equation for Derivation of Guide Wavelength

therefore, separated by one-half λ_a. Point H is on the left-hand wall, directly opposite point F, and HI is another wave front. By an argument similar to the one given above, the spacing between EF and HI is one-half λ_a. Then KL, the spacing between CD and HI, is exactly equal to λ_a. The wavelength of the pattern, however, is equal to KM, since this is the distance along the guide axis between two points where the phase of the signal differs by one cycle. The angle between KL and KM is B; therefore, the following relationships hold true:

$$KL = \lambda_a = KM \cos B = \lambda_g \cos B$$

and:

$$\lambda_g = \frac{\lambda_a}{\cos B}$$

13-661. GROUP AND PHASE VELOCITIES.

13-662. Just as two wavelengths, λ_a and λ_g, must be considered in connection with the transmission of energy along a waveguide, three separate velocities are of concern. The first of these is the velocity of electromagnetic waves in space, usually identified as the letter c, which is equal to 330 yards or 300 meters per microsecond. The other two are group velocity (vg), which is the velocity with which energy is transmitted along the waveguide, and phase velocity (vp), which is the velocity with which the field pattern moves along the waveguide.

13-663. DISTINCTION BETWEEN VELOCITIES.

13-664. Imagine a transmission equipment, similar to the one shown in figure 13-162, that consists of a vertical glass tube through which water is pumped. Simultaneously, air is introduced at the bottom and rises in the form of small, regularly spaced bubbles. Suppose, now, that clear water is pumped through the tube. The rising bubbles will form a pattern which moves upward at a pace that may be called the pattern velocity. If dyed water is then introduced, the boundary between the clear water and the colored water will rise at a speed that may be called the transport velocity (the velocity at which water moves through the tube). This equipment behaves in a manner similar to that of a waveguide. If a steady signal is sent through the guide, the field pattern will have a wavelength λ_g, analogous to the spacing between the bubbles. The field pattern will move with a phase velocity, just as the bubble pattern moves at the pattern velocity. If, then, the signal level is raised suddenly, the boundary between the low-level signal and the high-level signal will move along the waveguide at a group velocity, just as the boundary between the clear water and the dye solution moves at the transport velocity.

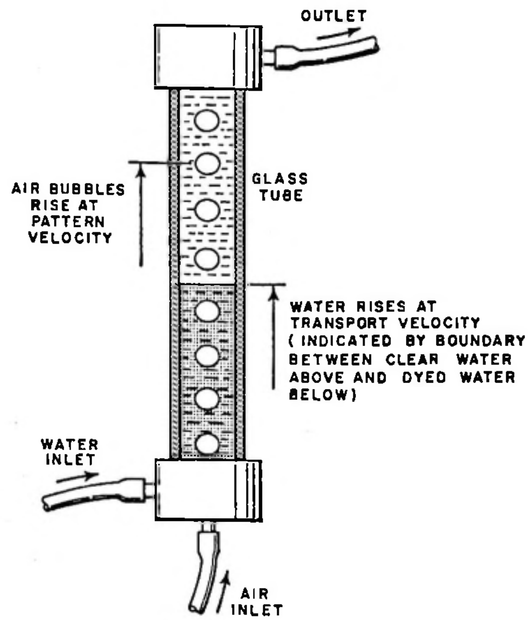


Figure 13-162. Pattern and Transport Velocities

13-665. VARIATION OF VELOCITIES WITH WAVELENGTH.

13-666. As explained in previous paragraphs, the cutoff wavelength, λ_c, is equal to 2b, and the angle B is determined by the following relation:

$$\sin B = \lambda_a/2b = \lambda_a/\lambda_c$$

When the operating frequency is extremely high, so that the wavelength of the signal (in air) is very much shorter than the cutoff wavelength of a waveguide, sin B and B are both very small and cos B is nearly 1. In this case the phase velocity is only slightly greater than c and the group velocity is only slightly smaller. At the other extreme, when the wavelength in air is only slightly shorter than the cutoff wavelength, sin D is very nearly 1 and B is nearly 90 degrees. Cos B is then very small, so that v_g becomes small and v_p becomes extremely large. Similar behavior can be observed at any straight lake front or ocean beach. Figure 13-163 shows a series of wave crests moving in a direction that makes an angle x with the edge of the beach. For example, assume that one of the wave crests moves from P to Q; the intersection of that crest with the beach moves from R to Q. By simple geometry, distance PQ is equal to distance RQ times cos X. Then, if the waves

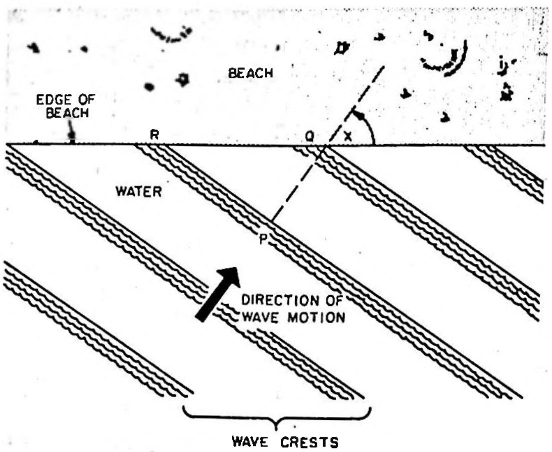


Figure 13-163. Water Waves Breaking on Beach

move forward with a velocity v_w, and the point at which the wave crest breaks on the beach moves along the edge of the beach with a velocity v_b, then:

$$PQ = v_w t; RQ = v_b t$$

where t is the time required for the wave crest to move from P to Q. Also:

$$v_w t = v_b t \cos X$$

Thus:

$$v_b = v_w / \cos X$$

The point of intersection of a wave crest with the edge of the beach, therefore, runs along the beach 1/cos X times as fast as the wave pattern moves.

13-667. OTHER MODES OF TRANSMISSION.

13-668. A number of individual electromagnetic waves can combine in many different ways to carry energy down the interior of a rectangular waveguide. The various ways are called modes, and each mode has a distinctive field pattern that is different from that of every other mode. Each mode also has its own cutoff frequency, cutoff wavelength, guide wavelength, group velocity, and phase velocity, which are not necessarily different from those of other modes. In a rectangular waveguide, it is customary to choose dimensions that make the cutoff frequency higher than the operating frequency for all modes except the mode being used. For all modes except this one, the operating frequency is below the cutoff frequency, and the operating wavelength is longer than the cutoff wavelength, so that transmission is impossible. The waveguide, therefore, operates only in the mode being used. In spite of this common engineering practice, the other modes should be considered briefly.

13-669. MODE TERMINOLOGY.

13-670. In some modes, the electric field is always transverse (where it does not vanish), whereas the magnetic field in these modes is sometimes transverse and sometimes longitudinal. This is true, for example, with the simple mode described above. Such modes are called TE (transverse electric) modes. In other modes, the magnetic field is always transverse, whereas the electric field is sometimes transverse and sometimes longitudinal. These are called TM (transverse magnetic) modes. In still other modes, both the electric field and the magnetic field are longitudinal at various points. These modes are not named, since a mode of this kind can always be considered as a combination of a given TE mode and a given TM mode. A mode is defined further by two subscripts, m and n, as a $TE_{m,n}$ or a $TM_{m,n}$ mode. The first subscript is the number of times a pattern repeats itself along the a dimension of the waveguide (from the top of the guide to the bottom), and the second subscript is the number of times the pattern repeats itself along the b dimension (from one side of the waveguide to the other). In the simple case described earlier, the field pattern is uniform from top to bottom, so that there is no repetition and the first subscript (m) is therefore equal to zero (0). There is variation, however, from one side to the other, but the pattern occurs only once; therefore, the second subscript (n) is 1. The simple mode is, accordingly, called the $TE_{0,1}$ mode (since the electric field is entirely transverse). The following discussion shows that the $TE_{0,1}$ mode has a longer cutoff wavelength than any other mode for a given rectangular waveguide. Because of this, it is possible to choose the waveguide dimensions so that λ_c is longer than λ_a for this mode, and shorter than λ_a for every other mode. The waveguide is thereby forced to operate in the $TE_{0,1}$ mode, as explained above.

13-671. TWO NONEXISTENT MODES.

13-672. It is impossible to send a single vertically polarized wave down a waveguide along a path that is parallel to the axis of the guide. The field pattern of such a wave will have an electric field which is parallel to each side wall at the surface of the wall. This electric field will cause currents to flow in the side walls, and the electromagnetic waves radiated by these currents will cancel the original signal in a very short distance. The same thing is true of a horizontally polarized wave that travels directly down the guide without reflection. Here there will be electric fields which are parallel to the top and bottom of the guide, and the same result will be obtained. Finally, any electromagnetic wave may be considered as a combination of vertically and horizontally polarized waves. It follows, therefore, that it is impossible to send energy for any great distance through a rectangular waveguide by means of any signal that travels directly down the guide without reflection. The $TE_{0,0}$ and $TM_{0,0}$ modes have field patterns that are uniform along both the a and b dimensions. The reflections from the side walls cause the field pattern to vary along the b dimension, whereas the absence of reflection from the top and bottom of the waveguide makes the pattern uniform along the a dimension. Thus, a field pattern which is uniform in the a and b dimensions can be obtained only if there is no reflection from any of the walls. However, it is impossible to transmit energy for any great distance through a rectangular waveguide by means of waves that are not reflected, and for this reason neither the $TE_{0,0}$ nor the $TM_{0,0}$ mode can exist, from a practical standpoint, in a rectangular waveguide.

13-673. $TE_{0,n}$ MODES.

13-674. It has been stated that energy can be carried along a waveguide by vertically polarized waves traveling parallel to the top and bottom of the guide if their paths make the proper angle with the side walls; it has also been established that the proper angle is given by either of the following equations:

$$2b \sin B = n \lambda_a$$

or:

$$\sin B = n \lambda_a / 2b$$

where B is the angle between the paths of the individual waves and the side walls. It has been shown that the $TE_{0,1}$ mode is produced when n is equal to 1, although the mode was not so labeled at the time. For any value of n (except zero), this arrangement of vertically polarized waves reflected from the side walls produces a $TE_{0,n}$ mode. This may be verified by examination of the diagram shown in figure 13-164, which treats the case where n is equal to 3. Here C and D are two points opposite each other on the side walls of a rectangular waveguide. The distance DF is n half-wavelengths (three in this case); therefore, wave front GD is three half-wavelengths behind the wave front CE.

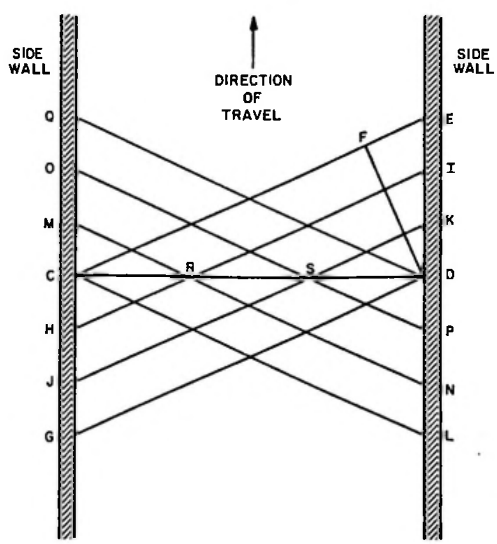


Figure 13-164. Geometry of $TE_{0,3}$ Mode

It is possible to construct additional wave fronts HI and JK, so that each is one half-wavelength from its neighbors. In the same way, wavefronts QD, OP, MN, and CL may be constructed, each being one half-wavelength from its neighbors. Now as one moves from C to R, the phase of the waves moving to the left changes one half-cycle in one direction, and the phase of the waves moving to the right changes an equal amount in the other direction. This leads to a pattern similar to that obtained by moving all the way across the guide in the $TE_{0,1}$ mode. Conditions at R are the same as those at C, except that the phases of both sets of waves are reversed. Therefore, the pattern between R and S is the same as the pattern between C and R, but the directions of the fields are reversed. The same thing happens between S and D, with the directions of the fields the same as those between C and R. Consequently, the pattern for a $TE_{0,n}$ mode is nothing more than a repetition of the $TE_{0,1}$ pattern, with the field directions reversed at each repetition.

13-675. Figure 13-165 illustrates the electric field pattern in a plane perpendicular to the axis of a rectangular waveguide; part A shows the $TE_{0,1}$ mode, part B shows the $TE_{0,2}$ mode, and part C shows the $TE_{0,3}$ mode. In part B, the direction of the electric field in the right half of the waveguide is opposite that in the left half. In one half, the field is directed upward, and in the other half it is directed downward. Similarly, in part C, the direction in the center of the guide is opposite the direction in either of the side portions of the guide. In every case, however, the field is either upward or downward. Figure 13-166 shows the combined fields for the $TE_{0,3}$ mode.

13-676. OTHER NONEXISTENT MODES.

13-677. At first glance, it would appear possible to develop the $TM_{0,n}$ mode by sending horizontally polarized signals along

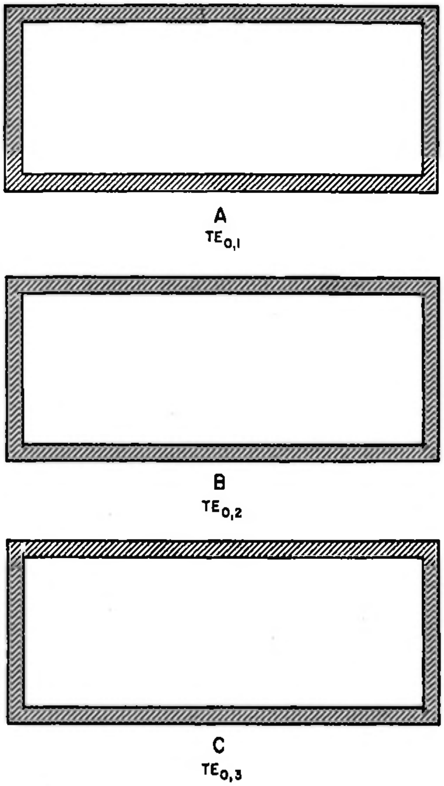


Figure 13-165. Electric Field Patterns for $TE_{0,n}$ Modes

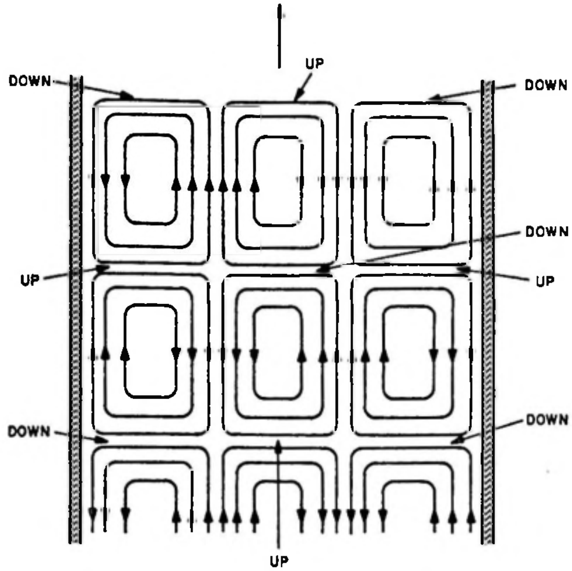


Figure 13-166. Combined Fields for $TO_{0,3}$ Mode

the paths prescribed for the signals that produce a $TE_{0,n}$ mode. This would produce a pattern of the desired type, but there would be longitudinal electric fields at the walls and the pattern would canceled by the waves radiated by induced wall currents. The same objection applies to the possibility of developing a $TM_{m,0}$ mode by sending vertically polarized waves along the paths that produce a $TM_{m,0}$ mode. For this reason, neither the $TM_{0,n}$ modes or the $TM_{m,0}$ modes can exist in a rectangular waveguide.

13-678. REFLECTION FROM FOUR WALLS.

13-679. All of the modes previously considered have patterns that are built up from two sets of waves; before considering the $TE_{m,n}$ and $TM_{m,n}$ modes, you should consider what happens when the signals are reflected from all four walls of the waveguide. The pattern will then be built up from four sets of waves, those moving downward to the right and left, and those moving upward to the right and left. To begin, consider the two sets of waves that move downward to the right and left. Their paths meet the side walls of the waveguide at some angle, B, which satisfies the following equation:

$$\sin B = n \lambda_a/2b$$

The two sets of waves combine to form a field pattern like that of the $TE_{0,n}$ mode, except that it is tipped downward and therefore travels downward. A similar pattern, tipped in the opposite direction and traveling upward, is formed by the two sets of waves that move upward to the right and left. These two patterns follow paths that meet the top and bottom of the waveguide at angle A. In figure 13-167, part A shows the paths from above and part B shows them from the side. Since each pattern is composed of two sets of electromagnetic waves which are reflected by the top and bottom of the waveguide, the pattern as a whole is reflected, and behaves generally, in a manner similar to one of the sets of waves in a $TE_{m,0}$ mode.

13-680. BEHAVIOR OF TIPPED PATTERNS.

13-681. Each of the patterns formed by combining two sets of waves has a value of $\lambda_a/\cos B$, where the wavelength is measured in a direction parallel to the path of the patterns. The pattern moves along the path, as shown in figure 13-167, with a velocity of c/cos B; energy moves along the path with a velocity of c cos B. The wavelength and the two velocities are equal to the guide wavelength and the phase and group velocities of the $TE_{0,n}$ mode. As might be expected, the pattern cannot exist unless λ_a is shorter than 2b/n, the cutoff wavelength for the mode. The two patterns combine in the same way as the two sets of waves that form a $TE_{m,0}$ mode, and angle A must satisfy the following equation:

$$\sin A = m\,\lambda/2a$$

where λ is the wavelength of each of the patterns, or $\lambda_a/\cos B$.

13-682. FIELD CONFIGURATION OF COMBINED PATTERN.

13-683. The general appearance of the combined pattern is shown in figure 13-168; the direction of the electric field is indicated by solid lines, the direction of the magnetic field is indicated by dashed lines. The pattern for vertically polarized waves is shown in part A of the figure, and the pattern for horizontally polarized waves is shown in part B. In each pattern, both m and n are

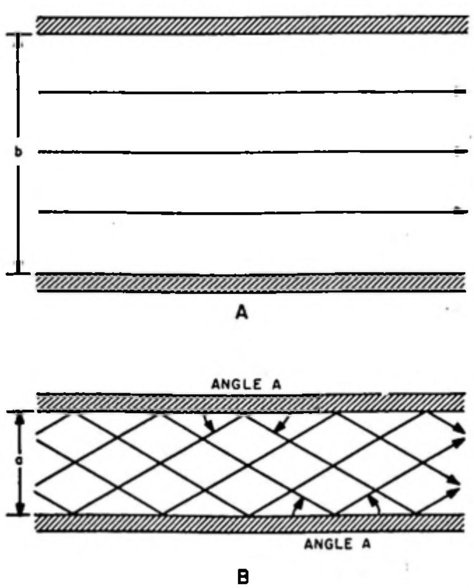


Figure 13-167. Paths of Two-Wave Patterns

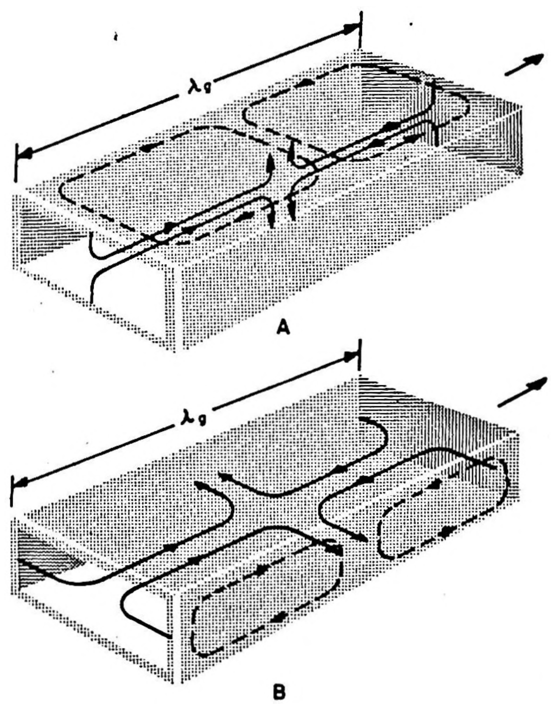


Figure 13-168. Configuration of Combined Pattern

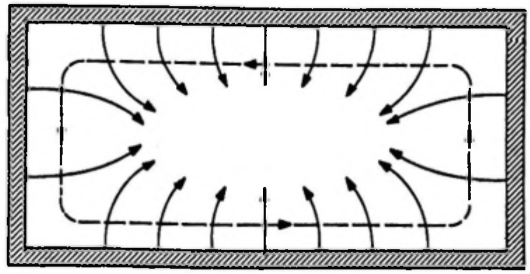
equal to 1. The pattern for higher values of m and n is simply a repetition of the simple pattern, either vertically or horizontally across the guide. To simplify the diagrams, most of the lines showing field direction have been omitted. The direction of the electric field is shown in a plane through the center of the guide, a vertical plane for the vertically polarized waves and a horizontal plane for the horizontally polarized waves. Similarly, the direction of the magnetic field is shown only near the top of the waveguide and near the right-hand wall. The pattern of the magnetic field near the bottom of the waveguide is the same as the pattern near the top, with the directions reversed. A similar situation, with identical patterns and reversed directions, exists for the magnetic fields near the two side walls.

13-684. $TE_{m,n}$ AND $TM_{m,n}$ MODES.

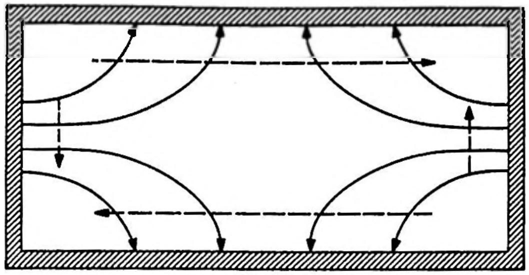
13-685. If eight sets of electromagnetic waves are present in a rectangular waveguide, both of the patterns shown in figure 13-168 will be present. If the phase relation between the two patterns is the same as that shown in the drawing, the longitudinal component of the magnetic field will be canceled and the result will be a $TM_{m,n}$ mode. If, on the other hand, the phase of one of the patterns is reversed (by reversing the phases of the four sets of waves that contribute to it), the longitudinal electric field will be canceled, and a $TE_{m,n}$ mode will result. In parts A and B of figure 13-169, the directions of the resulting electric and magnetic fields are shown in a plane across the waveguide at a point where the transverse field is maximum.

13-686. SINGLE-MODE OPERATION OF RECTANGULAR WAVEGUIDE.

13-687. Assume that the dimensions of a rectangular waveguide are chosen so that λ_c is shorter than λ_a for the $TE_{1,0}$ mode.


A

$TM_{1,1}$


B

$TE_{1,1}$

Figure 13-169. $TM_{1,1}$ and $TE_{1,1}$ Modes

The waveguide then cannot carry energy at the wavelength of λ_a by means of the $TE_{1,0}$ mode; therefore, this mode is said to be suppressed for λ_a and the corresponding frequency. The cutoff wavelength becomes shorter as either m or n is made larger. Consequently, all modes for which m is equal to or greater than 1 will be suppressed. (It is not necessary to specify the values of n because any value will be equal to or greater than zero). The only possible modes are, therefore, those for which m is zero. Since the $TM_{0,n}$ modes do not exist, the only possible modes are the $TE_{0,n}$ modes. If, in addition, the $TE_{0,2}$ mode is suppressed, all modes for which n is equal to or greater than 2 will also be suppressed. There remains only the $TE_{0,1}$ mode (since neither the $TE_{0,0}$ mode nor the $TM_{0,0}$ mode exists). This preceding procedure is normally followed when using a rectangular waveguide. Height a, shown in part B of figure 13-167, is made less than $\lambda_a/2$ so that the $TE_{1,0}$ mode, which has a cutoff wavelength of 2a, is suppressed. The $TE_{0,2}$ mode, which has a cutoff wavelength of b, is suppressed by making b less than λ_a. The cutoff wavelength for the $TE_{0,1}$ mode is 2b, and if the width is more than $\lambda_a/2$ (but less than λ_a), the waveguide will operate in this mode only.

13-688. DIMENSIONS OF RECTANGULAR WAVEGUIDE.

13-689. It is desirable to use waveguides which are as large as possible, provided that you suppress undesired modes. Large waveguides have less attenuation than smaller waveguides, just as large wire has smaller dc losses (because of the wire's lower resistance) than small wire, when both carry the same current. Modes other than $TE_{0,1}$ mode are suppressed for much the same reason that standing waves on a transmission line are suppressed. Both absorb power, and both interfere with the proper operation of the waveguide by changing the impedance. It is customary to use 1-1/2- by 3-inch rectangular waveguide for S-band operation, and 1/2- by 1-inch rectangular waveguide for X-band operation. Some of the characteristics of these two sizes of rectangular waveguides are given in table 13-4. In practice, the shortest operating electrical wavelength is 5 or 10 percent longer than the physical wavelength at which double-mode operation is possible. This insures that the waveguide will operate only in the $TE_{0,1}$ mode, even when there is a small error in the operating frequency. The longest operating wavelength is about 80 percent of λ_c for the $TE_{0,1}$ mode. The waveguide will operate at longer wavelengths (as long as they are shorter than λ_c), but the attenuation increases rapidly as the operating wavelength approaches the cutoff wavelength. A good rule of thumb for estimating the operating wavelength of an odd-size rectangular waveguide is that the nom-

Table 13-4. Rectangular Waveguide Characteristics

	NOMINAL OPERATING WAVELENGTH	
	(S-BAND) [a]	(X-BAND) [b]
Outside dimensions in inches	1.500 x 3.000	.500 x 1.000
Wall thickness in inches	.080	.050
Inside dimensions in inches	1.340 x 2.840	.450 x .950
Inside dimensions in centimeters	3.40 x 7.22	1.14 x 2.41
λ_c, $TE_{0,1}$ mode	14.44 cm	4.82 cm
λ_c, $TE_{0,2}$ mode	7.22 cm	2.41 cm
λ_c, $TE_{1,0}$ mode	8.80 cm	2.28 cm

a = S-Band—10 cm.
b =X-Band— 3 cm.

inal operating wavelength is about five-fourths of the larger dimension.

13-690. CIRCULAR VERSUS RECTANGULAR WAVEGUIDES.

13-691. GENERAL.

13-692. The behavior of circular waveguides is similar to that of rectangular waveguides. The field patterns, like those of rectangular waveguides, satisfy the following two boundary conditions, imposed by the fact that the waveguide wall is a good conductor: At the surface of a wall there can be no varying electric field parallel to the wall, and there can be no varying magnetic field perpendicular to the wall. A review of the material on rectangular waveguides will show that these boundary conditions are satisfied by the field patterns of all existing modes. Because of the curved walls of a circular waveguide, it is extremely difficult to examine the field patterns in the same detail as those of a rectangular waveguide; there is no simple method of building the field pattern from a small number of simple waves traveling along the guide. The calculation of cutoff wavelength for a given mode in a circular waveguide is beyond the scope of this manual; it will be necessary to accept the values given here without any demonstration of their validity. The general configuration of the field patterns may be deduced, however, from their similarity to corresponding patterns in a square waveguide (a rectangular waveguide for which dimensions a and b are equal).

13-693. TERMINOLOGY. As with rectangular waveguides, all circular waveguide modes are designated as TE modes, TM modes, or combinations of these two, and these modes have the same significance as they had in rectangular waveguides. The modes are described by the same two subscripts, so that in the circular waveguide there are $TE_{m,n}$ modes and $TM_{m,n}$ modes. In a rectangular waveguide, m is the number of times a field pattern is repeated from the top of the guide to the bottom, with a reversal in the directions of the electric and magnetic fields at each repetition, and n is the number of times the pattern is repeated from one side wall to the other. In circular waveguide, however, m and n have different meanings; m is the number of times the pattern is repeated around the waveguide, and n is the number of times the pattern is repeated between the center of the guide and the wall.

13-694. IMPORTANT MODES. In considering the behavior of a circular waveguide, it is necessary to examine only the $TE_{1,1}$ $TM_{0,1}$ and $TE_{2,1}$ modes. These are the three simplest modes for circular waveguide operation, and the only ones for which the cutoff wavelength is greater than the diameter of the guide. $TE_{1,1}$ is the dominant mode (the mode having the longest cutoff wavelength) for a circular waveguide; it has a cutoff wavelength of 1.71d, where d is the inside diameter of the waveguide. Part A of figure 13-170 shows the electric field pattern for the $TE_{1,1}$ mode in a circular waveguide, and part B shows the electric field for the $TE_{0,1}$ mode in a square waveguide (a rectangular guide for which a and b are equal). The resemblance between the fields in the circular and square waveguides is obvious,

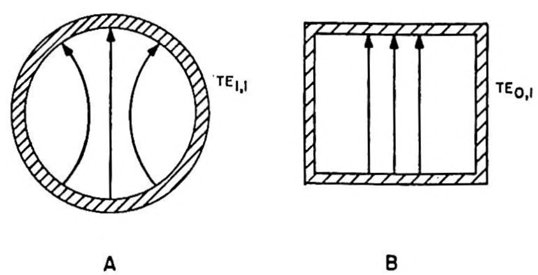


Figure 13-170. Dominant TE Modes in Circular and Square Waveguides

and both modes are dominant modes. The $TM_{0,1}$ mode in a circular waveguide, with a cutoff wavelength of 1.31d, is shown in part A of figure 13-171; the $TM_{1,1}$ mode in a square waveguide is shown in part B. The two diagrams are simplified by omitting the lines showing the direction of the magnetic field. The $TE_{2,1}$ mode in a circular waveguide, with a cutoff wavelength of 1.03d, is shown in part A of figure 13-172. The corresponding mode in a square waveguide, the $TE_{1,1}$ mode, is shown in part B of the figure. All other modes, both TE and TM, have cutoff wavelengths of 0.82d or less.

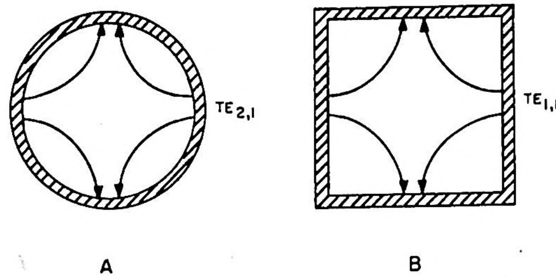


Figure 13-172. Simplest TE Modes in Circular and Square Waveguides

13-695. USES. The usefulness of circular waveguides lies in the perfect symmetry of the $TM_{0,1}$ mode. A rotating joint can be inserted in a section of circular waveguide without affecting the transmission of energy. Therefore, you can use a section of circular waveguide to join two sections of rectangular waveguide when it is necessary to rotate one with relation to the other, as shown in figure 13-173. Moving the upper rectangular waveguide from the position shown in part A of the figure to the position shown in part B has no effect on the transmission of energy. Although it is possible to use the $TE_{1,1}$ mode in a circular waveguide, it is difficult with this mode to control the plane of signal polarization; therefore, you will normally use the corresponding $TE_{0,1}$ mode in the rectangular waveguide instead.

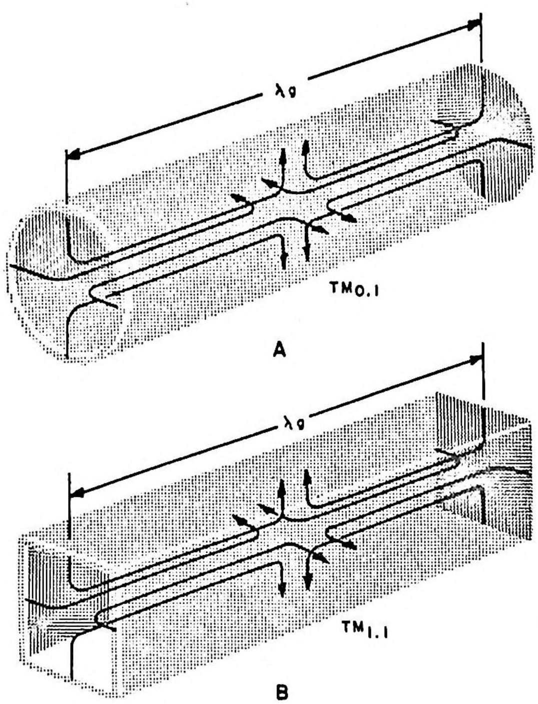


Figure 13-171. Simplest TM Modes in Circular and Square Waveguides

13-696. DIMENSION CRITERIA. In choosing a circular waveguide, you must always make sure that λ_a is shorter than 1.31d, the cutoff wavelength for the $TM_{0,1}$ mode. The higher TM modes are then automatically suppressed by choosing the waveguide diameter so that λ_a is longer than 0.82d. The method of exciting the waveguide suppresses the $TE_{1,1}$ and $TE_{2,1}$ modes unless the latter of these is already suppressed by your choice of diameter selection. It is customary to excite the guide with a small antenna at the guide axis, as shown in figure 13-174. The field radiated by this antenna has no longitudinal magnetic component (which must be present for any TE mode), and, consequently, neither the $TE_{1,1}$ nor the $TE_{2,1}$ mode is excited. Mechanically, it is desirable to use circular waveguides with a diameter equal to the width of the rectangular waveguide in the arrangement, so that d

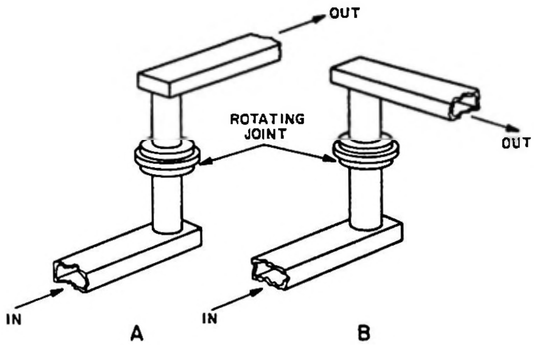


Figure 13-173. Rotating Joint in Circular Waveguide

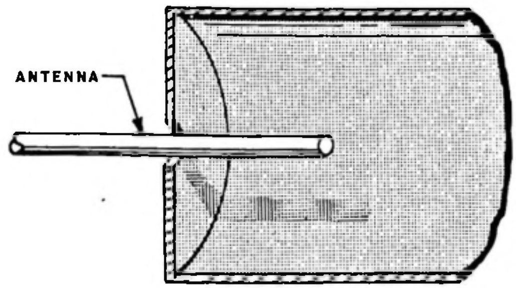


Figure 13-174. Excitation of Circular Waveguide

is normally about 0.8 λ_a, but this is not always done. The circular waveguide normally used in S-band operation has a diameter of about 3 inches, and that used in X-band operation has a diameter of about 1 inch.

13-697. WAVEGUIDE ADVANTAGES. The major advantage of a waveguide over a coaxial line is the much lower dc loss in a waveguide, resulting primarily from the elimination of the center conductor. At vhf and microwave frequencies, interaction between the guided waves in a transmission arrangement and the current-carrying electrons causes the current flow to be confined to the surface of the conductors. This skin effect, of course, is also present at lower frequencies, but in comparison with the dc resistive value of the conductor it is unimportant. However, at vhf and microwave frequencies, because of the high resistances, only the surface of the inner conductor of coaxial line is available for current flow, and its resistance is unavoidably high. The outer conductor has a larger surface (the inner surface is the one that carries the current), and therefore has a much lower resistance. The major portion of the loss, therefore, occurs on the inner conductor of the coaxial line. The inner conductor is not present in a waveguide; therefore, a waveguide is not subject to inner-conductor loss. A waveguide has other advantages over a flexable coaxial line, such as lack of radiation and dielectric losses, but a rigid coaxial line offers the same advantages. The greatest merit of a waveguide, aside from the fact that it has lower conductor losses (sometimes called copper loss) than a coaxial line, is its extreme simplicity of construction. By comparison with a waveguide, a rigid coaxial line with a quarter-wave stub support is a complicated mechanical device.

13-698. WAVEGUIDE IMPEDANCE.

13-699. The impedance of a transmission line at any point is the ratio between the voltage and the current at that point. It is necessary, of course, when computing this ratio, to consider the phase relationship between the voltage and the current. The impedance of a waveguide at any point is, similarly, the ratio between the transverse magnetic fields at that point. As with transmission line impedance, it is necessary, when computing the ratio, to take into account the phase relationship between the two fields. Part A of figure 13-175 shows the fields associated with a coaxial line, and part B shows the fields in a conventional two-wire line. In both, the transmission line acts as a guide for the waves that ac-

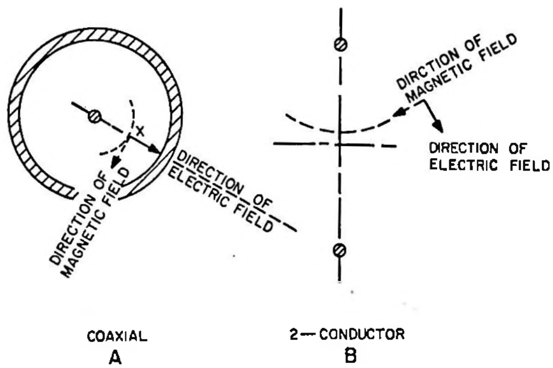


Figure 13-175. Fields Associated with Transmission Lines

company any transmission of energy. At any point where the guided wave is present, such as point X in part A, and point Y in part B, of the figure, electric and magnetic fields are associated with the guided waves. The transverse component of the electric field (the field in a plane perpendicular to the conductors) depends on the position of the point and on the voltage between the conductors. For any given point, the strength of the electric field is proportional to the voltage. In the same way, the transverse magnetic field depends on the position of the point and on the current carried by the conductors. For any given point, the strength of the magnetic field is proportional to the current. Both the transverse electric field and the transverse magnetic field depend in the same way on the position of the point, so that the ratio between the fields does not depend on the position of the point. The ratio between the two transverse fields is proportional to the ratio between the voltage and the current; it follows that the impedance of a transmission line can be defined either by the ratio between the voltage and the current or by the ratio between the transverse electric field and the transverse magnetic field. The second way of defining the impedance can be applied directly to a hollow waveguide. The waveguide, then, has a characteristic impedance associated with each mode (although several modes may have the same characteristic impedance). In the presence of a reflected wave, the impedance is different from the characteristic impedance and is also different at different points along the waveguide. The impedance is always the same at all points in a plane perpendicular to the axis of the waveguide, just as the impedance of a transmission line (computed from the electric and magnetic fields instead of from voltage and current) is the same, whatever point in a given plane is chosen as a place to measure the fields.

13-700. WAVEGUIDE TUNING DEVICES.

13-701. It seems inappropriate, at first glance, to discuss the impedance of a waveguide, since there is no point along the guide where the voltage and the current can be measured. The concept of waveguide impedance is, however, as useful as the concept of transmission line impedance.

13-702. SLUG TUNER.

13-703. A slug tuner is used with a waveguide transmission equipment in precisely the same way that the stub tuner is used with a conventional transmission line; the slug tuner is essentially the same device as the stub tuner. Most waveguide arrangements consist largely of a rectangular waveguide operating in the $TE_{0,1}$ mode; the double-slug tuner shown in figure 13-176 is designed for this type of waveguide. Part A of the figure shows a section through the tuner, and part B shows an end view. Each tuning slug projects into the waveguide so that it is parallel to the electric field, and acts as a small antenna. It is excited by the electric field in the waveguide, and, consequently, radiates a signal. The phase of the signal radiated may be altered by changing the length of the slug with the tuning screw. Ideally, the signals reflected

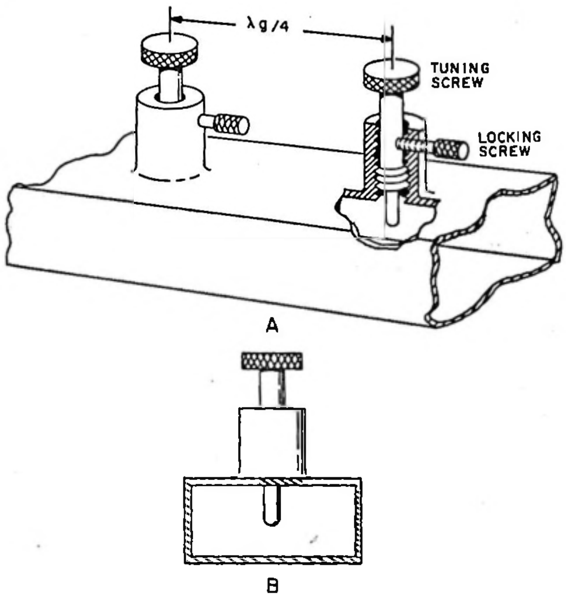


Figure 13-176. Double-Slug Tuner for $TE_{0,1}$ Mode in Rectangular Waveguide

from the slugs will cancel the signal reflected back from the load. The tuning stub has no reflection effect when it is exactly a quarter-wavelength long; its length may be varied in either direction from this point. The slug is ineffective when it is backed out to a position even with the waveguide wall, and it can be varied in only one direction from that point. Consequently, the double-slug tuner is limited to some extent, and three-slug tuners are frequently used with a spacing of 3/8 λ_g between the adjacent slugs. A double-slug tuner is adjusted by starting with both slugs retracted so that their tips are even with the wall of the waveguide, the position where they have the least effect. One of the slugs is then inserted by advancing the tuning screw. If this increases the amplitude of the reflection, that slug is retracted and the other is advanced. From this point, tuning proceeds by alternate adjustment of the two slugs for minimum reflection.

13-704. STUB TUNER.

13-705. Although actual tuning stubs are used occasionally, in most applications they offer little advantage over the slug tuner previously discussed. A typical tuning stub is shown in figure 13-177. It consists of a section of waveguide with an adjustable plunger of conducting material. To insure good contact between the plunger and the wall of the waveguide and, therefore, an effective short circuit for the termination, the plunger is grooved a short distance from its face. The groove, when seen from point B, appears as a short section of transmission line terminated by the short circuit at point A. The groove depth is chosen so that this section of line has an effective length of 1 quarter-wavelength. The impedance at point B, therefore, is the infinite impedance of a shorted quarter-wave section of line, and this is in series with the impedance of the rubbing contact between the plunger and the waveguide wall to the left of point B. Distance B to C is also 1 quarter-wavelength, and the impedance of the gap between the plunger and the wall at point C is that of a quarter-wave section terminated by the impedance at point B. Since the impedance at point B is always great (because of the impedance of the groove), the impedance at point C is almost zero, whether or not there is good contact between the plunger and the wall. The groove in the plunger is called a <u>choke groove</u>. The stub may be added to either wall of the waveguide.

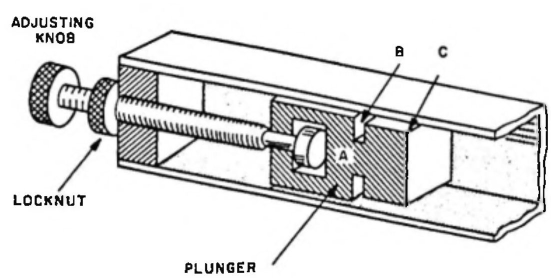


Figure 13-177. Adjustable Tuning Stub

When the junction is at the wall perpendicular to the electric field, it is called an E-junction, the stub is called a series stub, and the impedance of the stub is in series with that of the main line. When the junction is at the wall perpendicular to the magnetic field, it is called an H-junction, the stub is called a shunt stub, and the impedance of the stub is in parallel with that of the main line. Part A of figure 13-178 shows a stub connected with an E-junction, and part B shows the stub connected with an H-junction. In practice, the E-junction is normally used in preference to the H-junction.

13-706. WINDOW TUNERS.

13-707. It is always possible to tune a waveguide by any method that introduces the required reflection; this is sometimes done by means of window tuners, as shown in figure 13-179. Since it is desirable for the two obstructions to be symmetrical, it is rather difficult to construct an adjustable window tiner. However, window tuners are useful whenever they can be preset and left alone. Electrically, they are equivalent to shunt tuning stubs. The window tuners are nothing more than small conducting plates inserted in the guide. The tuner shown in part A of the figure is located where the electric field is relatively weak; thus, its major effect is imposed on the magnetic field. Since the magnetic field is analogous to current in a waveguide, a tuner of this kind behaves as a current device; it is approximately the equivalent of a shunt inductor (which does not affect the voltage) on a conventional transmission line. The tuner shown in part B affects the electric field more strongly than the magnetic field, and is thus approximately the equivalent of a shunt capacitor.

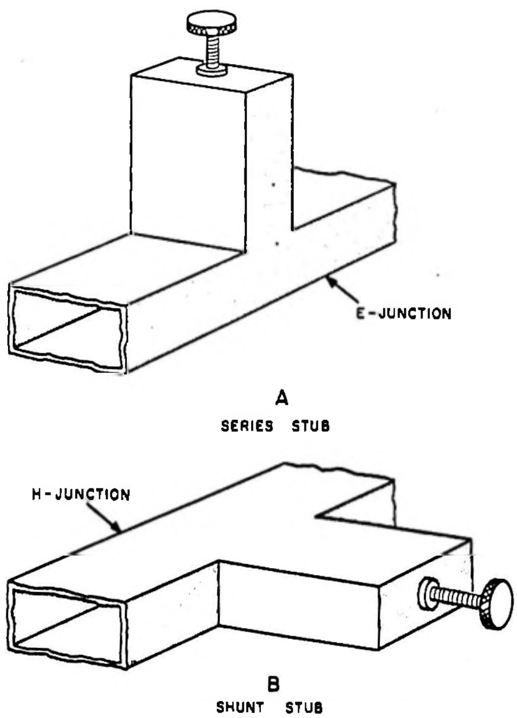


Figure 13-178. Shunt and Series Tuning Studs

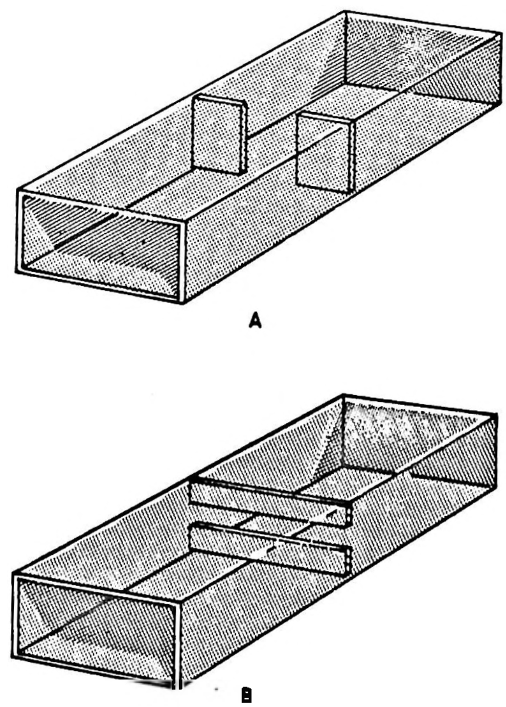


Figure 13-179. Window Tuners

13-708. WAVEGUIDE OBSERVATION AND MEASUREMENT TECHNIQUES.

13-709. SLOTTED SECTIONS.

13-710. Whenever a waveguide carries both direct and reflected signals, the strength of the electric field varies from point to point along the guide, just as the voltage varies in a conventional transmission line. A slotted section in connection with a probe is used to observe the load power. Following transmission line terminology, the ratio between the maximum and the minimum electric field strength is called the vswr. Figure 13-180 shows a cross section of the carriage which rides on a slotted section of rectangular waveguide. The probe, which is parallel to the electric field, is excited by the signal and develops a voltage at the operating frequency. A crystal rectifier is mounted in the carriage, and the output of the crystal is fed to a meter. You can generally use a sensitive meter directly, but an amplifier may be required. Figure 13-180 also shows a crystal holder with a fitting for a coaxial cable.

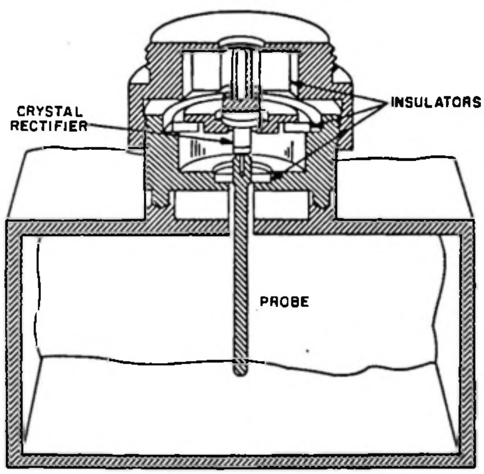


Figure 13-180. Slotted Section

13-711. DIRECTIONAL COUPLERS.

13-712. You can use a directional coupler to measure the reflected signal in a waveguide. The directional coupler requires no manipulation, and can be sealed to make it a permanent part of a pressurized waveguide arrangement. Figure 13-181 shows the basis for the theory underlying the use of the directional coupler. In this figure, a main waveguide has an auxiliary waveguide coupled to it through two small holes. Refer now to part A of figure 13-182; it is seen that when a signal travels from left to right in the main waveguide, some energy leaks into the auxiliary guide through the holes at points WX and YZ, and that the signal in the auxiliary guide at Z is composed of two parts. Some energy leaks through the hole at WX and travels from X to Z. Additional energy travels from W to Y and then leaks through the hole at YZ. Since distance WY and XZ are equal, the two parts of the signal will be in phase and some energy will be propagated from left to right in the auxiliary guide. Propagation of a signal in the opposite direction in the auxiliary guide, as illustrated in part B of the figure, depends on the spacing between the holes. Some energy leaks directly through the hole at WX. Addi-

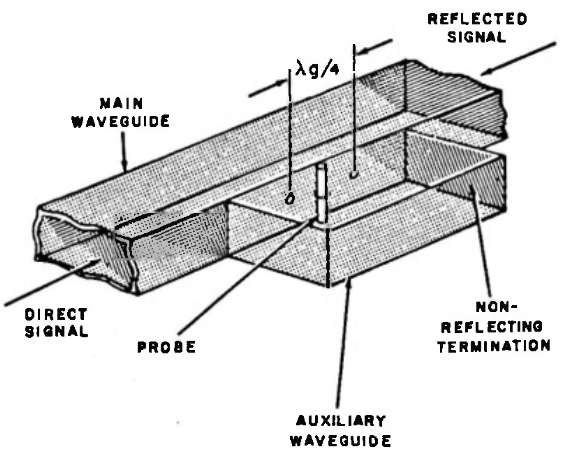


Figure 13-181. Directional Coupler

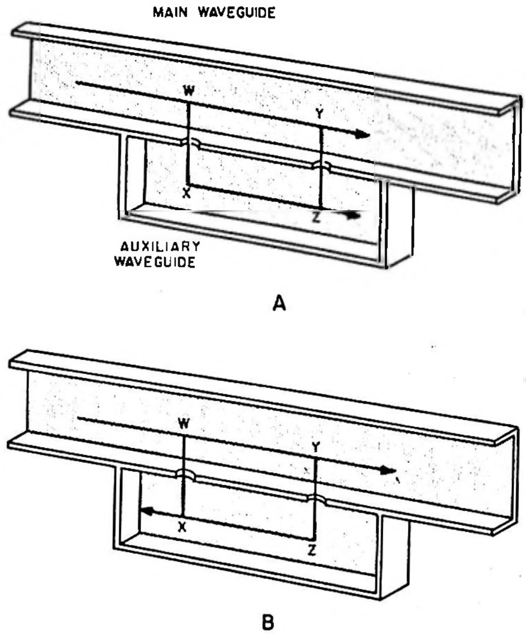


Figure 13-182. Theory of Directional Coupler

tional energy travels from W to Y, leaks through the hole at YZ, and then travels back from Z to X. The longer path differs from the shorter path by twice the spacing between the holes. If, then, the holes are separated by a distance $\lambda_g/4$, the two signals will be out of phase at X and no energy will be propagated from right to left in the auxiliary guide. Energy from the direct signal leaks into the directional coupler and travels in the same direction toward the right-hand end, where it is absorbed by a nonreflecting resistive surface termination, the impedance of which matches that of the waveguide (see figure 13-181). Energy from the reflected signal leaks into the directional coupler and continues to the left-hand end, where it excites the probe. Any reflection from the left-hand termination of the directional coupler will eventually reach the right-hand termination and be absorbed. The probe is fitted with a crystal holder. The signal picked up by the probe is proportional to the strength of the reflected signal in the main waveguide, and does not depend on the level of the signal traveling in the other direction. A waveguide arrangement is frequently fitted with two directional couplers, one to measure the strength of the desired signal and the other to measure the level of the reflected signal. It is also possible to feed energy into the directional coupler through the probe. Some of this energy will be absorbed by the resistive termination, and the rest will leak through the two holes into the main waveguide. Because of the difference in the path lengths through the two holes, the energy that reaches the main waveguide will be propagated in only one direction.

13-713. ATTENUATORS.

13-714. It is sometimes necessary for you to adjust the power level in a waveguide, and this is accomplished by means of an attenuator. Any of the various tuning devices will serve as an attenuator, since it can be deliberately adjusted for an impedance mismatch. The tuning device produces a reflected signal; only a portion of the incident signal in the waveguide will be transmitted, since the remainder is reflected. A more convenient device, however, is the shutter attenuator, shown in part A of figure 13-183. Part B of the figure shows how the conducting shutter closes off most of the waveguide so that only a small amount of energy can be passed. In many cases, it is desirable to attenuate the signal without introducing any reflection; you can accomplish this by using the dissipative attenuator, shown in part A of figure 13-184. This structure consists of a holder which carries a card coated with a resistive material. This card is inserted in the waveguide, as shown in part B, parallel to the electric field. The attenuator is designed to absorb and dissipate energy. If the resistance of the card is of the proper value, the dissipative attenuator will not introduce reflection. The resistance of a conducting surface of this kind is specified in ohms per square unit, without any restriction

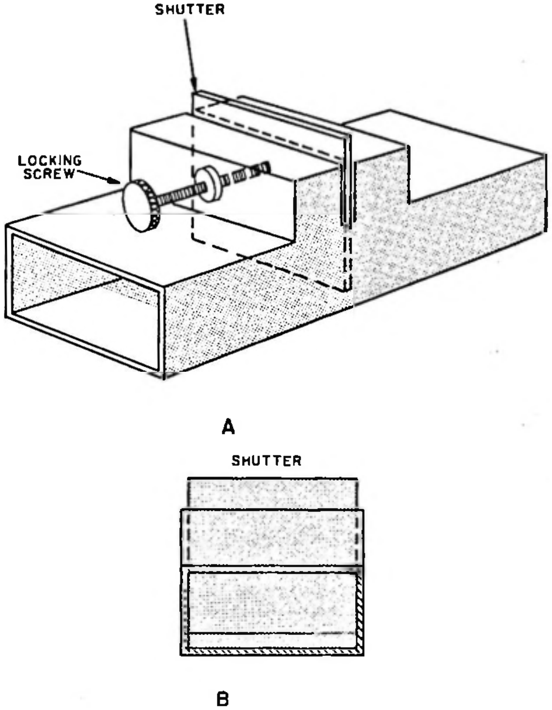


Figure 13-183. Shutter Attenuator

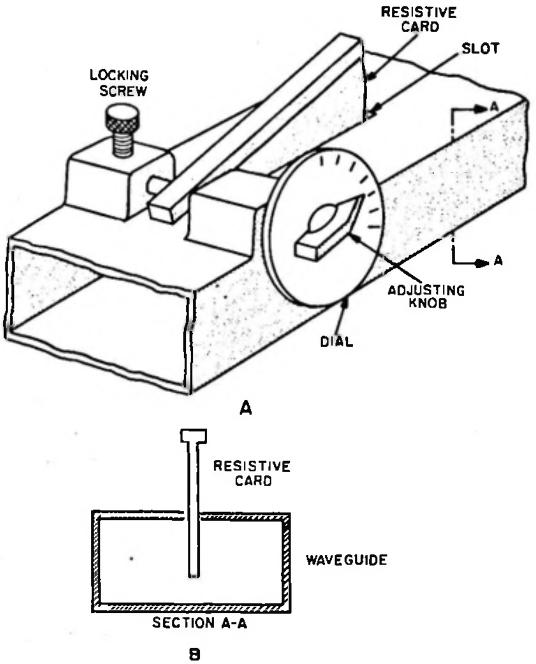


Figure 13-184. Dissipative Attenuator

as to how many units of distance are involved. In figure 13-185, part A shows a square of conducting surface which may have any convenient dimensions (1 inch by 1 inch, for example). Two conducting bars are fastened to opposite sides of the square, and the resistance between these is specified as R. In part B, the rectangle is specified as 1 inch by 2 inches, with a resistance of 2R. However, two of these rectangles in parallel form the larger square, 2 inches by 2 inches, as shown in part C; this square has a resistance of R. It follows that the resistance of any unit square will be the same as the resistance of any other unit square, and the behavior of the material is specified completely if the resistance per square unit is known. The impedance of a waveguide is specified at some plane perpendicular to the waveguide axis. The waveguide is, therefore, like a conducting surface, and its impedance is given in ohms per square unit.

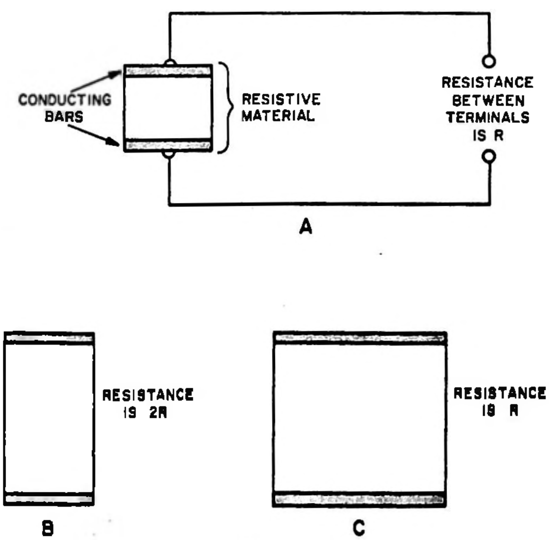


Figure 13-185. Resistance of Conductive Surface

13-715. TRANSMITTER COUPLING TO TRANSMISSION EQUIPMENT.

13-716. A microwave transmitter consists of an oscillator (together with its power supplies, modulator, and control circuits) operating at the desired output power level. Because of the difficulty of making a power amplifier for use at microwave frequencies, a power oscillator is used. (Power amplifiers are normally employed at lower frequencies.) The oscillator is normally a multicavity magnetron, although a reflex klystron is used frequently in low-power applications. In either case, microwave power is developed in a resonant cavity, or cavity resonator, which is an integral part of the tube structure. Energy is taken from the cavity on a short length of coaxial line, which is also a part of the tube structure, and delivered to the transmission equipment, which may be either a coaxial line or a waveguide.

13-717. MAGNETRON COUPLING.

13-718. In many S-band magnetrons, the coaxial line from the resonant cavity is terminated in fittings that allow it to be joined directly to a standard S-band coaxial transmission line. Figure 13-186 illustrates such a magnetron. In magnetrons used at the higher frequencies, and sometimes in S-band magnetrons, the coaxial line from the resonant cavity is coupled to a short section of waveguide that is also a part of the magnetron structure. This short section is connected to the waveguide transmission arrangement which carries the output of the magnetron. Although either a rectangular or circular waveguide may be used for the output fitting, most magnetrons use a rectangular waveguide. Figure 13-187 shows a magnetron of this type.

13-719. IMPEDANCE MATCHING BETWEEN MAGNETRON AND LINE.

13-720. A magnetron is designed to have an output impedance equal to the characteristic impedance of the transmission line with which it is used. (The term transmission line is used here means either a coaxial line or a hollow waveguide.) Therefore, the magnetron will operate with reasonable efficiency when it is connected directly to a transmission line of the it is designed to

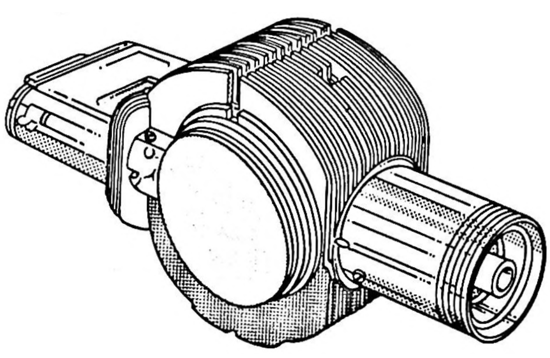


Figure 13-186. S-Band Magnetron with Coaxial Line Output Fittings

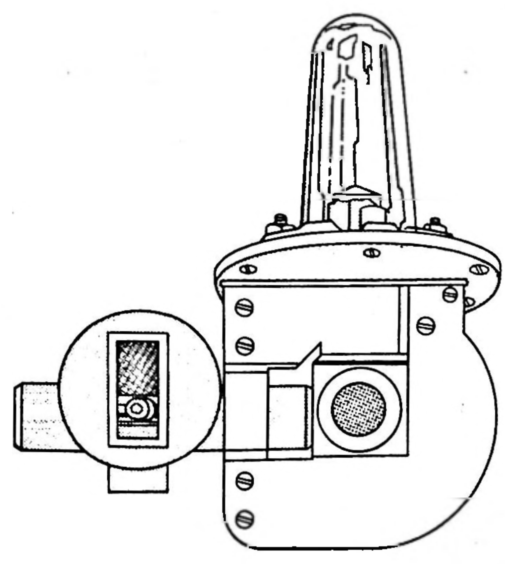


Figure 13-187. Magnetron with Waveguide Output Fitting

feed. For highest efficiency, however, it is usually necessary to insert a tuning device between the magnetron and the transmission line. The tuning device is either a stub tuner or a slug tuner, depending on whether the transmission line is a coaxial line or a waveguide. As stated in the discussion on the theory of tuning devices of this type, reflected signals are introduced into the coaxial line or waveguide and the input impedance of the tuner is made to take any desired value. When you adjust the tuner properly, the magnetron will deliver maximum power to the transmission line.

13-721. TRANSMISSION LINE TUNING.

13-722. GENERAL. A complete transmission equipment, coupling a magnetron to a load, is shown in figure 13-188. Tuner X matches the transmission line to the magnetron, and tuner Y matches the load to the line. You can measure the vswr at the slotted section (normally used with a coaxial-line arrangement), while you adjust tuner Y until the vswr is as close as possible to unity. With this adjustment, there is no reflected signal to the left of tuner Y, and the load is matched to the transmission line. The line to the left of tuner Y is said to be "flat" when there is no reflected signal. Following adjustment of tuner Y, tuner X is adjusted so that the magnetron delivers maximum power to the line, the signal level being measured at the slotted section. With a waveguide transmission arrangement, the slotted section is generally replaced by two directional couplers, although the slotted section can be used with a waveguide. One directional coupler reads the reflected signal, and tuner Y is adjusted so that the reflected signal has the smallest possible output. The other directional coupler reads the direct signal from the magnetron, and tuner X is adjusted so that the direct signal has the greatest possible output; tuner Y is always adjusted first.

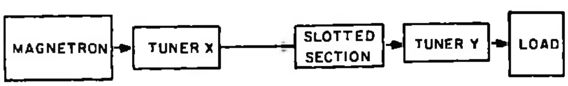


Figure 13-188. Transmission Line with Tuners

13-723. PREPLUMBED ARRANGEMENTS. In many cases, the load and the magnetron are designed to match the transmission line, and both tuners are omitted. More frequently, however, the tuner next to the magnetron is omitted. The portion of the equipment from which the tuner is omitted is said to be "preplumbed." In general, preplumbed equipments are used when maintenance is difficult to carry out. Although an equipment with tuners is slightly superior to a preplumbed equipment when the tuners are adjusted properly, it is inferior when the tuners are adjusted improperly. Because of this, the tuners are omitted whenever there is a likelihood that they will get out of adjustment between maintenance checks.

13-724. KLYSTRON TO LINE COUPLING.

13-725. The output of an S-band klystron normally appears at a coaxial fitting. A conventional polyethylene dielectric flexible coaxial cable (which is kept as short as possible) is used to carry the signal to a rigid S-band coaxial line. When a klystron of this kind is used to feed a waveguide, the coaxial cable is connected to a probe. Some S-band klystrons, and almost all klystrons operating at higher frequencies, use a short length of rigid, small-diameter, polystyrene dielectric coaxial line. When used as a transmitter, this type of klystron is normally coupled to a rectangular waveguide. Figure 13-189 illustrates this method of coupling for an X-band klystron. Although the stub shown on the left side of the illustration may be preplumbed in a transmitter, it is usually adjustable.

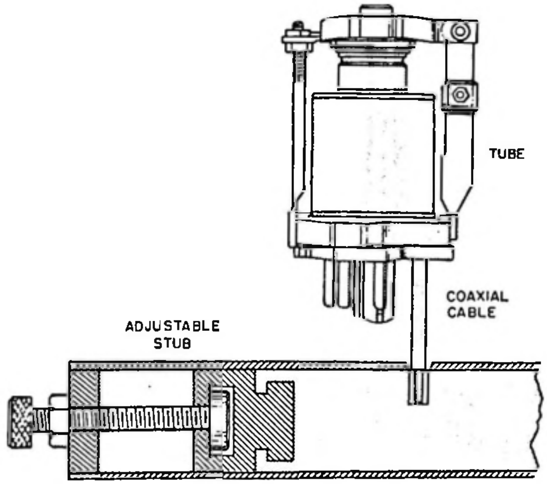


Figure 13-189. X-Band Klystron Coupled to Waveguide

13-726. RECEIVER COUPLING TO TRANSMISSION EQUIPMENT.

13-727. COUPLING FROM COAXIAL LINE.

13-728. Figure 13-190 shows a coaxial line receiver coupling assembly. The line is terminated at the crystal mixer on the left, and the i-f output is taken out through a conventional flexible coaxial cable. The local oscillator signal is brought to the assembly through a second flexible coaxial cable and coupled to the mixer through the capacitance between the local oscillator probe and the center conductor of the rigid line. The probe is sometimes adjustable, but is usually preplumbed. The crystal mixer impedance is normally not equal to the characteristic impedance of the rigid coaxial line, and a two-stub tuner is inserted between the receiver coupling assembly and the transmission equipment that feeds it.

13-729. COUPLING FROM WAVEGUIDE.

13-730. Figure 13-191 shows a waveguide receiver coupling assembly. This assembly is essentially the same as the coaxial-line assembly except that an adjustable stub is used to terminate the waveguide. The coupling is adjusted by first finding the best position for the adjustable stub and then setting the tuner so that the complete coupling assembly is matched to the waveguide that feeds it. The local oscillator probe is some-

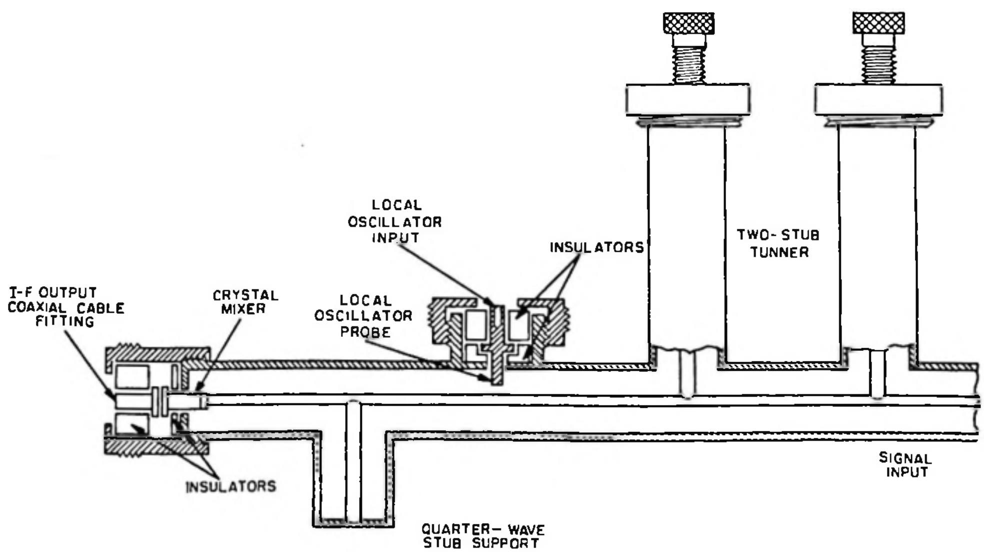


Figure 13-190. Coaxial Line Receiver Coupling Assembly

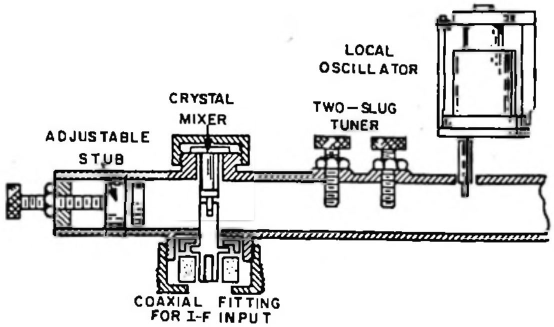


Figure 13-191. Waveguide Receiver Coupling Assembly

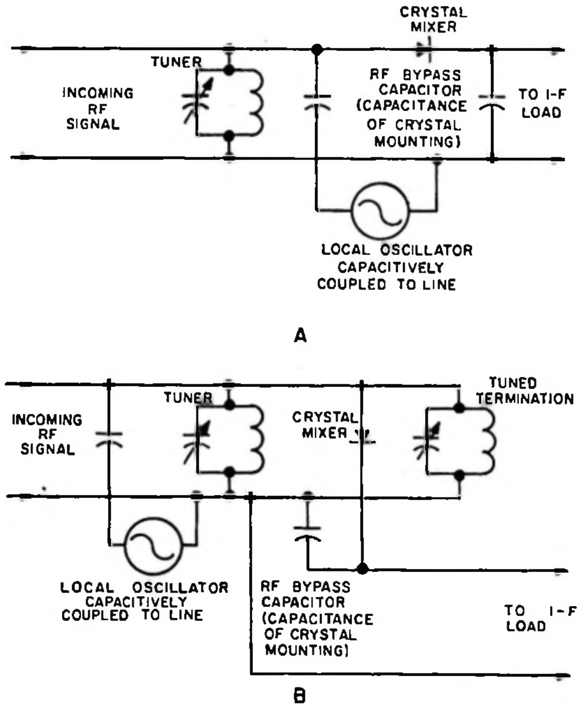


Figure 13-192. Equivalent Circuits of Receiver Coupling Assembly

times adjustable, but is usually preplumbed. The local oscillator signal is injected ahead of the tuner so that both the incoming signal and the local oscillator signal are changed in the same way by adjustment of the stub and tuner. Parts A and B of figure 13-192 show, respectively the equivalent circuits of the coaxial-line and waveguide receiver coupling circuits.

13-731. LOW-SENSITIVITY COUPLING.

13-732. In some cases, you may wish to measure the level of a fairly strong signal on a microwave transmission line without drawing an appreciable amount of power. This problem arises, for example, with a slotted section. The physical construction of a coupling which is suitable for this purpose is shown in part A of figure 13-193, and the equivalent circuit is shown in part B of the figure. No tuner is necessary, since the probe projects only a short distance into the interior of the coaxial line or waveguide Probe coupling of this type is used with de-

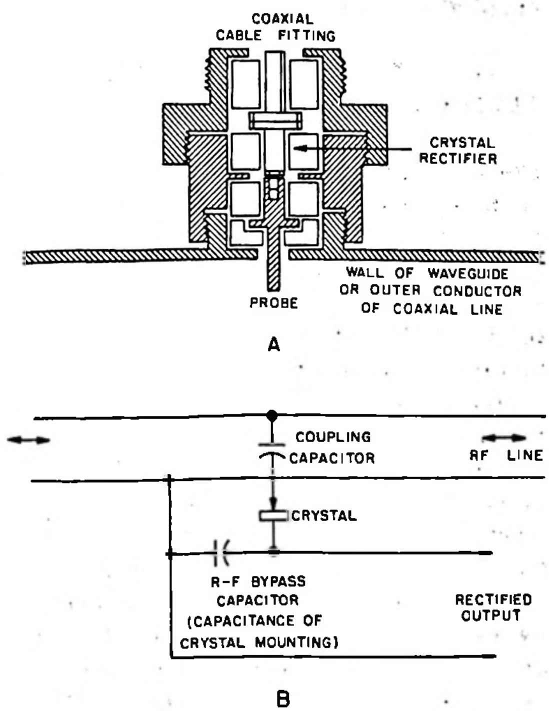


Figure 13-193. Low-Sensitivity Coupling and Equivalent Circuit

vices measuring the power level directly at the transmission line. The output from a directional coupler, however, is taken with the more sensitive arrangement for the crystal (figure 13-191), since the coupler itself is only loosely coupled to the transmission line. (The tuner and the adjustable stub are not used with a directional coupler).

13-733. COAXIAL LINE COUPLING TO WAVEGUIDE.

13-734. Coupling between the miniature coaxial line of an X-band reflex klystron and a rectangular waveguide, where the klystron functions as a transmitter, has been discussed; a similar coupling is used where the klystron functions as the local oscillator of a receiver (as shown in figure 13-191). Coupling a normal coaxial line to a waveguide is accomplished as described below. The use of proper fittings will allow either rigid or flexible coaxial line to be coupled to a waveguide in the same way.

13-735. COAXIAL LINE COUPLING TO CIRCULAR WAVEGUIDE.

13-736. A coupling of this type is seldom seen because there is little requirement for it. Part A of figure 13-194 shows a coupling that has the advantage of being simple, but is not very efficient. The coaxial line is terminated at the end wall of the waveguide, and the center conductor is extended to form a probe. A more efficient coupling—one in which there is a better impedance match—is shown in part B of the figure. This is rather bulky, since the tapered section is usually several guide wavelengths. The coupling shown in part B can be improved by the addition of a stub tuner, provided that the tuner is adjusted correctly. If there is any possibility of frequency variation under circumstances in which the tuner will not be readjusted immediately, it is preferable to omit the tuner entirely. Both couplings are designed for use with a circular waveguide operating in the $TM_{0,1}$ mode.

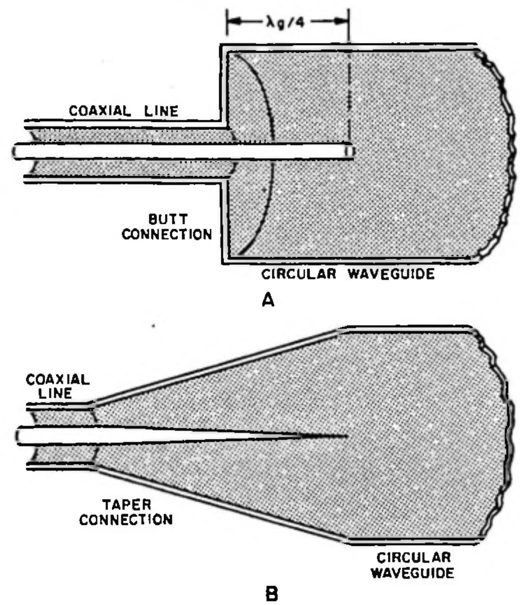


Figure 13-194. Coupling Between Coaxial Line and Circular Waveguide

13-737. COAXIAL LINE COUPLING TO RECTANGULAR WAVEGUIDE.

13-738. Part A of figure 13-195 shows a simple coupling between coaxial line and a rectangular waveguide. This coupling is reasonably efficient, and can be improved by use of either a waveguide or a coaxial line tuner. A more efficient coupling which employs a preplumbed tuner, shown in part B, is normally used with a magnetron fitted with a waveguide output fitting. It is possible to eliminate almost all reflection from this coupling by making both the waveguide stub and the coaxial stub adjustable, provided that the characteristic impedances of the waveguide and the coaxial line are reasonably similar. If the impedances differ widely, it is sometimes necessary to use a conventional tuner; in this case the simpler coupling shown in part A is used instead. Figure 13-196 shows a coupling that is used to transfer a small portion of the energy in the waveguide to a coaxial

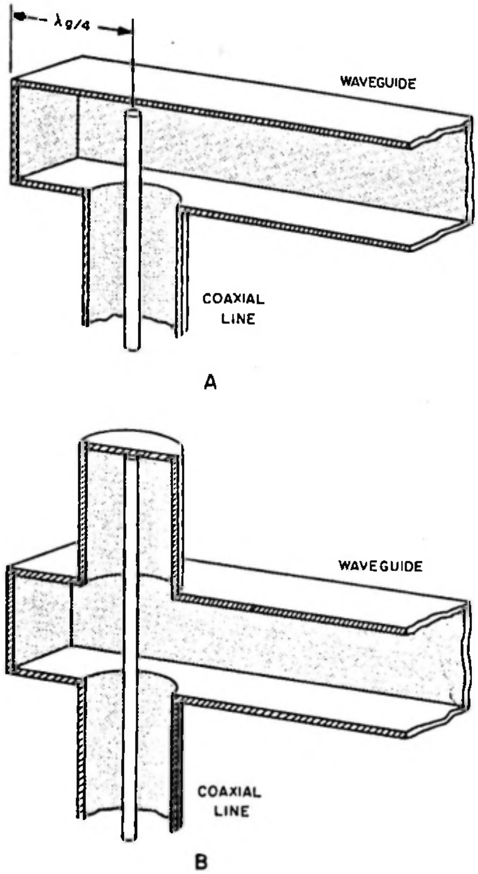


Figure 13-195. Coupling Between Coaxial Line and Rectangular Waveguide

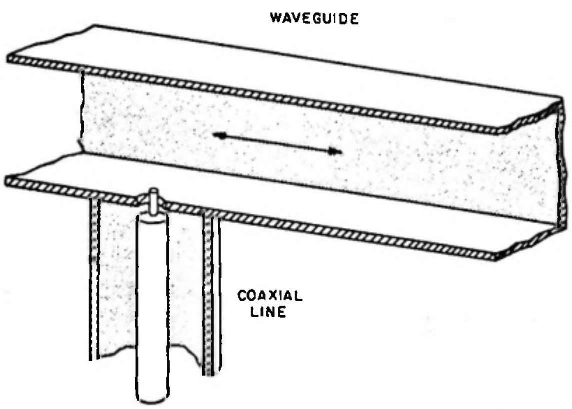


Figure 13-196. Loose Coupling Between Coaxial Line and Rectangular Waveguide

line, or to add a small amount of additional energy to that already present. You can use such a coupling, for example, to feed the signal from a local oscillator into a waveguide connected to the crystal mixer of a microwave receiver. It is also used to extract a small amount of energy from the output of a transmitter to feed a wavemeter. The probe extends only a short distance into the waveguide, and has little effect on its operation. There is a very poor impedance match, however, at the end of the coaxial line, since the line is terminated by a load that is almost a short circuit. The coaxial line is not tuned because the efficiency is rarely important in applications where this type of coupling is used.

13-739. RECTANGULAR TO CIRCULAR WAVEGUIDE COUPLING.

13-740. It is frequently necessary to couple rectangular and circular waveguides together, as shown in figure 13-197, in order to make use of the rotating joint mentioned previously. This type of coupling is frequently preplumbed, but if some adjustment is desired for tuning, the probe is made in the form of a screw so that its length can be varied. For precise matching, the probe is made so that its length is adjustable and the fixed termination shown in the right-hand side of

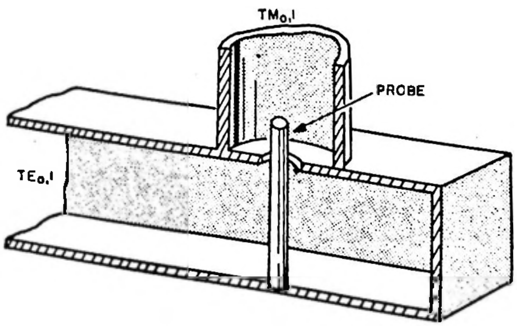


Figure 13-197. Coupling Between Rectangular and Circular Waveguides

the figure is replaced by an adjustable stub, but this refinement is ordinarily not required.

13-741. JOINTS BETWEEN SECTIONS OF TRANSMISSION EQUIPMENT.

13-742. ROTATING JOINTS.

13-743. The rotating joint used with a circular waveguide (see figure 13-198) consists of two circular plates, one carried on each of the rotating waveguide sections. One of the plates, called a choke plate, is grooved. The other, called a flange, is simply a flat disk. The groove is essentially a shorted section of coaxial line, and its depth is chosen so that its electrical length is a quarter-wavelength. The impedance at point X, looking out from the waveguide, is the impedance of the groove in series with the impedance of the gap between the choke plate and the flange. Since the groove is a quarter-wave line terminated in a short, its impedance is infinite. The total impedance at point X is then always infinite for reasonable separation between the choke plate and the flange. Since the mouth of the groove is a quarter-wavelength from the inner wall of the waveguide, the impedance at point Y is that of a quarter-wave section terminated by an open circuit, and this is zero. The waveguide, therefore, appears to have continuous walls as long as you do not separate the two plates by more than a quarter-wavelength. This is not affected by any electrical connection at the circumference of the plates; consequently, it is possible to seal the joint if the waveguide is to be pressurized. A poor electrical contact at the seal, or even an intermittent one, will not affect the operation of the joint. This method is similar to the one used at the outer conductor of the coaxial line rotating joint. The same method, using a choke groove, is used to insure good electrical termination in an adjustable waveguide stub.

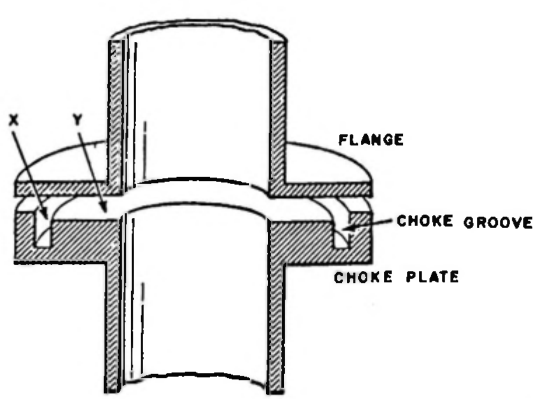


Figure 13-198. Waveguide Rotating Joint

13-744. FIXED JOINTS.

13-745. GENERAL. A microwave transmission system is normally made up of a number of sections that can be separated for inspection, cleaning, and possible replacement. Standard joint fittings are used to connect adjacent sections.

13-746. COAXIAL LINE. Details of the standard S-band coaxial-line joint are shown assembled in part C of figure 13-199, with male and female fittings illustrated in parts A and B, respectively. A clip fastening device holds the two parts of the joint together. Short sections of coaxial line have single quarter-wave-stub supports to hold the inner conductor in its proper position, and longer sections have a stub support at each end.

13-747. WAVEGUIDE. Two sections of waveguide may be joined by providing each one with a flange fitting and securing the joint with several bolts. This is normally not done, however, because undesired reflections are developed at the joint if it is misaligned (see figure 13-200). To minimize the possibility of reflections, a waveguide joint is usually constructed as shown in figure 13-198, where one section is fitted with a choke plate and the other is fitted with a flange. This is called a choke joint; it is preferable to the simple joint in figure

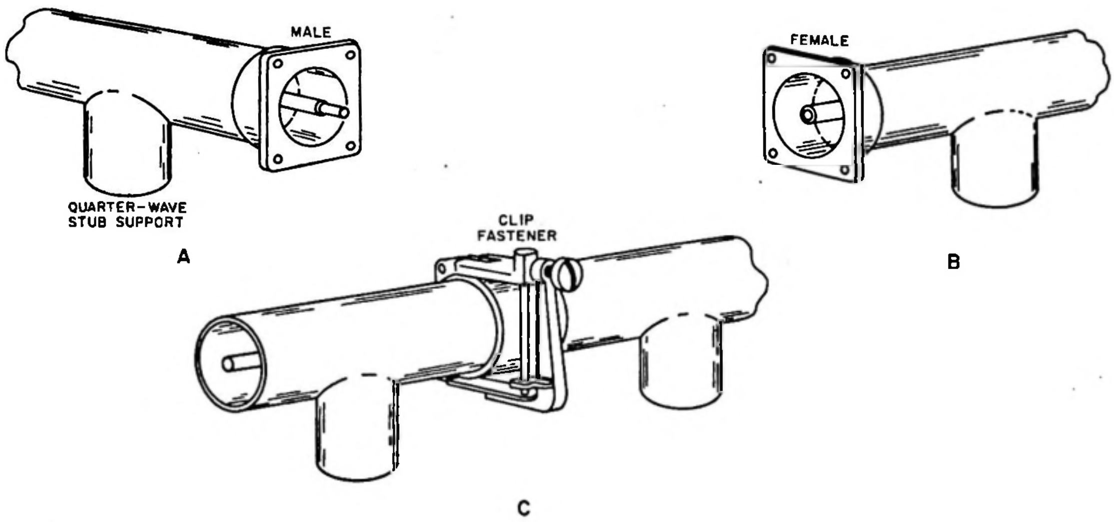


Figure 13-199. Joint for Coaxial Line

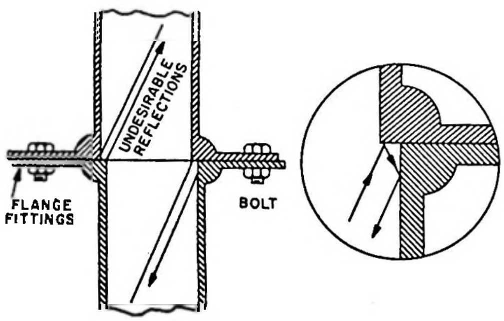


Figure 13-200. Effect of Misalignment in Waveguide Joint

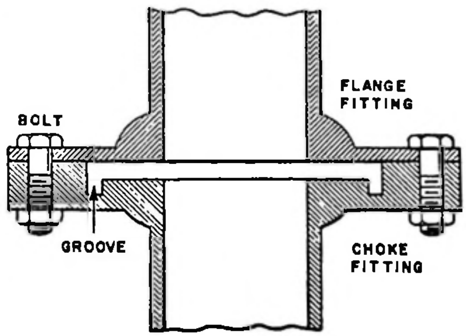


Figure 13-201. Choke Joint

13-200 because the walls of the waveguide are electrically smooth, even though they do not touch each other. Figure 13-201 shows a cross section of a choke joint. The theory of the choke joint is precisely the same as that of the rotating joint. The waveguide walls are electrically continuous at the joint, and the electrical distance between the groove and the inner surface of the waveguide wall is a quarter-wavelength. Part A of figure 13-202 shows a choke fitting for use with a circular waveguide. A choke fitting for use with a rectangular waveguide is shown in part B. The distance from the waveguide wall to the groove is, of course, different at different parts of the joint. In part B of figure 13-202, the distance X to Y, where the electrical field is strongest, is made electrically equal to a quarter-wavelength, and this is usually satisfactory. Part A of figure 13-203 shows an X-band rectangular waveguide with a choke fitting;

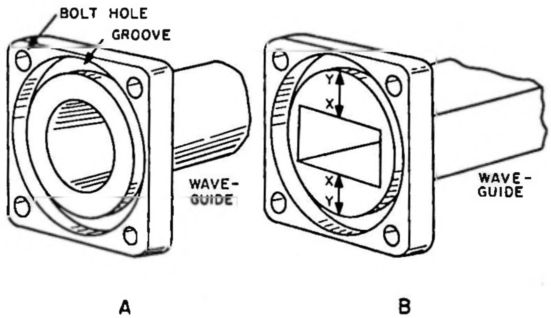


Figure 13-202. Choke Fittings

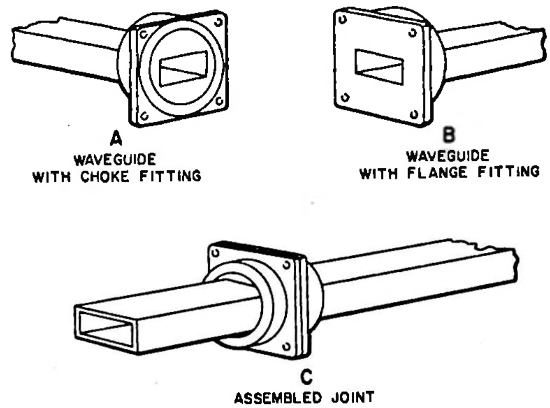


Figure 13-203. Rectangular Waveguide Joint

part B is the adjacent section with a flange fitting, and part C shows the assembled joint with the bolts in place.

13-748. WOBBLE JOINT.

13-749. You may desire to secure a waveguide transmission line to the structure of a ship or an aircraft while the unit which feeds the waveguide (a receiver, a transmitter, or a combination of both) is shock-mounted. The rf unit then will be free to move slightly in relation to the waveguide. It is possible to make a connection to the waveguide by using a flexible section, but the usual method is to use a wobble joint. A wobble joint is nothing more than a conventional choke joint in which the choke and flange fittings are separated by a distance of approximately 1/16 inch. It is possible to move one section of the wobble joint with respect to the other for a considerable distance without causing any ill effects, because the presence of the choke groove creates a smooth-walled joint having electrical continuity. In a typical wobble joint, a shift of 1/32 inch in any direction and a rotation of 2 degrees about any axis are permissible. It is customary to use a wobble joint with a rectangular waveguide. If a circular waveguide is used instead, the joint is electrically identical with a rotating joint, and either section may rotate through any angle about its electrical axis. Various positions of a wobble joint are shown, with some exaggeration, in figure 13-204.

13-750. BENDS, TWISTS, AND FLEXIBLE SECTIONS.

13-751. Special sections of coaxial line or waveguide are used when it is necessary to run transmission equipment around corners. These are usually bends of various kinds; in a rectangular waveguide, they are twisted sections. Representative types of bends, twists, and flexible sections are described in the following text.

13-752. COAXIAL LINE BENDS.

13-753. Part A of figure 13-205 shows a smooth bend in a coaxial line, and part B shows a right-angle bend. The quarter-wav stub supports, previously described, can be clearly seen. A smooth bend should be used in preference to a right-angle bend wheneve you have a choice, since it is less likely to produce reflected signals.

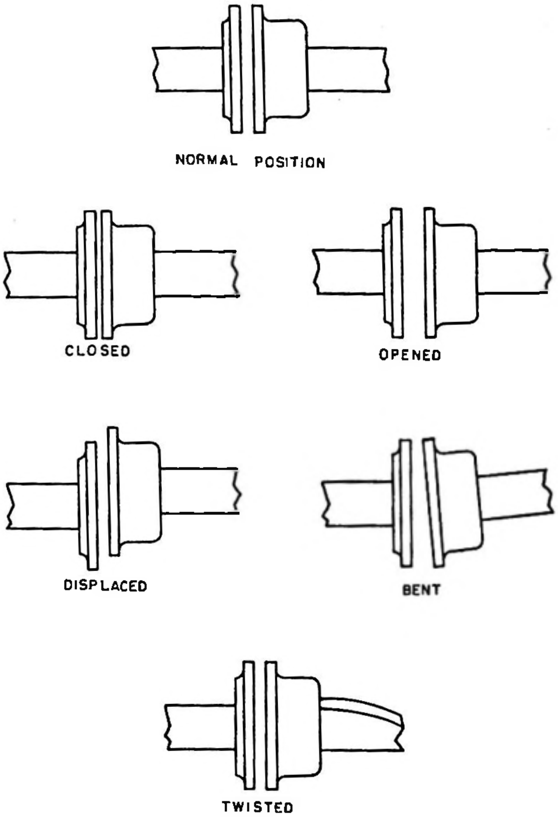


Figure 13-204. Various Positions of Wobble Joint in Rectangular Waveguide

13-754. WAVEGUIDE BENDS.

13-755. Bends in a rectangular waveguide may be classified as E-bends (in which the waveguide is bent up or down, in the plane of the electric field) and H-bends (in which the waveguide is bent sideways, in the plane of the magnetic field). Part A of figure 13-206 shows a smooth E-bend in an X-band rectangular waveguide, and part B shows a smooth H-bend; part C shows a smooth bend in a circular waveguide. The operation of bends of this type depends on the fact that any signal in the waveguide excites all the modes that are not suppressed. As long as the bend is smooth and gradual, very little reflection is produced. When it is necessary for the waveguide to bend sharply, a mitered bend is used. Part A of figure 13-207 shows a mitered E-bend, and part B shows a mitered H-bend. Mitered bends are rarely used with a circular waveguide. The operation of a mitered bend depends on reflection from the 45 degree wall; this inevitably sends some of the signal back along the waveguide, as shown in figure 13-208. A smooth bend should be used instead of a mitered bend whenever you have a choice.

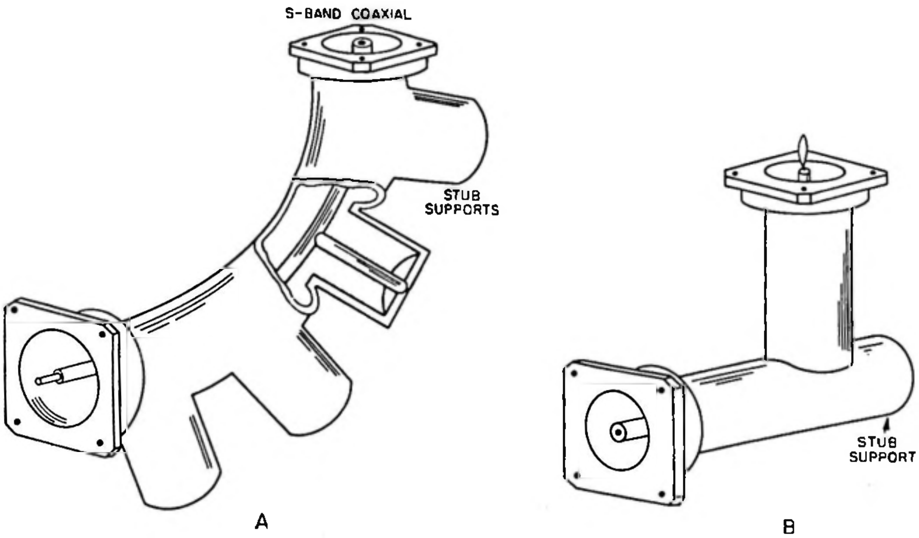


Figure 13-205. Coaxial Line Bends

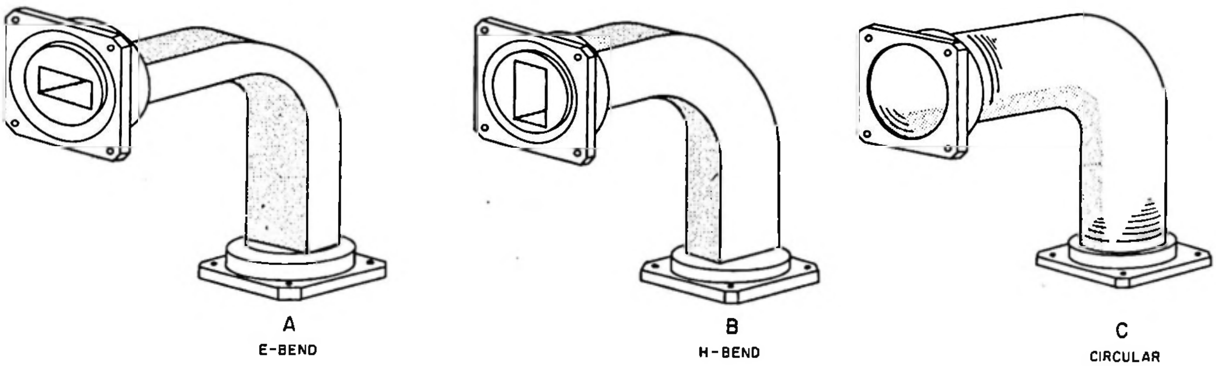


Figure 13-206. Smooth Waveguide Bends

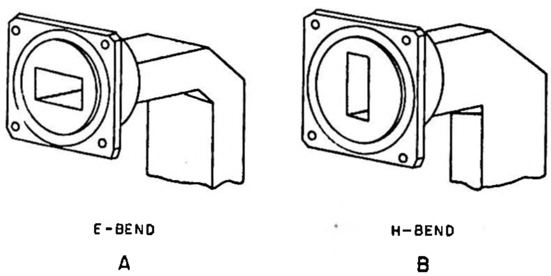


Figure 13-207. Mitered Waveguide Bends

13-756. TWISTS.

13-757. A twisted section of coaxial line can be used to change the plane of polarization in a rectangular waveguide. Figure 13-209 shows a twisted section of X-band rectangular waveguide, and figure 13-210 shows an installation using twisted sections.

13-758. FLEXIBLE SECTIONS.

13-759. Flexible coaxial line is generally familiar to everyone with experience in the field of communications. Its use at microwave frequencies is avoided wherever possible because it has much higher losses than either a waveguide or a rigid coaxial line. Flexible waveguide is available, but its use is limited for the reason stated—high losses. One type of flexible waveguide uses a woven shield braid conductor. This is formed in a circular or rectangular tube and surrounded by a molded rubber sheath of the same shape.

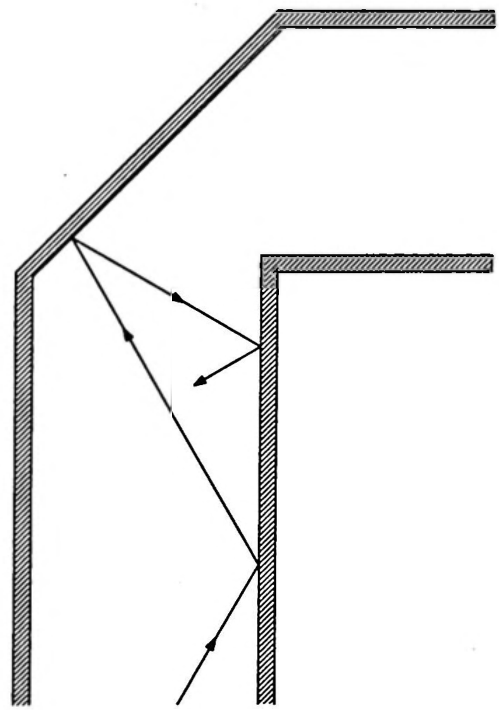


Figure 13-208. Reflected Signal from Mitered Bend

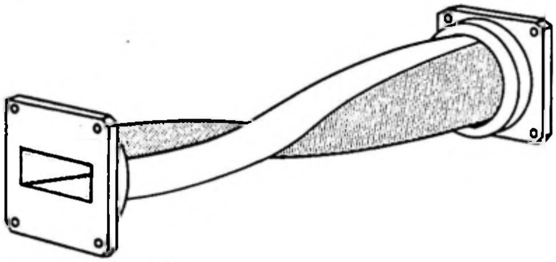


Figure 13-209. X-Band Waveguide Twist

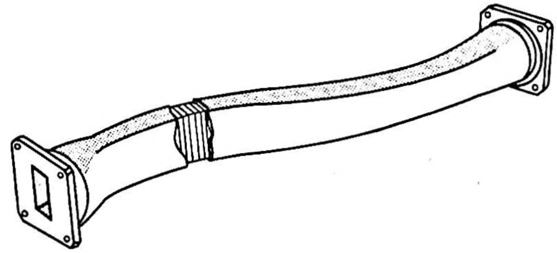


Figure 13-211. Flexible X-Band Waveguide

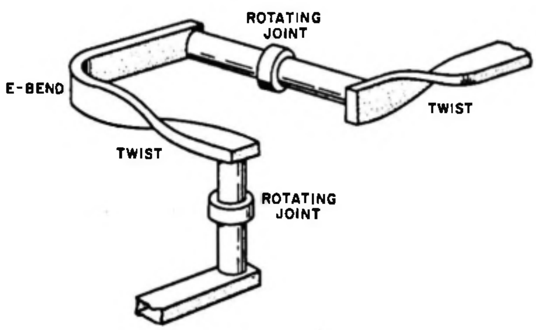


Figure 13-210. Waveguide Installation with Twists

The rubber sheath serves to hold the flexible conductor in the correct shape and main tains the desired circular or rectangular form of the waveguide. Another type consists of a spiral of flat metal ribbon, each turn being in electrical contact with its neighbors. As before, a rubber sheath surrounds the conductor and holds the turns in the proper relation to one another. Still another type of flexible waveguide makes use of a number of choke fittings. These are held in a rubber sheath and form a very flexible guide that is free from troubles resulting from poor electrical contact. In general, the use of a flexible waveguide is avoided when a rotating joint or a wobble joint can be used instead. One type of flexible waveguide is illustrated in figure 13-211.

13-760. JUNCTIONS.

13-761. TEE.

13-762. Tee junctions in a coaxial line have previously been discussed in connection with the stub tuner. The quarter-wave stub support is another example of a tee junction. Two types of tee junctions are used with a rectangular waveguide, as discussed below. Tee junctions are seldom used with a circular waveguide.

13-763. MAGIC TEE.

13-764. In addition to the junctions already described, there is an interesting double junction called a magic tee (see figure 13-212). The magic tee consists of a length of rectangular waveguide with an E-junction and an H-junction added at the same point. If a signal is fed in on branch A or B, there will be some output from each of the other branches of the tee. If, however, a signal is fed in on Branch C or D, there will be output only from branches A and B. Part A of figure 13-213 shows what happens when there is an input at branch C. The electric field is in opposite directions in the right and left portions of branch D, and attempts to excite either a TM or a $TE_{2,n}$ mode. Both of these are suppressed by your choice of waveguide dimensions, and no energy is propagated down the branch. Part B illustrates what happens when there is an input

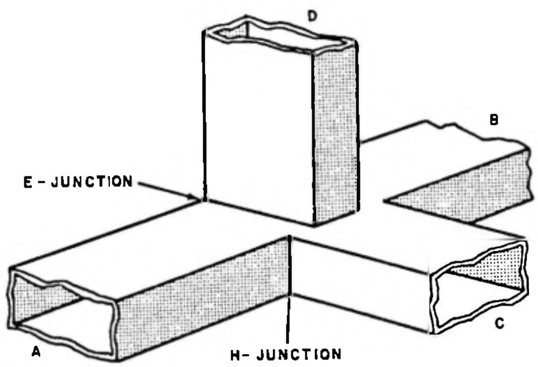


Figure 13-212. Magic Tee

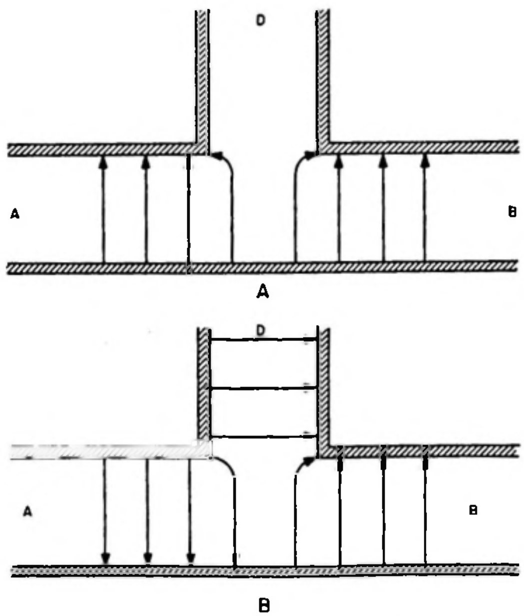


Figure 13-213. Operation of Magic Tee

at branch D. Here, the electric field attempts to excite the $TE_{0,2}$ mode in branch C, but this is suppressed as explained before. The junction will not behave in precisely the desired way unless branches A and B are flat. If any signal is reflected back in either of these branches, it will feed all of the other branches and there will be some coupling between branches C and D. A magic tee can be used in place of a directional coupler, and it is somewhat more sensitive in detecting reflected signals. The direct signal is fed in at branch C and out to the remainder of the transmission equipment from A. Branch B is terminated properly, and a detector is connected to branch D. So long as the transmission arrangement is flat (and branch A is terminated correctly), there will be no signal at D. If, however, there is any reflected signal, a portion of it will reach the detector through branch D.

13-765. HORN TERMINATION.

13-766. A transmission equipment, whether coaxial line or waveguide, usually runs from a transmitter to an antenna or from an antenna to a receiver. Any consideration of transmission equipments must, therefore, include some discussion of the devices used for coupling between the antenna and the transmission line. This discussion will be deferred until Chapter 14, which deals with microwave antennas. For the moment it is sufficient to note that there is always the problem of impedance matching at the antenna coupling. You can solve this problem by the use of a preplumbed tapered section, and you can add a tuner for more precise matching. One example of a tapered matching unit is the horn termination shown in figure 13-214.

13-767. TYPES OF WAVEGUIDES.

13-768. DIELECTRIC WAVEGUIDE.

13-769. In addition to the air-filled rectangular and circular waveguides previously discussed, some waveguides are filled with a solid dielectric. This increases the losses tremendously, but also raises the power-handling capacity by permitting the use of stronger electric fields because the breakdown voltage (the voltage at which an arc is formed) is higher for a good solid dielectric

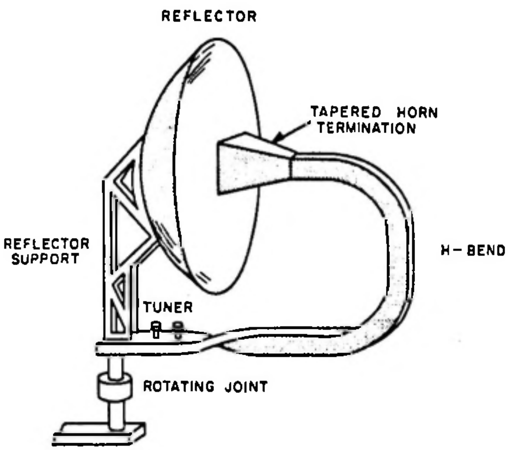


Figure 13-214. Tapered Horn Termination

than for air. This is the same principle that permits an oil-filled capacitor to operate at a higher voltage than an air capacitor. The behavior of a dielectric-filled waveguide is the same as that of a conventional air-filled waveguide, except that, in all calculations, λ_a must be replaced by λ_d (the wavelength of the signals in the dielectric) and λ_d is equal to $\lambda_a\sqrt{k}$, where k is the dielectric constant of the material. Another application of a solid dielectric is in the dielectric waveguide, which is nothing more than a solid rod of dielectric. This acts as a waveguide because signals traveling in the dielectrig are reflected at the boundary. Reflection takes place because of the difference between the impedance of the dielectric and that of the surrounding air; the reflection at an impedance discontinuity is similar to the reflection at an impedance mismatch on a transmission line. The dielectric waveguide seldom is used. Not only does it have high losses; it also radiates to some extent because the reflection at the boundary is not complete.

13-770. SURFACE WAVEGUIDE.

13-771. GENERAL. A single conductor used as a guiding surface for microwave frequencies is called a surface waveguide or open waveguide, or, more briefly, a G-line, from the initials of its originator, G. Goubau. It is described here to emphasize the fact that a single-conductor guiding structure need not be a hollow waveguide.

13-772. The open waveguide that has been most thoroughly investigated is the conventional two-wire line. This type of line is useful at frequencies up to and including the vhf range. The small spacing required between the wires at higher frequencies and the difficulty of maintaining perfect symmetry to prevent excessive radiation make the two-wire line less useful at higher frequencies. The single-conductor line used as an open waveguide is of particular interest, since it appears to be simple in construction and is most useful in the microwave ranges. Open waveguides have two advantages: they are less bulky and less expensive, and, since the field is not confined to a small area, as in a coaxial line or closed waveguide, the field density is low. This, in turn, has the effect that, under comparable conditions, the resistive and dielectric losses of open guides are much smaller than those of closed guides. The only practical wave mode that is guided by a single conductor is a radially symmetrical transverse magnetic mode. The radial extensive of the field, caused by reducing the phase velocity, can be controlled by modifying the conductor surface. The simplest modification of the conductor surface is the application of a dielectric layer such as ordinary enamel or polyethylene. A very thin layer of dielectric will reduce the extension of the field considerably. Since the part of the energy that is propagated within the dielectric layer is small, the loss resulting from the dielectric material is almost negligible. A practical method of launching the surface wave on a single conductor becomes evident if the field of this wave (see figure 13-215) is compared with the field of a wave in a coaxial line. Starting with a coaxial-line section, with the inner conductor

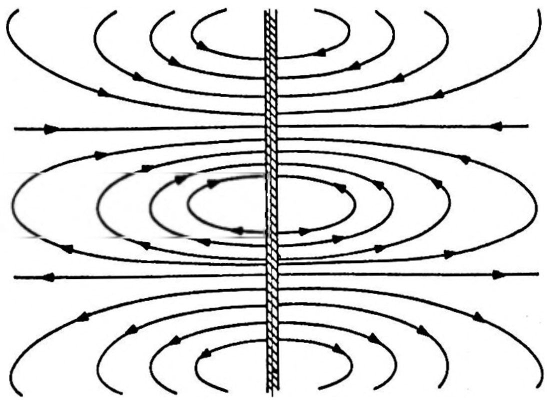


Figure 13-215. Single-Conductor Open-Guide Field Pattern

having a dielectric coating, or layer, and increasing the outer conductor of the coaxial section gradually until it is so large that it no longer affects the field very much, the field distribution becomes that of the open waveguide. The field lines in part A of figure 13-216 give a rough idea of how the surface wave develops. The complete setup for a surface wave transmission line is shown in part B. The single-conductor surface-wave line is useful throughout the microwave frequency range. The line must be kept straight and clear of field obstructions as determined by the diameter of the launching horn. In addition, the line must be free of bends and kinds to prevent radiation loss.

13-773. COUPLING BETWEEN G-LINE AND WAVEGUIDE. A convenient method of feeding a G-line is from a section of waveguide, and the energy is received from the G-line through another section of waveguide. The coupling at each end of the G-line is the same, and consists of a tapered horn which matches a coaxial section to the surface waveguide. In order to support the surface guide, the coaxial section is made short and is coupled to a rectangular waveguide. The probe which couples the rectangular guide to the coaxial section is then used as a support for the G-line conductor, as shown in figure 13-217.

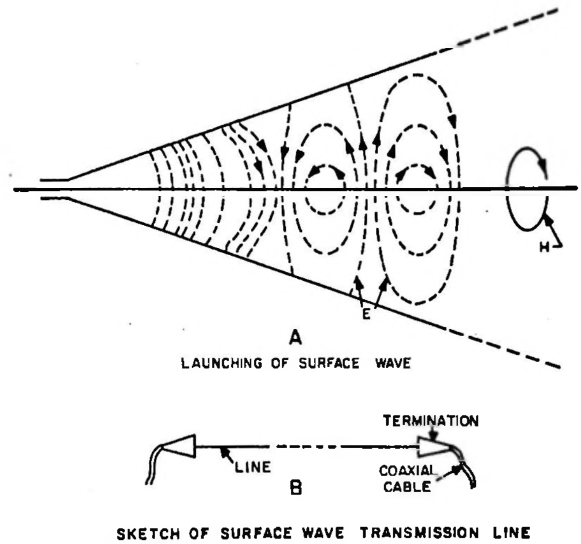


Figure 13-216. Surface Wave

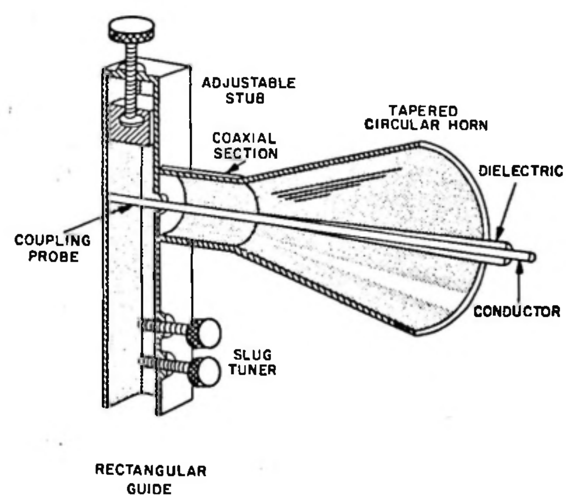


Figure 13-217. Coupling to G-Line

SECTION XI

CAVITY RESONATORS

13-774. GENERAL.

13-775. A cavity resonator, sometimes called a resonant cavity, is nothing more than a space bounded by conducting walls (a hollow conductor) having dimensions that make it resonant at a particular frequency. Cavity resonators are used, at microwave frequencies, in place of the lumped LC (inductance-capacitance) circuits and transmission line stubs employed at lower frequencies. The behavior of cavity resonators is discussed in detail in the following paragraphs. Before proceeding to this analysis, it is desirable to examine the behavior of a simple cavity, to gain an understanding of how it resonates.

13-776. SIGNAL REINFORCEMENT.

13-777. Consider the section of rectangular waveguide excited in the $TE_{0,1}$ mode by a small probe as shown in figure 13-218. The transmission equipment that supplies energy to the probe is not shown, since it does not enter the discussion. The waveguide extends a distance, X, to the left of the probe, where it is terminated by a conducting wall, and it extends indefinitely to the right of the probe. The probe, of course, radiates signals to the right and to the left, and at any instant a signal moving to the right is made up of two parts. One part is simply the signal that the probe radiates to the right; the other part is the signal that has been reflected from the termination at the left (that is, the signal that was radiated to the left by the probe a short time before). Now, if distance X is equal to $\lambda_g/4$, the signal radiated to the left must travel a guide half-wavelength before it returns to the probe. Its phase is reversed on reflection at the conducting wall, so that it reaches the probe with a phase shift of 180 degrees. Because it travels a guide half-wavelength, its traveling time is exactly 1 half-cycle, and during this time the phase of the signal radiated by the probe also changes by 180 degrees. If, therefore, distance X is one-quarter of λ_g (the guide wavelength), the signal radiated to the right by the probe will be reinforced by the reflected signal, originally radiated to the left by the probe, which is traveling to the right. If distance X is changed, the traveling time is also changed. The phase shift in the reflected signal, however, is always 180 degrees, since it results from reflection at the terminating wall, and not from traveling time. It follows, therefore, that the reflected signal will reinforce the signal radiated to the right whenever the traveling time is an odd number of half-cycles, since the phase shift at the probe during this time is 180 degrees plus some whole number of cycles. Reinforcement thus occurs whenever 2X is an odd number of guide half-wavelengths, or whenever X is an odd multiple of $\lambda_g/4$.

13-778. SIMPLE RESONATOR.

13-779. Now, consider the resonant cavity shown in figure 13-219. This consists of a section of waveguide similar to the one discussed previously, except that it is terminated in both directions by conducting walls.

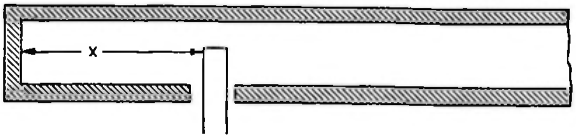


Figure 13-218. Waveguide Terminated by Conducting Wall

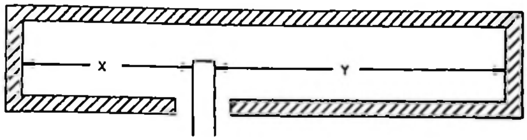


Figure 13-219. Simple Resonant Cavity

If, now, both X and Y are odd multiples of $\lambda_g/4$, signals radiated to the right will be reinforced by those reflected from the left-hand wall, and signals radiated to the left will be reinforced by those reflected from the right-hand wall. The bulk of the signal in each direction, consequently, is supplied by reflection from the opposite end, and only a small amount of energy need be supplied by the probe. For a given signal level in the cavity, the required input power is a minimum when X and Y are odd multiples of a guide quarter-wavelength. Also, for a given signal input, the signal level in the cavity is greatest when X and Y are odd multiples of $\lambda_g/4$. The wavelengths or frequencies for which this is true are said to be characteristic frequencies of the cavity, and the cavity is said to resonate at these frequencies. The relation between X, Y, and λ_g is tabulated in table 13-5 for the four characteristic frequencies having the longest guide wavelengths. Note that there are several possible positions for the probe at the higher characteristic frequencies. The longest guide wavelength is the one for which the length of the cavity is $\lambda_g/2$, and the guide wavelengths of the other characteristic frequencies are in the ratio 1, 1/2, 1/3, 1/4, etc. Unfortunately, the relation between λ_g and λ_a is not the same for all of the characteristic frequencies, and the wavelengths in air do not satisfy this simple relation, nor are the characteristic frequencies related by the reciprocals of these ratios (1, 2, 3, 4, etc). Note that a second probe can be added to take energy out of the cavity. If, then, the input probe is excited

Table 13-5. Resonant Cavity Distance Versus Wavelength

X	Y	X + Y	λg
$\lambda_g/4$	$\lambda_g/4$	$\lambda_g/2$	2(X + Y)
$\lambda_g/4$	$3\lambda_g/4$	λ_g	(X + Y) = 2 (X + Y)/2
$3\lambda_g/4$	$\lambda_g/4$	λ_g	(X + Y) = 2 (X + Y)/2
$\lambda_g/4$	$5\lambda_g/4$	$3\lambda_g/2$	2(X + Y)/3
$3\lambda_g/4$	$3\lambda_g/4$	$3\lambda_g/2$	2(X + Y)/3
$5\lambda_g/4$	$\lambda_g/4$	$3\lambda_g/2$	2(X + Y)/3
$\lambda_g/4$	$7\lambda_g/4$	$4\lambda_g/2$	2(X + Y)/4
$3\lambda_g/4$	$5\lambda_g/4$	$4\lambda_g/2$	2(X + Y)/4
$5\lambda_g/4$	$3\lambda_g/4$	$4\lambda_g/2$	2(X + Y)/4
$7\lambda_g/4$	$\lambda_g/4$	$4\lambda_g/2$	2(X + Y)/4

at a constant signal level, the signal level in the cavity will be maximum when the cavity resonates (the resonant frequency of the cavity may be changed by tuning it, or the frequency of the input signal may be changed to obtain resonance), and the energy transferred to the output probe will also be maximum at resonance, since the output probe takes energy from the cavity.

13-780. PRINCIPLES OF RESONANT CIRCUITS.

13-781. Since the behavior of a cavity resonator resembles that of an LC circuit as well as that of a transmission line section, it is convenient to introduce the detailed analysis of resonator behavior by examining the behavior of the two conventional circuits.

13-782. RESONANT LC CIRCUIT.

13-783. Consider the shunt-resonant, zero-resistance LC circuit shown in figure 13-220. This circuit consists of an inductor, L, in parallel with a capacitor, C. When an alternating voltage is applied to the terminals of this circuit, the phase of the current through the inductor is 90 degrees behind the phase of the applied voltage. The phase of the current through the capacitor is 90 degrees ahead of the phase of the applied voltage, and the two currents (through the inductor and the capacitor), therefore, differ in phase by 180 degrees. At any frequency, the two currents cancel each other to some extent; however, at the resonant frequency, f_r, the two currents are equal in magnitude (and opposite in phase). Consequently, they cancel each other completely and no current flows through the LC circuit from the external source, although heavy circulating current flows in the circuit loop. The resonant frequency is defined by the following relation:

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

At the resonant frequency, the impedance of this parallel circuit is infinite; thus the circuit passes no current. In practice, the inductor has some resistance and the capacitor may have both lead resistance and leakage. The impedance at resonance in a practical case is, therefore, not infinite, but rather very great. For the purpose of this discussion, it is sufficient to know that the impedance is very high at the resonant frequency.

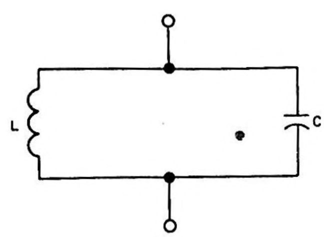


Figure 13-220. Resonant LC Circuit

13-784. RESONANT TRANSMISSION-LINE SECTION.

13-785. The circuit shown in part A of figure 13-221 is nothing more than a section of transmission line, of length X, terminated by a short circuit. The previous discussion of stub tuning makes it clear that there will be a reflected signal whenever an alternating voltage is applied to the input terminals of the shorted section of transmission line. The amplitude of the reflected signal will be equal to that of the direct signal and, at the termination, their phases will differ by 180 degrees. This is true because the terminating impedance is zero, since the direct and reflected signals have equal amplitudes and differ in phase by 180 degrees. The current, however, is not zero, since the currents of the direct and reflected signals add directly; forward current for the direct signal flows in the same direction as backward

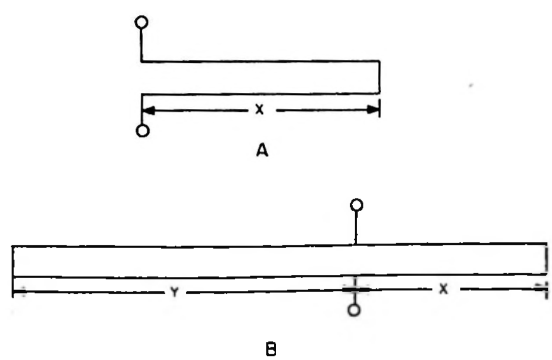


Figure 13-221. Resonant Transmission Line Sections

current for the reflected signal. At a point a quarter-wavelength from the termination, the phase of the direct signal is advanced by 90 degrees, and that of the reflected signal is retarded by 90 degrees. The two signals, which differ in phase by 180 degrees at the termination, are therefore in phase with each other at a point a quarter-wavelength from the termination. The voltages of the two signals then add, whereas the currents cancel each other (the forward currents of the direct and reflected signals flow in opposite directions). Since the current is zero, the impedance at a point a quarter-wavelength from the termination is infinite.

13-786. The same condition obtains at all points an integral number of half-wavelengths from this point (points an odd number of quarter-wavelengths from the termination), since moving a half-wavelength along the line changes the phase of each signal by 180 degrees and, therefore, leaves the signals in phase with each other. In practice, of course, the impedance is very high but not infinite. The behavior of this circuit, then, is described by the statement that the input impedance is infinite whenever the frequency of the applied signal is such that X is an odd number of quarter-wavelengths. Now, consider the circuit shown in part B of figure 13-221, which is a section of transmission line terminated at each end by a short circuit, with input terminals connected at a suitable point. This circuit will have an infinite input impedance if X is an odd number of quarter-wavelengths and Y is also an odd number of quarter-wavelengths. The total length of the section of transmission line, for an infinite input impedance at a suitable point, is, then, (X + Y), which is an even number of quarter-wavelengths (since the sum of two odd numbers is an even number) or some multiple of a half-wavelength. Consequently, there will be a point (or a number of points) an odd number of quarter-wavelengths from each end of the section at which the impedance is infinite; this is true whenever the frequency is such that the total length of the section is an exact multiple of a half-wavelength. The characteristic wavelengths for which the input impedance is infinite are given by the following relation:

$$\lambda_k = 2(X + Y)/k$$

where k is a positive integer (1, 2, 3, etc), and where the wavelength is the wavelength on the transmission line. The dominant mode of resonance is the one having the greatest wavelength. The characteristic wavelength for the dominant mode is, therefore, as follows:

$$\lambda_1 = 2(X + Y)/1 = 2(X + Y)$$

The wavelength for the k mode, is, of course, λ_k, and this is 1/k times λ_1. It follows, therefore, that f_k, the characteristic frequency of this mode, is approximately k times f_1. The characteristic frequencies are very nearly exact multiples of characteristic frequency f_1 of the dominant mode. The harmonic relation (exact multiples) is precise if the ratio between the wavelength on the transmission line and the wavelength in air is the same at all frequencies. The behavior of the circuit shown in part B of figure 13-221 is described by the statement that the input impedance is infinite at every characteristic frequency. Each characteristic frequency, f_k, is associated with a characteristic wavelength, λ_k, which is given by the following relation:

$$\lambda_k = 2Z/k$$

where Z is the length of the transmission line section.

13-787. RESONANT CAVITY.

13-788. It is impossible to write a simple expression for the characteristic frequencies of a general cavity resonator. If, however, the resonator consists of a section of waveguide terminated at each end by a conducting wall, it is precisely equivalent to the section of transmission line just described. Here, the characteristic guide wavelengths for which the input impedance of the resonator (at a suitable point) is infinite are given by the following relation:

$$\lambda_{gk} = 2Z/k$$

This means that the wavelength in the guide, λ_{gk}, is equal to twice the length of the wavelength, 2Z, divided by k, which is any positive integer. Trouble arises as soon as an attempt is made to compute the wavelength in air and the corresponding frequency which is associated with a given λ_{gk}. In order to determine λ_{ak} from λ_{gk}, it is necessary to know the dimensions of the waveguide (its height and width if it is rectangular, and its diameter if it is circular), in addition to the length of the section. It is also necessary to know the mode in which the waveguide is operating. Finally, the relation between λ_a and λ_g may be the same for more than one mode; therefore, there may be two distinct field patterns for a given characteristic frequency. For the moment, it is necessary only to point out that there is neither a single characteristic frequency, as in the LC circuit, nor a simple series of characteristic frequencies, as in the transmission line section. It is stated frequently that a cylindrical resonator may be thought of as an infinite number of quarter-wave shorted stubs (see figure 13-222). This implies that the characteristic frequencies are related to each other in the same way as those of the transmission line section previously described. This explanation of resonator behavior completely neglects the existence of various modes in waveguides, and is so oversimplified that it is of practically no value. The only constructive way of thinking about a cavity resonator is to consider it as a section of waveguide.

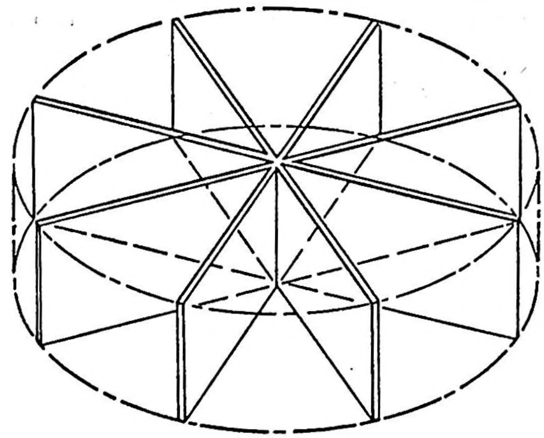


Figure 13-222. Characteristics of Cylindrical Resonator

13-789. RECTANGULAR RESONATORS.

13-790. Since the mathematics of rectangular cavity resonators is based on the mathematics of a rectangular waveguide, such resonators are not difficult to understand. It is possible to discuss rectangular cavity resonators in considerable detail, using simple mathematics, whereas this is not true of other resonators. It must be kept in mind that the cavity consists of a section of rectangular waveguide terminated at each end by a conducting wall, which is assumed to be a perfect conductor.

13-791. BASIC THEORY.

13-792. Consider a rectangular cavity resonator having a smallest dimension (height), a, an intermediate dimension (width), b, and a longest dimension (length), c. The use of c for the physical length of a rectangular cavity resonator should not be confused with its use for the electrical length of a light wave or a radio signal. For the moment, then, consider this resonator as a waveguide section of length c, height a, and a width b. Suppose, now, that the resonator is excited at a frequency for which the following relation exists:

$$\lambda_g = 2c/p$$

where p is a positive integer. Because of the boundary conditions at a conducting surface, which require that the electric field parallel to the surface and the magnetic field perpendicular to the surface dissipate at the surface, the signal in the waveguide will be reflected from the end of the cavity with a reversal of phase. The direct and incident field patterns, then, will be 180 degrees out of phase, and will, of course, be traveling in opposite directions.

13-793. At the end of the cavity, the transverse electric fields of the direct and reflected field patterns in the waveguide will cancel each other, as will the longitudinal magnetic fields. The longitudinal electric fields and the transverse magnetic fields, however, will add, because the forward direction for the direct signal is opposite the forward direction for the reflected signal. At a quarter-wavelength from the end of the cavity, the direct and reflected field patterns are in phase. At this point, the transverse electric fields and the longitudinal magnetic fields add, and the longitudinal electric fields and the transverse magnetic fields cancel. At a quarter-wavelength farther from the end of the cavity, the earlier situation (the phase relation that is present at the end wall) is repeated. The result is a field pattern, formed by the direct and reflected signals, which is repeated every half-wavelength along the cavity. All of the distances are measured, of course, in terms of the guide wavelength. If, now, the waveguide is operating in either the $TE_{m,n}$ or the $TM_{m,n}$ mode, and the length of the cavity is p times $\lambda_g/2$, the resulting field pattern in the cavity will be repeated m times along the a dimension, n times along the b dimension, and p times along the c dimension.

13-794. MODES OF OPERATION.

13-795. In this discussion, it has been assumed that the rectangular cavity was operated by making waveguide signals travel from one end of the cavity to the other. It is, of course, possible to send signals across the cavity, or between the top and the bottom. The length of the waveguide section and the dimensions of the waveguide are tabulated in table 13-6 for a cavity of height a, width b, and length c.

13-796. OPERATION WITH TE AND TM SIGNALS.

13-797. Suppose that a rectangular cavity, of height a, width b, and length c is excited by sending $TE_{m,n}$ signals along the cavity, from end to end, at a wavelength for which λ_g is equal to 2c/p. At this point, consider what happens if the $TE_{m,n}$ signals are re-

Table 13-6. Waveguide Section Dimensions

DIRECTION OF WAVEGUIDE SIGNALS	LENGTH OF WAVEGUIDE SECTION	HEIGHT OF WAVEGUIDE	WIDTH OF WAVEGUIDE
End to end	c	a	b
Side to side	b	a	c
Top to bottom	a	b	c

placed by $TM_{m,n}$ signals having the same guide wavelength, so that λ_g remains equal to 2c/p. It is assumed that m and n are the same for the TE and the TM signals. $TE_{3,5}$ signals, for example, are replaced by $TM_{3,5}$ signals. Assume that a waveguide pattern has the following guide wavelength:

$$\lambda_g = \lambda_a/\cos A \cos B$$

This waveguide can be built up from four sets of vertically polarized waves, the paths of which make angles A with the top and bottom of the waveguide, and angles B with the side walls of the guide; also, a similar pattern can be built up from horizontally polarized waves having the same λ_a. If the same values of m and n are used, both patterns will have the same guide wavelength. Although angles A and B are different for the two patterns, the product cos A cos B is the same for both. Thus, if m, n, and λ_a are the same for both patterns, λ_g will also be the same. It has been pointed out that these patterns can be combined with one phase relation to obtain the pattern of a $TE_{m,n}$ mode, and with another phase relation to obtain the pattern of a $TM_{m,n}$ mode. It follows, therefore, that a $TE_{m,n}$ mode and a $TM_{m,n}$ mode have the same guide wavelength if m, n, and λ_a are the same for both modes. Alternatively, if m, n, and λ_g are the same for both modes, they will have the same λ_a. It is clear, then, that λ_a is unchanged when the $TE_{m,n}$ signals, having λ_g equal to 2c/p, are replaced by $TM_{m,n}$ signals having the same m, n, and λ_g. Thus, λ_a for either case is given by the following relation:

$$\lambda_a = 1/\sqrt{(m/2a)^2+(n/2b)^2+(p/2c)^2}$$

13-798. CHARACTERISTIC WAVELENGTHS.

13-799. The characteristic wavelength (the value of λ_a) was previously calculated for a cavity of height a, width b, and length c excited by $TE_{m,n}$ or $TM_{m,n}$ signals traveling along the cavity, from end to end, and having a wavelength for which λ_g is equal to 2c/p. This mode of exciting the cavity produces a field pattern that is repeated m times along the a dimension, n times along the b dimension, and p times along the c dimension. An examination of the expression for λ_a shows that it involves a, b, c, m, n, and p, but makes no mention of the direction in which the waveguide signal travels. It follows, therefore, that λ_a will have the same value (in a given cavity) for every mode of operation in which the field pattern is repeated m, n, and p times along the a, b, and c dimensions, respectively. The characteristic wavelengths are, then, those for which the following relation holds:

$$\lambda_a = \lambda_{m,n,p} = \frac{}{1/\sqrt{(m/2a)^2+(n/2b)^2+(p/2c)^2}}$$

where there may be more than one mode of operation of the cavity at a given characteristic wavelength. The longest characteristic wavelength is, to some extent, equivalent to the cutoff wavelength of a waveguide, since the cavity cannot resonate at longer wavelengths. The parallel is not exact, however, because the waveguide will operate at any frequency above cutoff, whereas the cavity resonates only at its characteristic frequencies.

13-800. TERMINOLOGY OF MODES.

13-801. Modes in a rectangular cavity resonator are described as TE modes if they are excited by TE waveguide signals, and as TM modes if they are excited by TM waveguide signals. Each mode is described further by three subscripts which show how many times the field pattern is repeated along the height, width, and length of the cavity. There are, therefore, $TE_{m,n,p}$ modes and $TM_{m,n,p}$ modes. For example, if the electric field is perpendicular to the top and bottom of the cavity, it is possible to have a pattern that is uniform along one dimension (in this case, the height), so that one subscript may be zero. Here, m is zero. A field that is perpendicular to two walls is, however, parallel to the other four. Therefore, a pattern that is uniform in one dimension must vary along both of the other dimensions in order that the electric field may satisfy the boundary conditions by dissipating at the surface of each of the four walls to which it is parallel. It is thereby impossible to have a mode in which the field pattern is uniform along more than one direction, and there are no modes for which more than one subscript is zero.

13-802. Field patterns that are uniform along one dimension are excited by waveguide signals like those of the $TE_{0,n}$ modes or the $TE_{m,0}$ modes in a rectangular waveguide, and are, accordingly, TE modes. The cavity resonator, therefore, may operate in either a $TE_{m,n,p}$ mode or a $TM_{m,n,p}$ mode at a characteristic wavelength, $\lambda_{m,n,p}$, for which all of the subscripts are nonzero, and it may operate only in a $TE_{m,n,p}$ mode at a characteristic wavelength, $\lambda_{m,n,p}$, for which one of the subscripts is zero. There are no characteristic wavelengths for which more than one subscript is zero.

13-803. FIELD PATTERNS OF MODES.

13-804. The dominant characteristic wavelength (the longest one) is $\lambda_{0,1,1}$. The corresponding mode is the $TE_{0,1,1}$ mode, the field pattern of which is shown in part A of figure 13-223. This mode may be excited by a $TE_{0,1}$ signal traveling along the resonator, from end to end, or by a $TE_{0,1}$ signal traveling across the resonator from side to side. The field pattern is the same in both instances, so that there is, in fact, only a single $TE_{0,1,1}$ mode. Similar patterns are obtained in the $TE_{1,0,1}$ mode, at $\lambda_{0,1,1}$, where the electric field is perpendicular to the sides of the cavity (in the $TE_{0,1,1}$ mode the electric field is perpendicular to the top and bottom of the cavity), and in the $TE_{1,1,0}$ mode, at $\lambda_{1,1,0}$, where the electric field is perpendicular to the ends of the cavity. All TE modes having one subscript equal to zero have field patterns that are repetitions of the simple $TE_{0,1,1}$ pattern. A single example, the pattern of the $TE_{0,1,2}$ mode, is shown in part B of the figure.

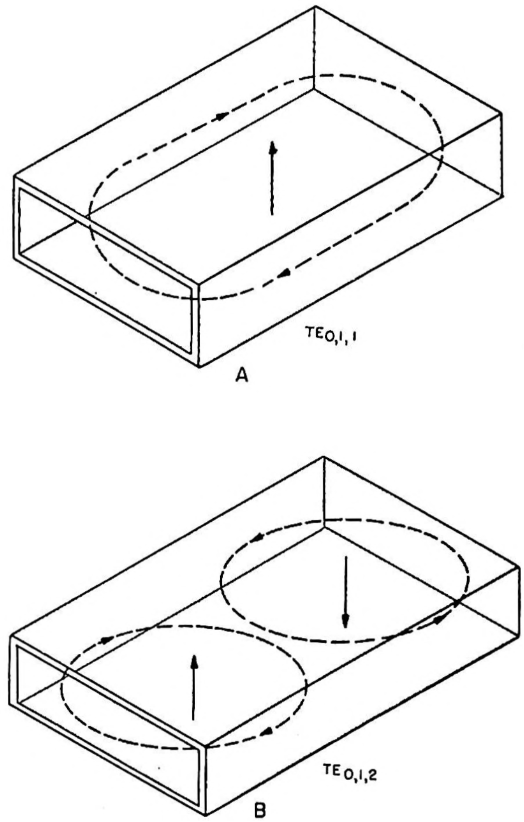


Figure 13-223. Simple Resonator Modes

13-805. The $TE_{0,1,2}$ mode may be excited by a $TE_{0,1}$ signal traveling along the resonator or by a $TE_{0,2}$ signal traveling across the resonator. The pattern, however, is the same in each case. For any characteristic wavelength having one subscript equal to zero, there is a single TE mode at which the cavity may operate. There are several possible modes, however, for a characteristic wavelength having three nonzero subscripts. At $\lambda_{1,1,1}$, for example, the cavity may be excited by a $TE_{1,1}$ signal traveling along the cavity from end to end. This develops a $TE_{1,1,1}$ mode, with the pattern shown in part A of figure 13-224. Note that the electric field is parallel to the ends of the cavity. A second $TE_{1,1,1}$ mode may be developed by a $TE_{1,1}$ signal traveling across the cavity. This will have a similar pattern, but the electric field will be parallel to the sides of the cavity. Finally, a third $TE_{1,1,1}$ mode may be developed by a $TE_{1,1}$ signal traveling from the top of the cavity to the bottom. Here, the electric field will be parallel to the top and bottom of the cavity. There are also, for the same characteristic wavelength, three $TM_{1,1,1}$ modes, all excited by $TM_{1,1}$ signals. In one, shown in part B, the signals travel along the cavity, from end to end, and the magnetic field is parallel to the ends of the cavity. In the two other modes, the signal travels across the cavity, with the magnetic field parallel to the sides of the cavity, or it travels from top to bottom, with the magnetic field parallel to the top and bottom of the cavity. The various patterns for characteristic wavelengths with subscripts greater than 1 are repetitions of the simple patterns. For any characteristic wavelength having three nonzero subscripts, there are three TE modes and three TM modes in which the cavity may operate.

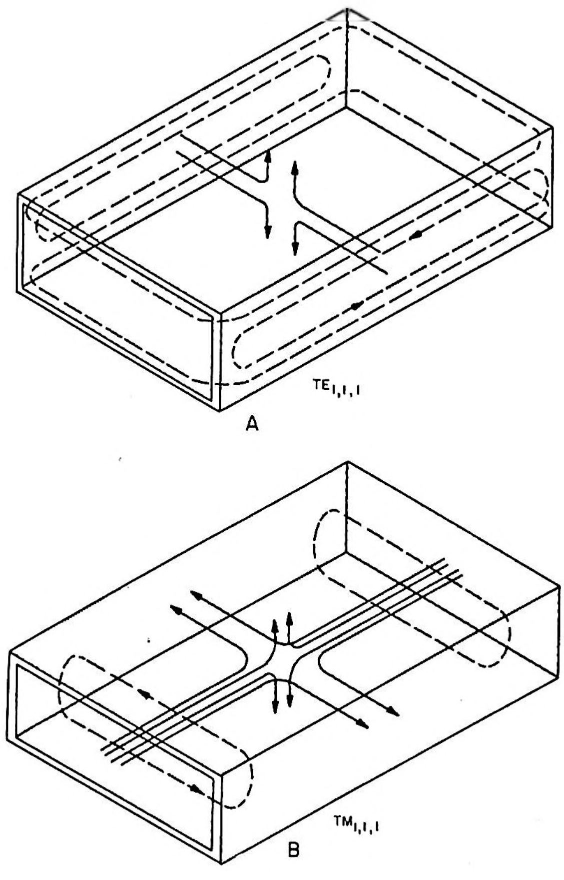
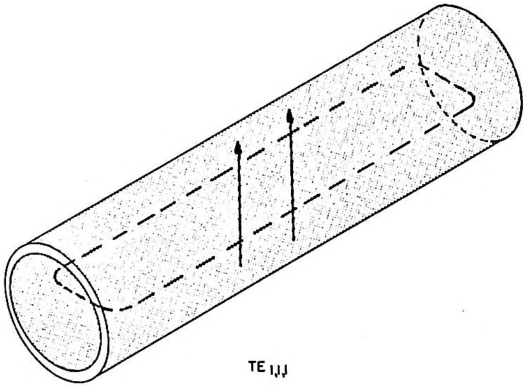


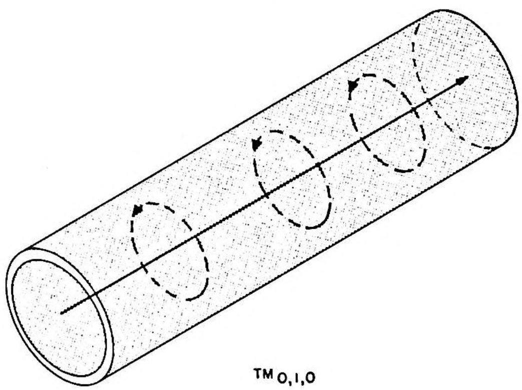
Figure 13-224. Higher Resonator Modes

13-806. CYLINDRICAL RESONATORS.

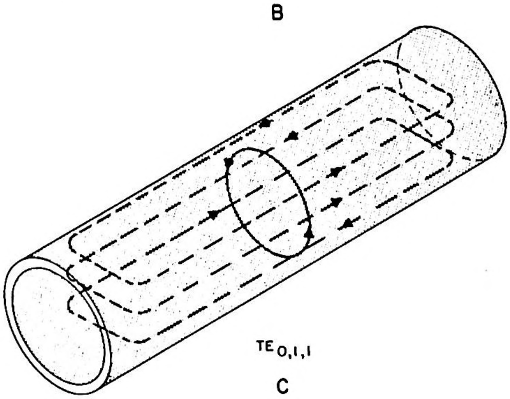
13-807. A detailed discussion of cylindrical resonators, like a discussion of circular waveguides, involves extremely complicated mathematics beyond the scope of this manual. Therefore, the subject is discussed only briefly. As with a rectangular resonator, the characteristic wavelengths of a cylindrical resonator are distinguished from one another by three subscripts, m, n, and p. There is, unfortunately, no simple expression that relates $\lambda_{m,n,p}$ to the dimensions of the resonator. The subscripts, m and n, like those used to describe a mode in a circular waveguide, refer to the number of times the field pattern is repeated around the circumference of the cylinder, m, and the number of times the pattern is repeated from the axis of the cylinder to its circumference, n. As in the rectangular resonator, the third subscript, p, refers to the number of times the field pattern is repeated along the length of the resonator (along the axis of the cylinder). The simplest mode in a cylindrical resonator is the $TE_{1,1,1}$ mode, shown in part A of figure 13-225, which corresponds to the $TE_{0,1,1}$ mode in a rectangular resonator. Another simple mode is the



A



B



C

Figure 13-225. Field Patterns in Cylindrical Resonators

$TM_{0,1,0}$ mode, shown in part B, which corresponds to the $TE_{1,1,0}$ mode in a rectangular resonator, and is frequently seen in resonators that have a modified cylindrical form. Among such resonators are the cavity of an X-band klystron, discussed in the following paragraph, and the cavities used for TR and ATR switches. Another important mode is the $TE_{0,1,1}$ mode, shown in part C. A cavity resonator operated in this mode has extremely low losses. Therefore, its input impedance at resonance is extremely high, and the variation in impedance with change in frequency is very sharp. Microwave wavemeters are often cylindrical cavities operating in this mode.

13-808. RE-ENTRANT CAVITY.

13-809. Cavity resonators are seldom simply rectangular or cylindrical. The rectangular resonator has been discussed in detail to illustrate the basic theory behind all cavity resonators, and the cylindrical resonator has been described briefly because most actual resonant cavities operate in modes similar to the $TM_{0,1,0}$ and $TE_{0,1,1}$ modes of a cylindrical resonator. Three representative re-entrant cavities are illustrated in figure 13-226. The term re-entrant indicates that a portion of the wall protrudes into the space within the cavity. Part A shows the resonant cavity of an X-band reflex klystron, part B shows a TR cavity, and part C shows a re-entrant wavemeter cavity. All three cavities operate in modes similar to the $TM_{0,1,0}$ mode of a cylindrical resonator. In each, the direction of the electric field is indicated.

13-810. CAVITY RESONATOR TUNING.

13-811. One method of tuning a cavity resonator is to change its dimensions by moving one of its walls in much the same way that the termination of a tuning stub is moved. This is actually done in some wavemeter cavities. Another method is to distort the cavity by the application of pressure at appropriate points. A third method is to distort the electric field or the magnetic field, thus changing the effective dimensions of the cavity. The wavemeter cavity shown

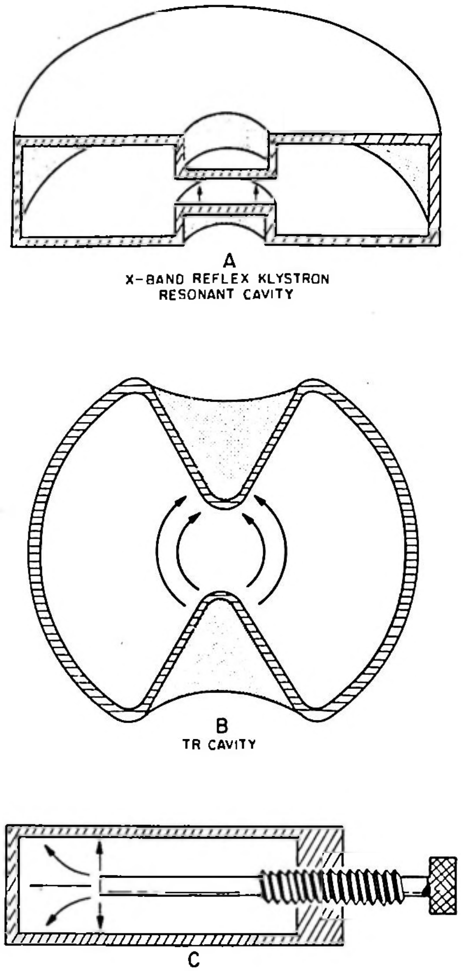


Figure 13-226. Cavity Resonators of Various Shapes

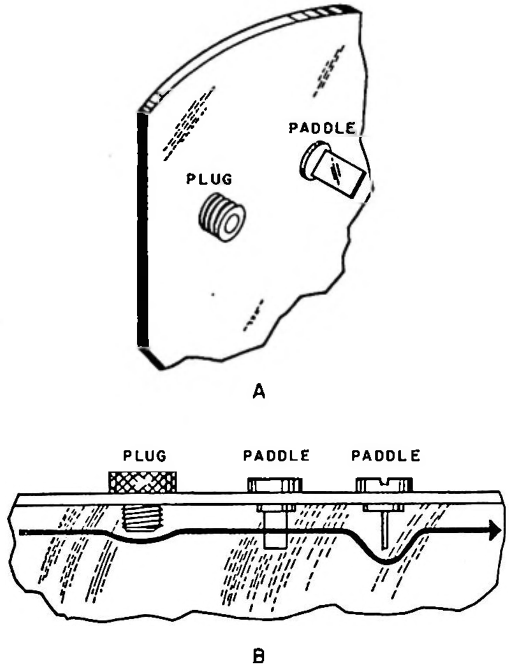


Figure 13-227. Tuning Plug and Paddle

in part C of figure 13-226 is tuned in this manner; changing the position of the pin alters the electric field. The magnetic field in a cavity is usually changed by inserting plugs which extend a short distance from the wall. For large changes in the magnetic field, tuning paddles are frequently used instead of plugs. A tuning plug and a tuning paddle are shown in the wall of a cavity in part A of figure 13-227. Part B of the figure shows the consequent distortion of the magnetic field. The operation of the plug and paddle are illustrated. So long as the plane of the paddle is parallel to the direction of the magnetic field, it has no effect; when the plane of the paddle is perpendicular to the direction of the field, the distortion is greatest.

13-812. COUPLING TO CAVITY RESONATORS.

13-813. Energy may be fed into or taken from a cavity resonator in three ways—by using a probe, a loop, or a slit.

13-814. PROBE COUPLING.

13-815. Probe coupling is seldom used with resonant cavities. If the probe extends deep into the cavity, like the probe used to feed a waveguide, it is likely to load the cavity so heavily that the resonance peak is neither very high nor very sharp. If, on the other hand, the probe extends only a small distance into the cavity, there is usually some difficulty in exciting the proper operational mode. For example, part A of figure 13-228 shows the configuration of the electric field in a rectangular cavity resonating in the $TE_{0,1,1}$ mode, and part B shows the configuration of the electric field set up by a probe. Because of this, the use of a probe is usually limited to applications where tight coupling is desired, such as to and from a waveguide, from a directional coupler, or where the signal is set up by some other method and only a small amount of energy is to be taken out (as from slotted sections of waveguide or coaxial line).

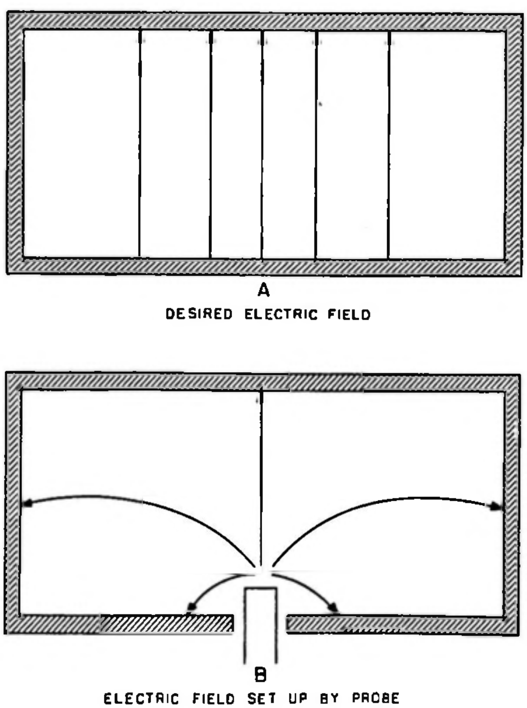


Figure 13-228. Effect of Probe Coupling on Electric Field

13-816. LOOP COUPLING.

13-817. Loop coupling is more common than any other type. Figure 13-229 shows a coupling loop used to excite the $TE_{0,1,1}$ mode in a rectangular resonator. The basic advantage of loop coupling lies in the fact that the coupling loop does not deform the electric field pattern to the same extent that a probe does. The degree of coupling is controlled by the size of the loop and sometimes, but less frequently, by the placement and orientation of the loop. If the loop is turned so that it lies in a plane parallel to the magnetic field of a given mode, it is not coupled at all to that mode, although there may be undesired coupling to other modes. Its operation in this respect is similar to that of the tuning paddle. One of the major uses of loop coupling is to take energy from the resonant cavity of a magnetron or a klystron; loop coupling is also used to couple to and from the resonant cavities of some wavemeters.

13-818. SLIT COUPLING.

13-819. Slit coupling, which may involve small holes instead of slits, is used mostly

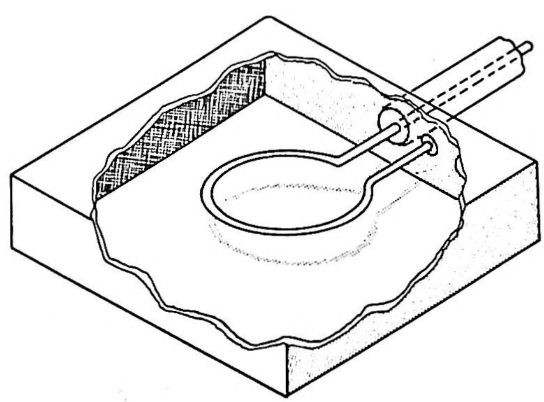


Figure 13-229. Loop Coupling to Cavity Resonator

in tr and atr switches. Energy leaks into or out of the resonant cavity by means of the slits or holes in the same manner that it leaks from the main waveguide through two small holes into the nonresonant auxiliary cavity of a directional coupler. The principal merits of slit coupling are mechanical simplicity and, in the case of tr and atr switches, the relatively smooth transition between the waveguide and the cavity when the coupling is tight.

13-820. APPLICATIONS OF CAVITY RESONATORS.

13-821. WAVEMETERS.

13-822. Most microwave wavemeters are essentially similar. A resonant cavity is coupled loosely to the source of signals, and a detector is coupled to the cavity. The cavity is then adjusted to resonate at the operating frequency (resonance being indicated by a maximum reading at the detector), and the frequency is read from a scale on the tuning adjustment. Figure 13-230 shows the equivalent circuit of such a wavemeter, and figure 13-231 shows a cutaway view of the tunable cavity with its input and output coupling loops.

13-823. SHARPNESS OF RESONANCE PEAK.

13-824. The impedance of a resonant circuit, whether it is a cavity resonator, a section of transmission line, or an LC circuit, reaches a maximum at its resonant frequency (or at each of its characteristic frequencies), and falls off at frequencies above and below this. As the losses in the equipment are increased, the peak becomes lower and less sharp. It is customary to describe the sharpness of the resonance peak by means of a quantity, Q, which is defined as follows: Let Z_r be the impedance at the resonant frequency (or the character-

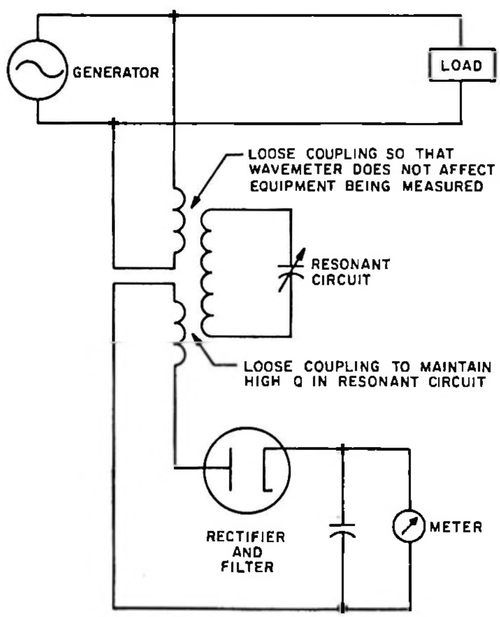


Figure 13-230. Equivalent Circuit of Wavemeter

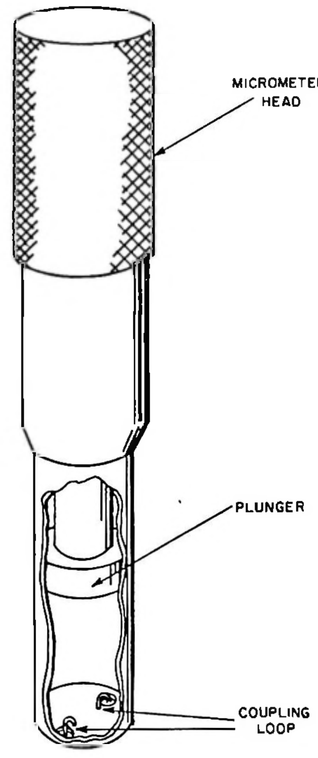


Figure 13-231. Wavemeter Tunable Cavity

istic frequency), f_r. At some frequency above f_r the impedance will have a magnitude equal to $Z_r\sqrt{1/2}$, or $.707Z_r$. Let this frequency be f_1. At some other frequency, below f_r, the impedance will have the same magnitude. Let this frequency be f_2. Then, by definition:

$$Q = f_r/(f_1 - f_2)$$

The Q of a cavity resonator is commonly several thousand.

13-825. EQUIVALENT CIRCUIT.

13-826. In figure 13-230, the tuned circuit is coupled loosely to the transmission line in order to avoid disturbing the arrangement whose frequency is being measured. In addition, the detector is coupled very loosely to the resonant circuit in order to avoid lowering the Q. The detector shown consists of a conventional rectifier and filter circuit which feeds a meter; at microwave frequencies, the diode rectifier is replaced by a crystal rectifier, and the filter capacitor is part of the capacitance of the crystal holder. In some cases, an amplifier may be inserted between the rectifier output and the meter. The resonant circuit is adjusted so that the rectifier output is maximum. When several frequencies are present, the meter will show a maximum as the resonant circuit is tuned through each frequency.

13-827. TUNABLE CAVITY.

13-828. The tunable cavity shown in figure 13-231 consists of a cylindrical cavity of adjustable length operating in the $TE_{0,1,1}$ mode. The upper stop is moved by a conventional micrometer head until resonance is reached. The position of the plunger is read on a scale at the micrometer head, and the frequency is determined from a calibration curve. A cavity of this kind is a precision instrument and must be treated with great care. At the lower microwave frequencies, a simple cylindrical cavity becomes unduly large; thus, various modified forms are used. Two of these types are shown in figure 13-232. Various forms of coupling are used, loop coupling being the most common.

13-829. ABSORPTION WAVEMETER.

13-830. Another method of using a resonant cavity to determine frequency is shown in figure 13-233. Here a resonant cavity is coupled tightly to a waveguide so that it absorbs a substantial amount of energy when it is tuned to resonance. As the cavity is tuned to the operating frequency, the signal in the waveguide will decrease.

13-831. ECHO BOXES.

13-832. GENERAL.

13-833. An echo box is a piece of test equipment used to check the over-all per-

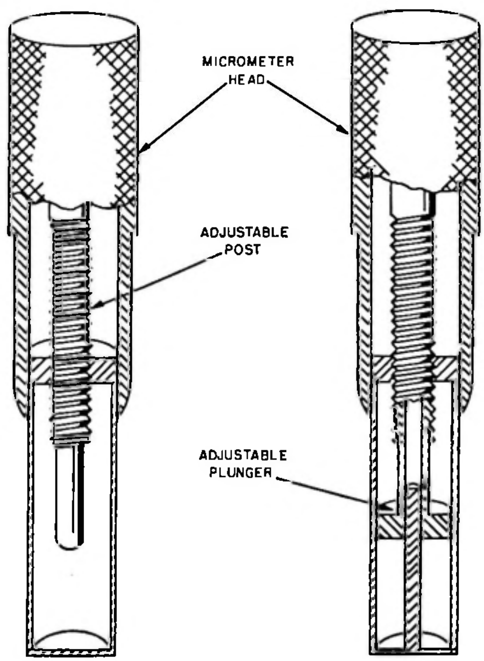


Figure 13-232. Wavemeter Tunable Cavities for Low Microwave Frequencies

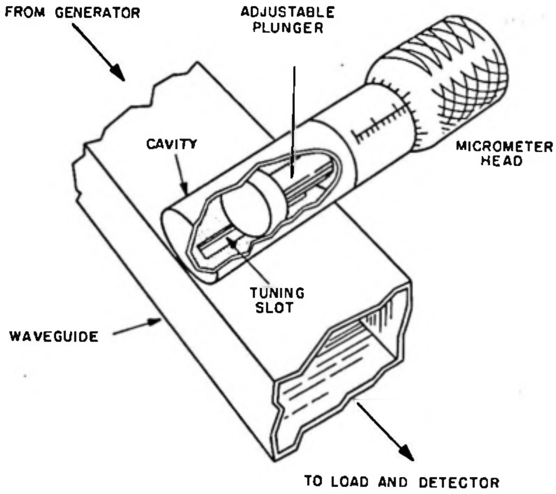


Figure 13-233. Resonant Cavity Used in Absorption Wavemeter

formance of a radar facility. It consists of a tunable resonant cavity connected through a transmission arrangement to a small antenna. The cavity is usually cylindrical, of adjustable length for tuning, and is operated in the $TE_{0,1,1}$ mode in order to obtain the highest possible Q. An S-band echo box, shown in figure 13-234, uses a small dipole antenna. This is connected through flexible coaxial cable to the resonant cavity, where a small loop is used for coupling between the cable and the cavity. An X-band echo box generally uses a small dipole antenna at the center of a parabolic reflector. The cavity is mounted on the back of the reflector, and a short length of rigid coaxial line carries energy between the dipole and the coupling loop in the cavity. In most cases, the resonant cavity is the tuning knob in a micrometer head, and the echo box is provided with a calibration chart so that you can use it as a wavemeter. Occasionally, a second coupling device is added to the cavity. This coupling device is generally a probe which is part of a crystal holder. The crystal rectifier output is fed to a meter so that the signal level in the cavity resonator can be measured.

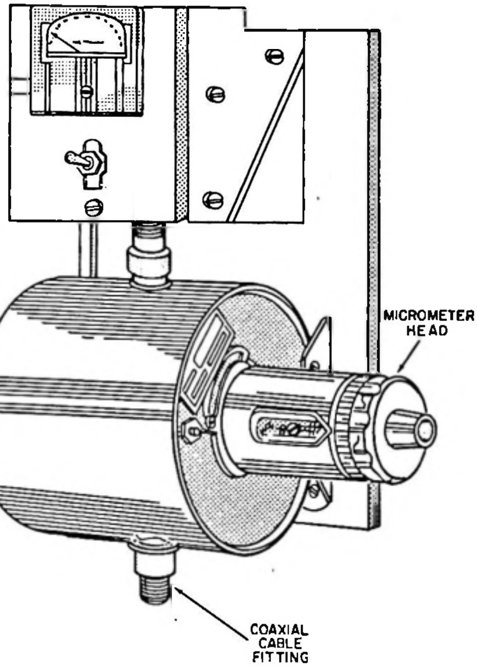


Figure 13-234. Echo Box

13-834. If a cavity resonator is excited by an external signal, the level of the signal in the cavity will rise, as shown in part A of figure 13-235. Similarly, if the excitation is stopped suddenly, the signal level in the cavity will decay, as shown in part B of the figure. It is assumed that the exciting signal is at the resonant frequency of the cavity. The decay time is related to the Q of the resonator, and the signal level will fall to .01 of its original value in 1.5Q cycles. Similarly, the rise time is related to Q, and the signal level will rise to .99 of its ultimate value (or will reach a point 1 percent short of its ultimate value) in 1.5Q cycles. The rise and decay times can also be expressed in terms of the time constant of the cavity, which is the time during which the signal level decays to about one-third (actually .368) of its original value, or during which it rises to about two-thirds (actually .632) of its ultimate value. In practice, the time constant of the cavity is so long that the signal level cannot rise to its ultimate

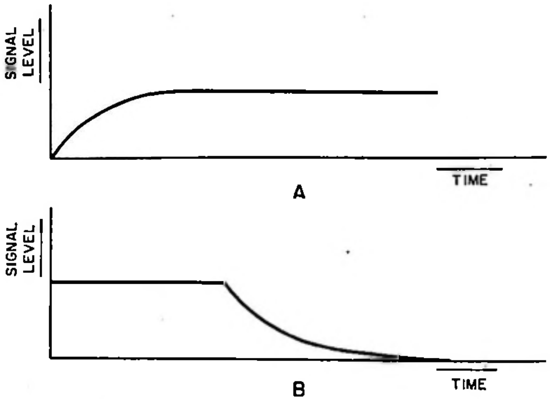


Figure 13-235. Rise and Decay of Signal in Resonator

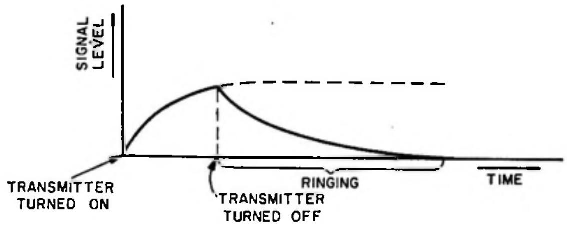


Figure 13-236. Signal Level in Echo Box

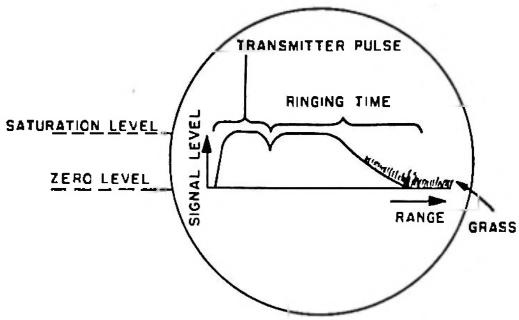


Figure 13-237. Echo Box Signal Waveform on A-Scope

value during the transmitter pulse. A graph of signal level against time is shown in figure 13-236.

13-835. APPEARANCE OF ECHO BOX SIGNAL.

13-836. Figure 13-237 shows the appearance of the echo box signal on an A-scope, which displays receiver output on the vertical axis versus range or time on the horizontal axis. During the transmitter pulse, the receiver is saturated and the output signal reaches its maximum level. Following the transmitter pulse, the receiver output results only from the signal picked up from the echo box (assuming that there are no short-range echoes). The output drops at first because the receiver sensitivity has been somewhat reduced as a result of saturation by the transmitter pulse, and also because the TR and ATR switches take some time to recover. The receiver, however, quickly recovers its normal sensitivity, and the signal from the echo box is then sufficient to drive the receiver to full output.

13-837. After a short interval during which the receiver is saturated, the echo box signal decreases so far that the receiver output also decreases. From this point, the signal from the echo box and the output from the receiver both decay until the receiver output is lost in the grass (electrical noise) on the A-scope. The ringing time is, of course, the time interval between the end of the transmitter pulse and the point where the echo box signal is lost in the grass.

13-838. It is customary to express the ringing time in whatever range units (yards or miles) are shown on the A-scope. The ringing time may be increased in three ways:

a. Raising the transmitter output causes the echo box resonator to be excited at a higher level; it then takes longer for the echo box signal to decay to a point where the receiver output is lost in the grass.

b. Increasing the duration of the transmitter pulse has the same effect, since the signal level in the resonator will rise farther toward its ultimate value if the period of excitation is short as compared with the time constant of the resonator. As before, a higher initial signal level will increase the time required for the signal to decay to the point where the receiver output is lost.

c. An increase in receiver sensitivity will allow the receiver to see smaller signals; consequently, the period of decay will be longer to the lower critical level where the receiver output is lost.

13-839. In this connection, it is necessary to distinguish between receiver sensitivity and receiver gain. Increasing the sensitivity raises the ratio of signal output to noise output for a given level of the input signal. Increasing the gain, however, raises both the signal output and the noise output without affecting their ratio. This will affect the appearance of the picture on the A-scope, but will not alter the ringing time. The appearances of the echo box signals on a B-scope, which displays range versus azimuth, and on a PPI scope (plan-position-indicator oscilloscope), which displays range versus direction, are shown in parts A and B, respectively, of figure 13-238. In each case, the receiver output controls the brightness of the spot on the scope. The ringing time is always taken from the greatest peak, which is obtained when the antenna is pointing at the echo box. When the antenna is pointing in any other direction, the ringing time decreases because the transmitter delivers a weaker signal to the echo box and also because the receiver output is reduced for a given signal from the echo box. The small peaks in the ringing time pattern result from side lobes in the antenna pattern. Although most of the energy is radiated in the main beam, a portion is radiated in directions near the beam. The behavior of the antenna for receiving is much the same. The strongest response from the receiver is obtained when the signal source lies in the main beam of the antenna, but signals can be picked up weakly from other directions transmitted close to the beam. This subject is discussed more fully in chapter 14, which deals with microwave antennas.

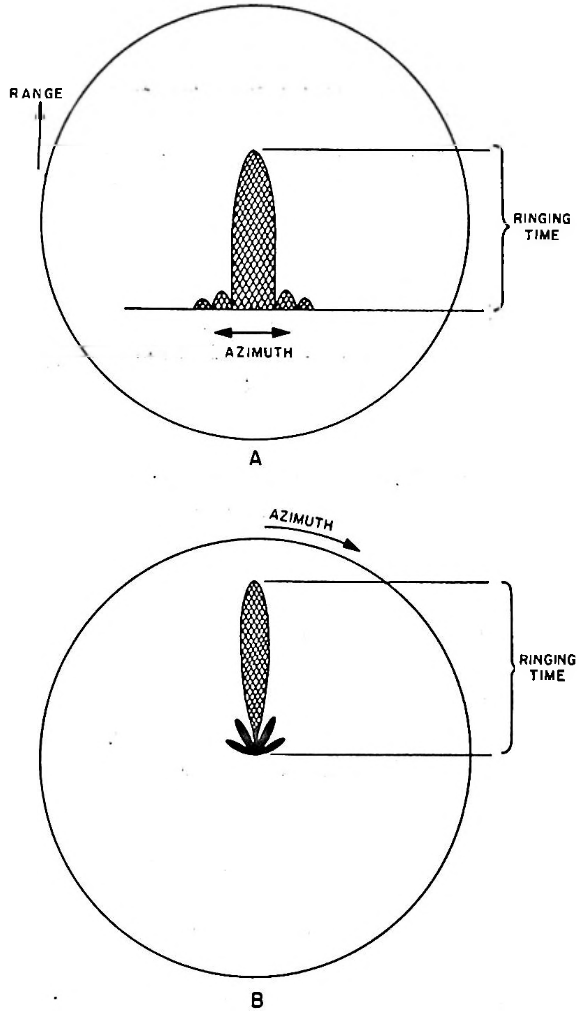


Figure 13-238. Echo Box Signal Waveforms on B-Scope and PPI

13-840. INTERPRETATION OF ECHO BOX TEST RESULTS.

13-841. In order to interpret the results of an echo box test, it is desirable to recall the expression developed for a given radar installation. Operating against a given target, the maximum range (unless limited by the horizon) is proportional to the fourth root of P/p', where P is the power output of the transmitter, and p' is the power input required by the receiver to develop a useable receiver output—an output signal that can be observed in spite of the electrical noise. This shows that an increase in the transmitter output or an increase in the receiver sensitivity, which is a decrease in p', will increase the maximum range against a given target and also will increase the ringing time with a given echo box. Therefore, any adjustment of the radar facility which increases the ringing time, except an increase in the duration of the transmitter pulse, also will improve the facility performance. The extent of the improvement can be calculated, either from the time constant of the resonator or from its Q, but the process is rather complicated and the calculation is not particularly useful. However, it is possible to estimate the improvement in performance when the echo box signal is displayed on an A-scope. First, note the range at which the echo box signal is just as strong as the noise signal on the scope. Then, after making an adjustment which increases the ringing time, note the new level of the echo box signal at the range where it was previously equal to the noise signal.

13-842. The percentage increase in range against targets of a given type will be about half of the percentage increase in the ratio of echo box signal to noise signal. The initial ratio between the two signals may be considered as 1, since the point of observation is chosen as the point where the signals are equal. If, after improvement, the echo box signal is 10 percent greater than the noise signal, the range on targets of a given type will be improved by about 5 percent. Although the echo box is one of the most valuable pieces of test equipment for tuning up a radar installation, it is dangerous to place much faith in comparisons between different ringing time tests unless the echo box is always placed in exactly the same position with the same orientation. Even then, any change in the temperature of the cavity resonator will alter its Q and, therefore, will change the ringing time, as will any change in the duration of the transmitted pulse. The arrangement should be considered satisfactory if the ringing time is within about 20 percent of its usual value; the echo box should then be used as an aid for the precise adjustment of the various tuning devices.

13-843. PROCEDURE FOR RINGING TIME TEST.

13-844. Considering the many different types of echo boxes, the ringing time test procedure is more complicated to describe than to carry out. In fact, the ringing time test is particularly useful just because it is so easy. In any case, adjust the transmitter first. The proper adjustment of the transmitter insures adequate output from the magnetron at the proper pulse duration. If the echo box is equipped with a detector, your next step is to tune it to the transmitter (by adjusting for maximum output from the detector). At this point, you should check the transmitter frequency and readjust the transmitter, if necessary, so that it operates on the proper frequency. (Unless the transmitter uses a tunable magnetron, you cannot alter its frequency; the frequency check is merely one step in the check for proper over-all operation.) Next, tune the receiver to the echo box signal, with the automatic frequency control arrangement turned off, and adjust the various tuning devices in the transmission equipment to obtain the longest ringing time. If the echo box is not equipped with a detector, adjust the transmitter first; then, by observing a local echo, tune the receiver to the transmitter with the automatic frequency control arrangement turned off. Next, tune the echo box to the transmitter by observing the ringing time while you tune the echo box for the longest ringing time. Thereafter, you adjust the tuning devices in the transmission equipment to increase the ringing time.

13-845. TR AND ATR SWITCHES.

13-846. Two separate antennas and transmission lines are sometimes used for transmitting and receiving rf signals (one for each function). By using the same transmission line and antenna for both functions, the equipment and installation of equipment can be simplified to a large degree. With a modified transmission line, simultaneous reception and transmission, using one antenna arrangement, can be accomplished. An important point to consider when the transmitter and receiver both operate on the same frequency is that you must keep the transmitted signal out of the receiver. The high transmitter power could damage the sensitive receiver to a point where it could not receive at all.

13-847. PARALLEL DUPLEXING EQUIPMENT.

13-848. In radar, where the receiver and transmitter use the same frequency, a duplexer equipment makes single antenna operation possible. The duplexer equipment contains tr (transmit-receive) and atr (anti-transmit-receive) circuits. The purpose of the tr circuit is to prevent the transmitter power from reaching the receiver. The purpose of the atr circuit is to disconnect the transmitter during the period the antenna is receiving. Part A of figure 13-239 shows a transmission line drawn as a two-wire line including the duplexer components just mentioned. The transmission line may be either a coaxial line or a waveguide; both operate on the same principle. In part A of figure 13-239, the receiver transmission line is in parallel with the transmitter transmission line. A spark gap is placed in the receiver line a quarter-wavelength from the junction. This is the tr spark gap. One quarter-wavelength from the receiver line junction, toward the transmitter, an additional quarter-wavelength line is placed in parallel with the transmitter. This line is also terminated in a spark gap, the atr spark gap. During transmission, both spark gaps fire (part B of figure 13-239); this causes the tr and atr circuits to act as shorted quarter-wave transmission lines, reflecting an open circuit condition at the point of transmission line connection so that no transmitted energy will enter either the tr or atr circuit.

13-849. The open circuit (maximum impedance) is reflected at both junctions to the main transmission line; therefore, the transmitted power is conducted to the antenna without loss because none of the power enters the receiver, tr, or atr line. During reception, the amplitude of the received power is not sufficient to fire either gap. Under this condition the atr circuit is now a quarter-wave transmission line terminated in an open circuit. One quarter-wave back from the open circuit termination a short circuit is in effect at the point of the atr junction with the transmission line, as shown in part C of figure 13-239. The current from this short circuit will "see" an open circuit at the point of the receiver line junction with the transmission line. The received signal, looking toward the transmitter, sees an open circuit and thus enters the lower resistance of the receiver circuit with little loss.

13-850. Part A of figure 13-240 shows the physical components of the circuit shown in figure 13-239. Part B of figure 13-240 illustrates the electrical relationships involved, and should be compared with the circuit shown in part A of figure 13-239. The

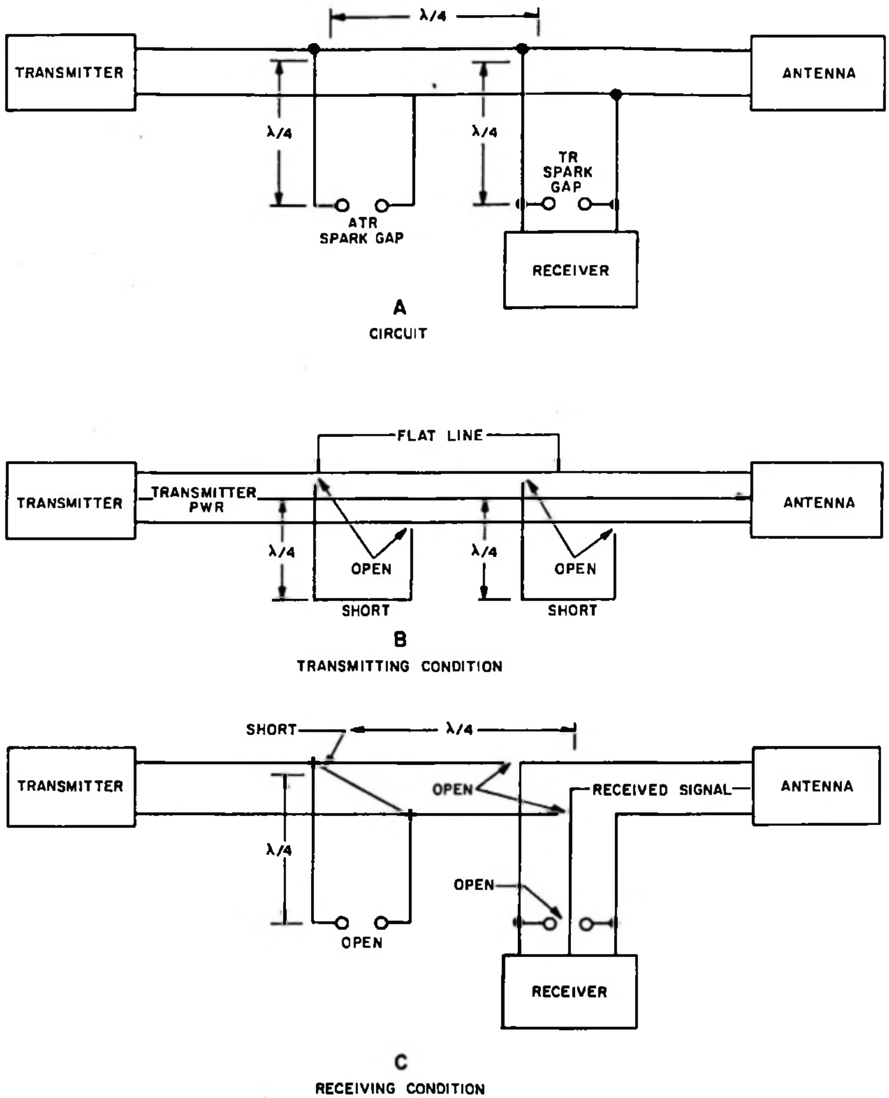


Figure 13-239. TR and ATR Arrangement

tr and atr tubes are placed an odd number of quarter-wavelengths apart, and a quarter-wavelength from the waveguide. When the transmitter fires, both tubes fire. At points A and B in the waveguide sections shown in part B of figure 13-240, the rf signal from an apparent open circuit is reflected. This permits the transmitted power to be conducted to the antenna without loss. Since an electrical short is across the tr tube, the transmitted power does not enter the receiver. During reception, neither the tr tube nor the atr tube fires. The atr section of waveguide now represents an open quarter-wave stub, which has the same effect as a shorted half-wave section. Therefore, it reflects a short to point A of part B in figure 13-240. Since the distance between

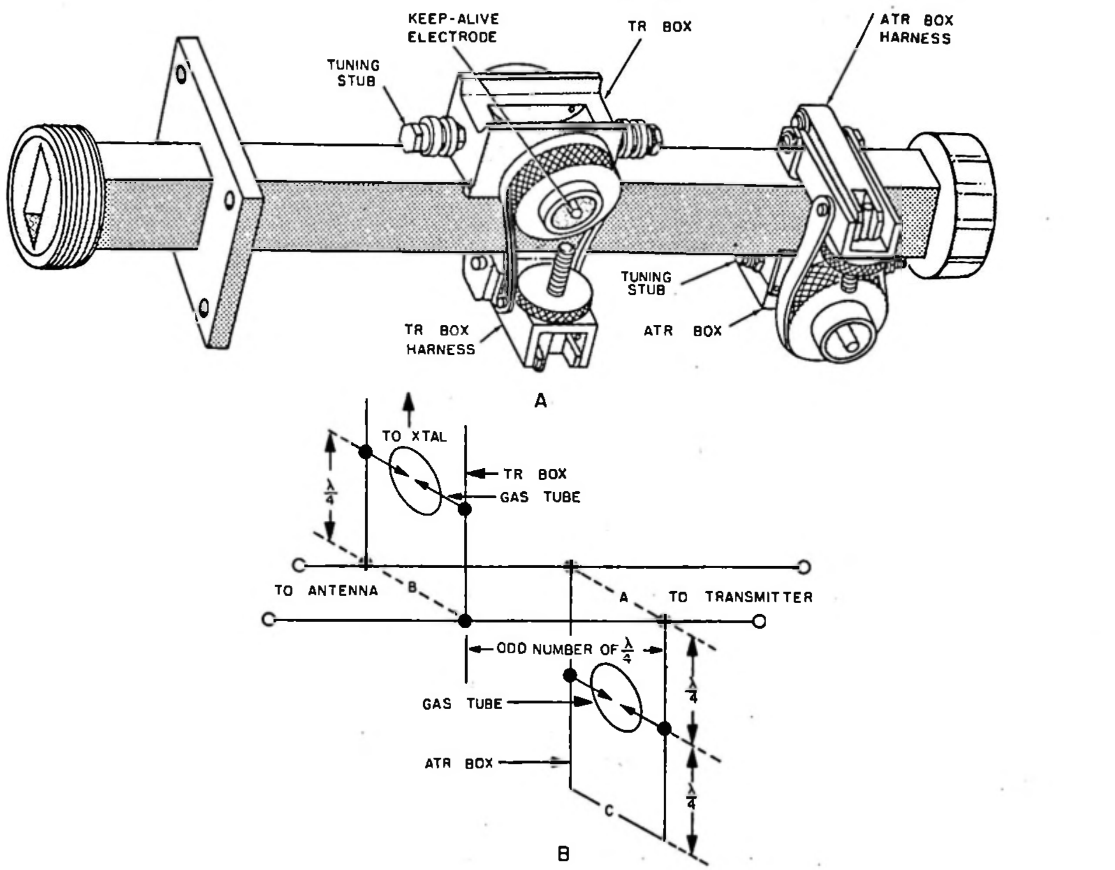


Figure 13-240. TR and ATR Components in Waveguide

points A and B is a quarter-wave, the short at point A is reflected as a high impedance to point B. Point C does not affect impedance values as the "C" distance represents waveguide width. This effectively breaks the line from the receiver junction toward the transmitter, and prevents received energy from entering the transmitter and being dissipated. Therefore, the received signal passes into the receiver.

13-851. SERIES-PARALLEL DUPLEXER.

13-852. The action of the series-parallel duplexer is similar to that of the parallel duplexer except that the atr tube is placed in series with the transmitter and the tr tube is placed in parallel with the transmitter. Figure 13-241 illustrates the series-parallel duplexer. The atr circuit is in series with the transmitter, and is placed one wavelength from the junction of the tr circuit. The tr tube, as before, is a quarter-wavelength from the junction with the transmission line. During transmission, both the tr and atr tubes are ionized. A short circuit is seen at the opening of the atr box, which effectively seals the opening, and the transmitted energy passes on down the line. The tr tube is now a short circuit, which reflects a high impedance at the junction with the transmission line. The transmitted energy passes on down the line to the antenna. During reception, the atr tube does not ionize; therefore, it acts as a shorted quarter-wave stub. A high impedance is reflected at point A in part A of figure 13-241. As the distance from point A to

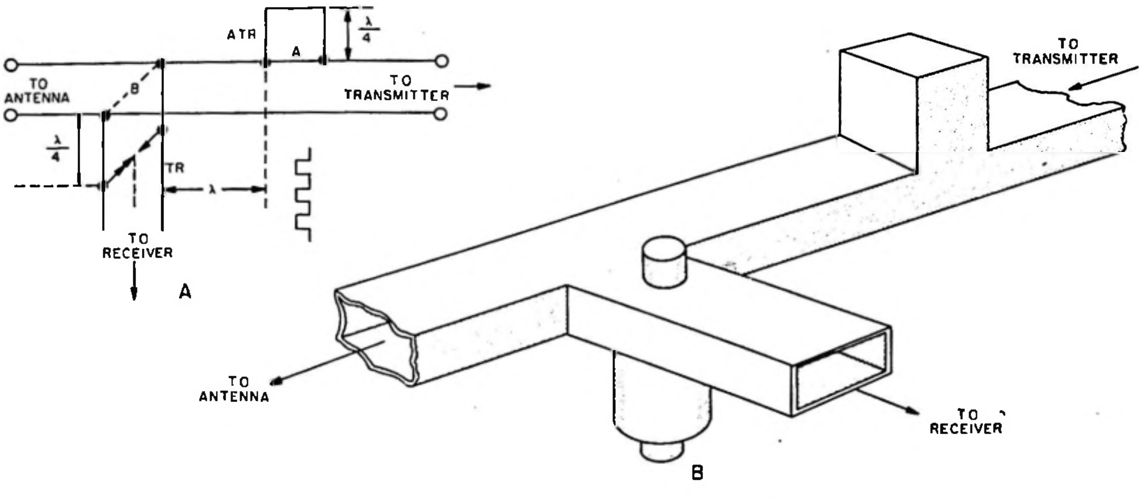


Figure 13-241. Series-Parallel Duplexer

point B is a full wave, a high impedance is reflected at point B in part A of the figure. The received energy is prevented from entering the transmitter section and thus passes into the receiver section.

13-853. TR TUBES.

13-854. A spark gap is not ideal in its operation. When not fired, it acts as a high impedance; when fired, it acts as a nonlinear resistance with a low voltage across itself. The characteristics of a spark gap change with use. To increase its efficiency, the spark gap is enclosed in a partially evacuated glass envelope and is connected to the transmission line through a resonant cavity. Figure 13-242 illustrates the tr tube and the connection to the waveguide. The purpose of the tr tube is to prevent the transmitted energy from entering the receiver. If the transmitted energy entered the receiver, the crystal mixer of the receiver would be damaged severely or would be completely burned out. To protect the crystal mixer, the tr tube must fire, or ionize, immediately after the transmitter fires. The majority of tr tubes are filled with a mixture of argon gas and water vapor. To aid them in ionization, a negative voltage is applied to an electrode placed near one of the gap electrodes. A tr tube showing the keep-alive electrode is illustrated in figure 13-243.

13-855. The negative voltage maintains a steady glow discharge that provides a continuous supply of ions and free electrons. Thus, when rf energy from the transmitter is applied to the electrodes, the tube fires more quickly. In this manner the keep-alive electrode acts as a cathode, while the back of one of the other electrodes acts as the anode. There are different requirements for different radars, but a typical example is that the tr tube must fire, or ionize, within .01 microsecond after application of the transmitted pulse. By the same token the tube must de-ionize 6 microseconds after termination of the transmitted pulse. This interval is called the recovery time. To prevent the crystals in the receiver from being damaged when the radar set is turned off, some radar sets are provided with a

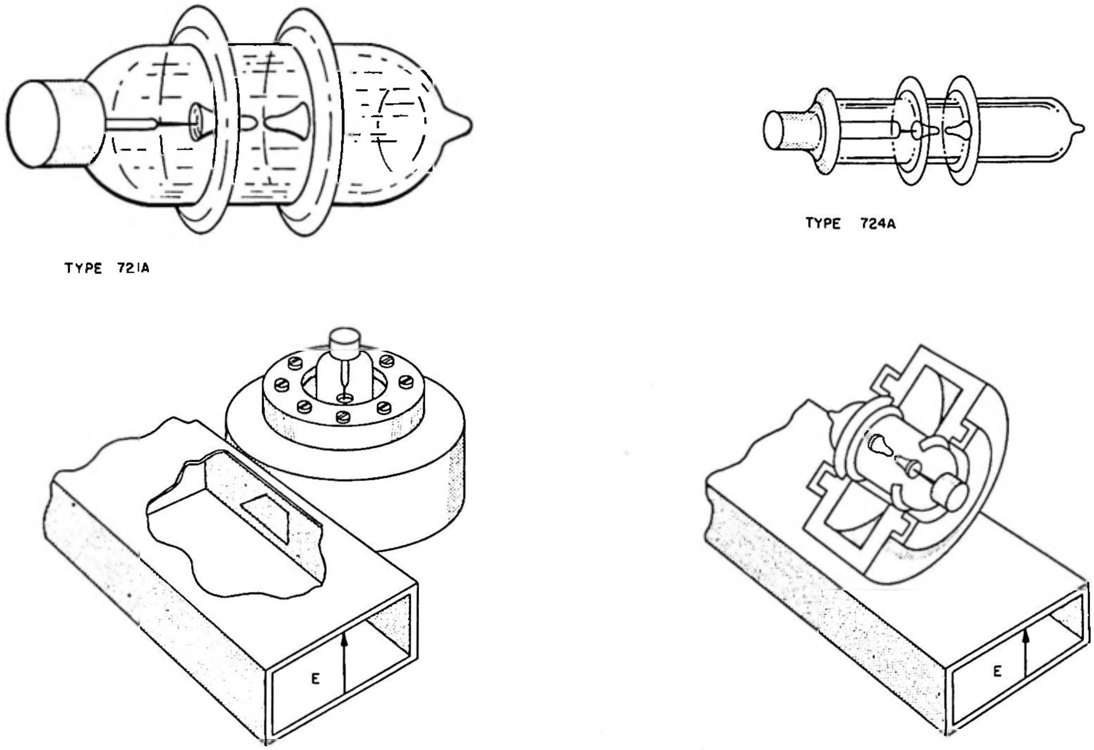


Figure 13-242. TR Tubes

crystal shutter, or gate. This normally consists of an automatically positioned metal vane across the waveguide when the set is turned off.

13-856. ATR TUBES.

13-857. When an atr tube is used in parallel with the transmitter, it is generally of the same construction as the tr tube, but it has no keep-alive voltage applied. When the atr device is used in series with the transmitter, it is of different construction than the tr tube. Figure 13-244 illustrates the 1B35 atr box, which actually replaces a portion of the broad side of the waveguide. The atr box consists of a cavity filled with argon gas and water vapor. During transmission, the gas within the box ionizes and, in effect, completes the section of the waveguide that it replaces.

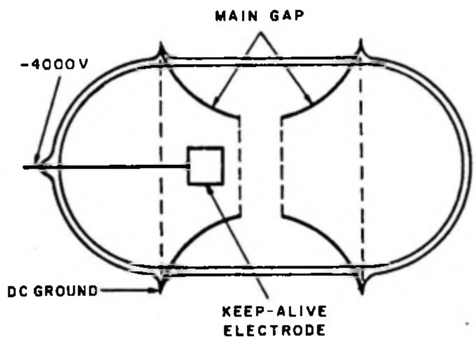


Figure 13-243. TR Tube Showing Keep-Alive Electrode

13-858. POLARIZATION-SHIFTING DUPLEXER.

13-859. One type of duplexer used with a number of radar equipments is termed a

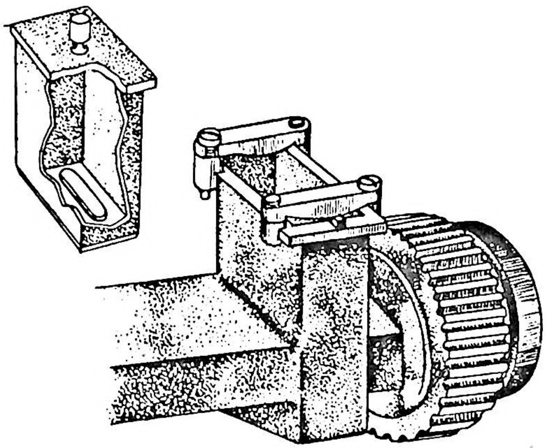


Figure 13-244. 1B35 ATR Box and Mounting

polarization-shifting duplexer (see figure 13-245). In this particular type of duplexer the rectangular waveguide at the antenna end of the duplexer is oriented with its transverse dimensions 90 degrees from those at the end of the duplexer nearest the transmitter. The receiver arm is oriented with its electric vector parallel to the electric vector in the waveguide through which the received signal arrives at the antenna. Both the transmitted and received signals are carried in a rectangular waveguide and arrive at the duplexer with a $TE_{0,1}$ mode of transmission, that is, with the electric vector in the rectangular waveguide at all times parallel to the short dimension of the waveguide cross section. Energy being propagated toward the antenna on transmission has its polarization shifted 90 degrees in passing through the duplexer, and readily continues in the $TE_{0,1}$ mode through the waveguide to the antenna. It does not couple into the receiver arm, except to a limited degree, because the polarization of electric vectors in the transmitted wave and the receiver arm differ by 90 degrees. The energy of the received wave is coupled into the receiver arm, but cannot pass through to the transmitter because of the transverse orientation of the waveguide at that end and because of the action of the atr device.

13-860. The duplexer consists of a section of circular waveguide which is converted at each end to the rectangular section waveguide by transition sections. Within the duplexer (that is, the cylindrical section), 20 gas-filled quartz tubes are placed at regular intervals along the axis of the cylinder. Each is positioned on a diameter of the cylinder. Four of these tubes (the first four encountered from the transmitted signal input end) are all parallel to the long dimension of the rectangular section waveguide at that end. These are shown vertically in figure 13-245. The fifth quartz tube is placed at an angle of 5.63 degrees counterclockwise from the first four (as viewed from the transmitter end). Each of the remaining 15 tubes is turned an additional 5.63 degrees ccw (counterclockwise), so that the last tube is at right angles to the first four. The transmitted wave enters the duplexer transition in the $TE_{0,1}$ mode; the electric field is parallel to the short dimension of the waveguide. Since the duplexer is circular in cross section, the wave proceeds in the $TE_{1,1}$ mode, but the orientation of the transmitted energy is maintained, assisted by transverse shorting bars inside the transition section, near the end of the duplexer. Just beyond the branch arm for the received signal the transmitted energy strikes the first of the gas-filled quartz tubes. These tubes are filled with argon under pressure. The high-power rf energy ionizes the argon molecules, breaking down the initial resistance of the tube and causing it to conduct. This forms a very low-impedance path across the cylindrical duplexer, and the first four "in-line" tubes perform essentially the same function as metal shorting rods. They tend to maintain the existing polarization of the transmitter wave.

13-861. The high-power wave now meets the fifth tube, which has an angular dis-

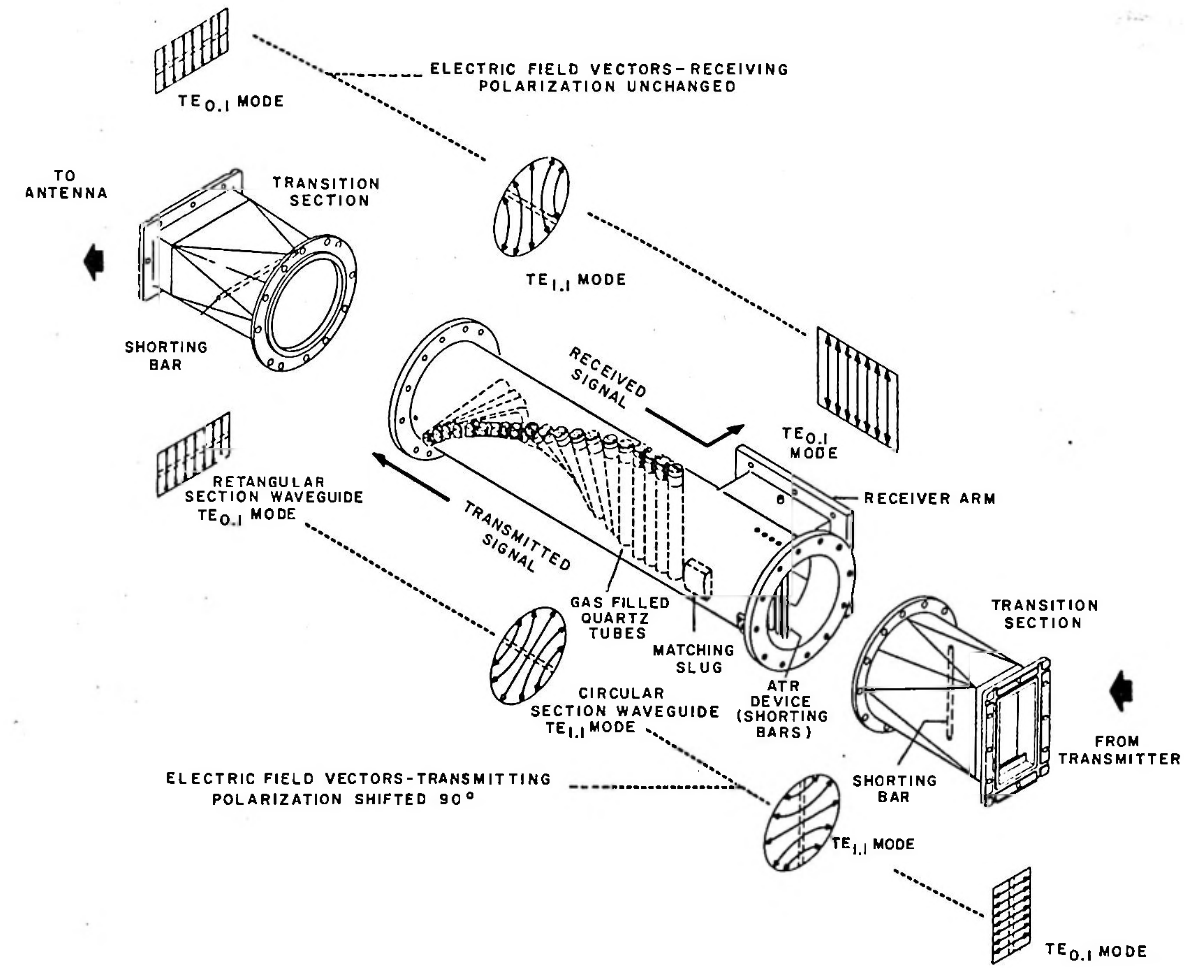


Figure 13-245. Polarization-Shifting Duplexer

placement of 5.63 degrees from the first four. The direction of the E vectors is shifted by this amount, and the rotation will continue through the 16 tubes until the transmitted wave arrives at the twentieth tube. The polarization has been shifted through a full 90 degrees; this means that the polarization is at right angles to that of the energy entering the duplexer. The net steady-state result of field interaction within the "fired" duplexer is a nearly uniform axial conducting surface twisting through 90 degrees. The purpose of the matching slug on the side of the duplexer opposite the opening into the receiver arm is to compensate for the effect upon the transmitted wave of the discontinuity in the walls of the duplexer caused by the opening into the receiver arm. The received signal from the antenna arrives at the duplexer with a polarization determined by the orientation of the rectangular waveguide through which it has passed. It is too weak to fire the gas-filled quartz tubes placed spirally across the duplexer. Thus, they do not function as they do in transmission, and the received wave passes through them with its polarization unchanged.

13-862. For two reasons the received wave is coupled into the receiver arm. First, the orientation of the rectangular waveguide of the receiver arm is aligned with the polarization of the incoming wave, so that the wave can proceed into the receiver arm without serious attenuation. Second, the received wave is prevented from passing directly through the duplexer toward the transmitter by the five shorting bars across the duplexer just beyond the receiver arm. These bars serve to reflect the received signal and to set up standing waves with a current minimum at the receiver branch leading to the tr cavity. This provides maximum coupling of the received signal into the tr cavity. These shorting rods, therefore, constitute an atr device. Figure 13-245 illustrates a typical polarization-shifting duplexer.

13-863. COMPLETE WAVEGUIDE EQUIPMENT.

13-864. GENERAL.

13-865. In a complete waveguide rf equipment the signal source is the transmitter unit. The antenna that transmits the output is located as near as possible to the radar transmitter. A principal requirement in an efficient radar facility is that energy must be transferred from the transmitter to the antenna with minimum losses, because any losses that occur will shorten the maximum range and thereby decrease the efficiency of the entire radar facility. Since the transmitting antenna is pointed at the target at the time the transmitter pulse leaves, and since it is conveniently located, it is preferable to use it as the receiving antenna also. Similarly, it is possible to use the same waveguide arrangement for transferring the received energy to the radar receiver. When the same elements are used for both transmitting and receiving, it is necessary to use a high-speed automatic switch to prevent the powerful transmitter pulse from going into the receiver circuit and causing serious damage. The tr and atr boxes discussed earlier perform this function in many radar sets. The highly directional antenna is made rotatable so that it will cover a horizontal 360-degree area. To make rotation possible, signals are coupled to the antenna through vertical and horizontal rotating joints.

13-866. Since airborne equipment is sometimes operated at high altitudes, and some ground radar facilities are also located at either high altitudes or in high-humidity areas, temperature changes will cause moisture condensation inside the waveguide. As moisture will cause extremely high radiation loss, the interior of the waveguide must be maintained at a higher pressure than the exterior to drive out the moisture and keep it out. Furthermore, all joints have to be airtight to maintain this pressure. Figure 13-246 shows a waveguide equipment that uses the components just discussed. The source of the rf energy is the magnetron tube shown at the right in the lower part of the figure. A probe couples it to the waveguide, which operates in the $TE_{0,2}$ mode. A very short section of waveguide is fixed permanently to the magnetron output. When the magnetron is installed, its waveguide section is connected to the end of the main waveguide. In operating conditions the transmitter energy travels down the main waveguide and enters a section which has two T-junctions—one called the atr box, and the other the tr box and mixing chamber.

13-867. In each T-junction, there is a spark gap at a distance of one quarter-wavelength from the center of the main guide. Both of these gaps arc-over when the transmitter is on. Thus, no energy goes past either spark gap tube into the T-junction waveguides. The energy flow is continuous down the main guide to the antenna. Each section is connected to the next with a choke joint. In going to the antenna, the signal

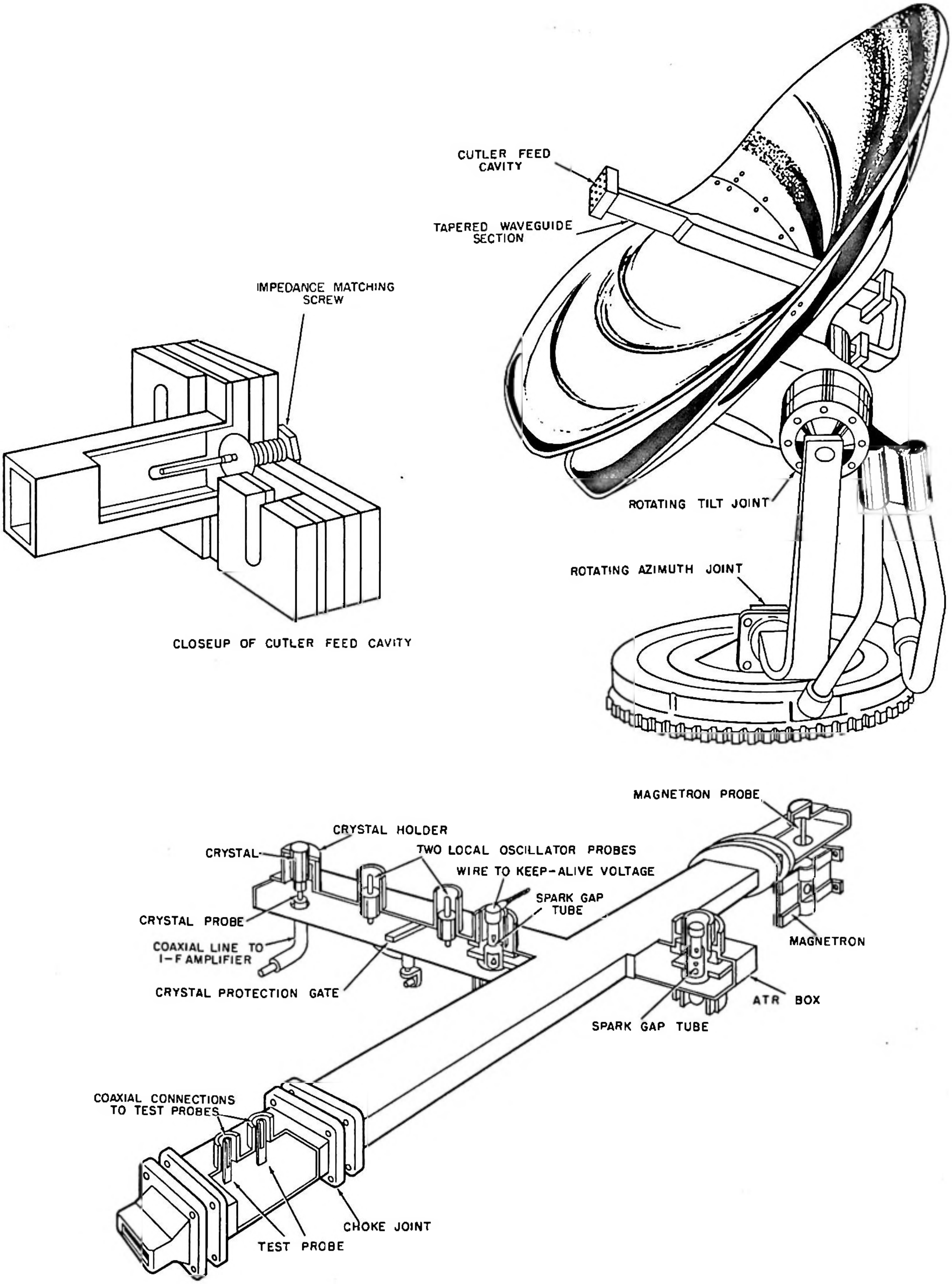


Figure 13-246. Complete Waveguide Equipment

passes through a special section that contains a pair of probes spaced one quarter-wavelength apart. Test equipment can be inserted at each probe to measure the signal strength at that probe. The ratio of the two signal strengths is the standing-wave ratio. In going through the stationary transmitter unit to the moving antenna, the signal goes through a flexible section of waveguide that insulates the transmitter unit from the mechanical vibration of the antenna. The antenna is designed to rotate horizontally and to tilt vertically. Thus, the signal is transferred to a rotating coaxial joint twice before being radiated, and each time it is immediately transferred back to a waveguide again. The antenna itself is a resonant cavity containing a pair of slots one half-wavelength apart. As you can see in the figure, the waveguide is tapered in the narrow dimension to fit between the slots. The effect of the cavity is as if the waveguide were split in half and each side bent through 180 degrees. Tapering the waveguide prevents the reflections that would occur if the size were changed abruptly. Tapering in the narrow dimension does not affect the operation of the waveguide, since this dimension is not critical. The actual impedance match is accomplished by adjusting the matching screw. This adjustment is made at the factory and is soldered in place.

13-868. The energy comes out of the back of the cavity and goes to the large reflector. The reflector projects the energy forward in a narrow beam. The energy which returns from the radar target is reflected by the reflector into the slots. It passes through the slots into the waveguide and travels toward the magnetron, where it stops at the atr box. The length of the T-junction is one half-wavelength from the center of the waveguide; thus, the closed end of the section reflects a short circuit at the outer-center of the main waveguide. This reflects the signal back along the guide to the mixing chamber. Of course, neither spark gap is fired because the signal is too weak to produce an arc. The signal enters the mixing chamber, where it strikes the end wall, and reflections are again set up. The chamber becomes a cavity resonator. An additional signal is introduced from another oscillator—the heterodyne oscillator for the receiver. The signal is injected with a probe at the center of the E-field in the cavity. Both signals cause current flow through a pickup probe one quarter-wavelength from the end of the cavity. This probe is connected to a crystal mixer. The second oscillator is set at a different frequency, and produces the correct intermediate frequency when the returning signal is from a radar beacon transmitter. Since the crystal is easily damaged by high-powered signals, the tr box spark gap protection must be certain. To insure that it will arc-over at once when the transmitter signal appears in the mixing chamber or when other nearby radar transmitters accidentally send in a weaker—but still damaging—signal, a keep-alive voltage is provided. This is a high dc voltage that causes the spark gap to be partially ionized. Then the additional voltage of any signal that is stronger than a radar echo will cause complete ionization, or an arc, between the electrodes. When the set is turned off, no keep-alive voltage is present, and the crystal is vulnerable to signals from other nearby radar transmitters. A spring-operated gate closes the waveguide when the set is turned off. No signal can get to the crystal from the antenna. When the set is turned on, an electrically operated relay opens the gate. The entire equipment is assembled with rubber gaskets at the joints, or is otherwise sealed to permit pressurizing the interior of the waveguide. The small windows or slots in the antenna cavity are covered with mica. The mica permits rf energy to pass through, but keeps moisture out.

13-869. WAVEGUIDE ISOLATORS.

13-870. A device which is very useful in waveguide transmission is the so-called isolator. This is a waveguide element which effectively permits transmission through a guide in one direction only. It can thus be used to prevent energy, reflected from a transmitting antenna or other discontinuity, from traveling backward through the guide to points where it might adversely affect the operation of vital circuit components, such as klystron oscillators and traveling-wave tube amplifiers. The non-reciprocal behavior of isolators depends upon certain peculiar properties of the magnetic ferrites. Because of their extremely high resistivity, ferrite materials do not normally have any appreciable effect on the propagation of high-frequency electromagnetic waves. If inserted within a waveguide, they are practically transparent to microwave energy traveling through the guide. In the presence of an externally applied dc magnetic field, however, the behavior of the ferrite is quite different. Within the ferrite, because of its unique magnetic structure, there is now definite interaction between the applied field and the magnetic field of the traveling microwave energy. This effectively changes the permeability of the ferrite, which results in some distortion or displacement of the electric field of the traveling microwave energy. Moreover, the interactions are such that the permeability change in the ferrite is different for the two directions of microwave transmission. The electric field displacements, accordingly, are also different.

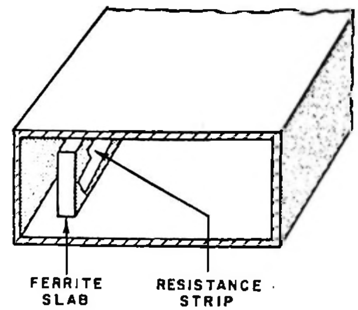


Figure 13-247. Field Displacement Isolator

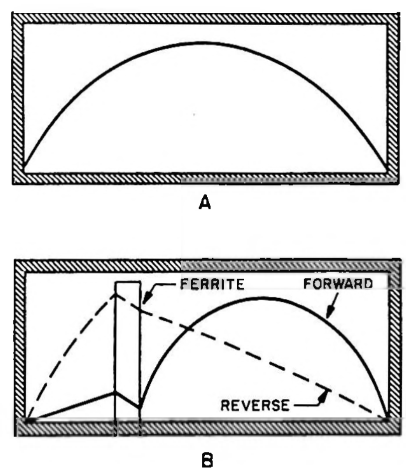


Figure 13-248. Electric Fields in Rectangular Waveguide

13-871. Figure 13-247 shows a section of rectangular waveguide in which a thin slab of ferrite, which is typically a few inches in length, is placed in an off-center position. It is assumed that a dc magnetic field is applied to the ferrite in a direction at right angles to the magnetic field of the traveling microwave energy. Such an arrangement will displace the electric fields of forward and reverse traveling microwave energy in some manner such as that indicated in figure 13-248. Part A of the figure shows the normal electric field intensity distribution in a cross-section of rectangular waveguide, and part B shows the way in which this field is displaced by the magnetized ferrite for both forward and reverse waves. You should note that the field intensity of the forward wave is minimum at the ferrite slab, while the intensity of the reverse wave is maximum at the same point. Now, if a coating of resistive material such as graphite is applied to one side of the slab, as shown in Figure 13-247, it will absorb most of the energy of the reverse wave. The forward wave, on the other hand, will be attenuated only slightly by the resistive coating because of its low field intensity at that point. The magnitude of the discrimination between forward and reverse waves depends rather critically on such factors as the length and width of the ferrite slab, its location with respect to the guide wall, the strength of the applied magnetic field, and particularly the length, shape, and exact location of the resistive strip. In practice, representative losses introduced by field displacement isolators of this type are about 0.2 db in the forward direction of transmission and 30 to 35 db in the reverse direction.

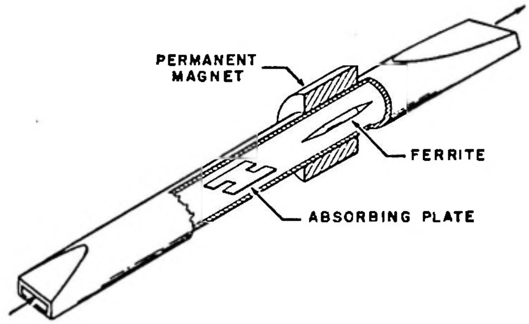


Figure 13-249. Faraday Rotation Isolator

13-872. FARADAY ROTATION ISOLATOR. In a circular waveguide, you may take advantage of another special property of magnetized ferrite to effect microwave isolation. When a short rod or pencil of ferrite material is centered in a circular waveguide and magnetized in a longitudinal direction by an external magnet, the plane of polarization of transverse electric microwave energy in the guide is rotated to a new angular position as it passes the ferrite. The extent of the rotation is accurately determined by the size and shape of the ferrite pencil and the strength of the applied magnetic field. Of particular interest, moreover, is the fact that the field rotation is in the same angular direction, regardless of the direction in which the microwave energy is traveling through the guide. This remarkable behavior is called "Faraday rotation" because of its close analogy to a comparable optical phenomenon first noted by Michael Faraday more than a century ago.

13-873. How this Faraday rotation effect is employed in microwave practice to secure isolation may be understood by reference to figure 13-249. Here a thin pencil of ferrite, tapered at both ends to minimize reflections, is held centered in a section of circular waveguide. An axial magnetic field is established in the ferrite pencil by means of a permanent magnet surrounding the section of waveguide, as shown. It is assumed here that the dimensions of the ferrite and the strength and direction of the magnetic field are such that the polarization of a wave in the guide will be rotated exactly 45 degrees in a counterclockwise direction, as viewed from the left end of the guide. At each end, the waveguide is transformed from circular to rectangular cross section by a gradual transition that prevents impedance irregularity. The transition at the right end, however, is so made that the rectangular guide there is turned through 45 degrees with respect to the rectangular guide at the left end.

13-874. VERTICAL POLARIZATION. From the general discussion of waveguide transmission, you will recall that for normal $TE_{1,0}$ transmission in a rectangular guide, the electric field lines always extend between the longer walls and are thus parallel to the shorter walls. In figure 13-249, this configuration may be designated as vertical polarization of the electric field. In a circular waveguide, on the other hand, the dominant mode, $TE_{1,1}$, may be transmitted equally well in any plane of polarization. If the desired direction of transmission in figure 13-249 is from left to right, a signal entering the guide at the rectangular left end will necessarily be vertically polarized. As it passes through the transition into the circular guide, it will continue to be polarized vertically. Passage through the ferrite, however, will rotate the polarization 45 degrees counterclockwise. But since the rectangular guide at the right end is also turned through 45 degrees, the rotated wave will pass into the rectangular guide at the right end still in the vertical plane with respect to the guide. Transmission is thus effected without appreciable attenuation of the wave. However, if the transmitted wave is partially or completely reflected by some irregularity farther along in its path, the reflected wave will encounter a different set of conditions. It will return still vertically polarized until it reaches the ferrite. Here it will again be rotated through 45 degrees, but also in a counterclockwise direction as viewed from the sending end. Its plane of polarization is nowturned exactly 90 degrees from that of the original input wave. The rectangular guide cannot transmit a wave in this horizontal polarization; therefore, the wave must be totally reflected when it reaches the transition. If desirable, this second reflection can be largely avoided by inserting a plate of energy-absorbing material between the ferrite and the transition, as shown in the drawing. This may consist of a small card of insulating material which has been given a thin coating of a "lossy" substance, such as carbon or aquadag, that readily soaks up microwave energy. Mounted in a horizontal plane, as indicated, the card causes high attenuation of the reflected wave while offering practically no opposition to the transmission of the vertically polarized forward wave.

13-875. In practice, it may not always be feasible to control the rotation of the wave polarization to the precise angle required for 100 percent isolation. A simple device such as that pictured here, however, may be expected to introduce a loss of only about .25 db in the desired direction of transmission, and from 30 to 35 db in the opposite direction. When more complete isolation is desired, any of several arrangements employing two or more ferrite units in tandem can be used.

13-876. HIGH-SPEED WAVEGUIDE SWITCHES. The Faraday rotation effect is also used to produce the high-speed waveguide switches used in certain microwave radio equipments. Here the wave rotation is controlled by an electromagnet rather than a permanent magnet, and the magnetization is of such a value as to rotate the wave through 90 degrees. When you insert a section of circular guide so equipped in a rectangular guide, as shown in figure 13-250, it has no appreciable effect on transmission as long as the electromagnet is not energized. Simply closing the circuit through the electromagnet results in complete blocking of transmission through the guide. Thus, two such devices make possible almost instantaneous switching of the wave transmission from one path to another.

13-877. MICROWAVE MEASUREMENTS AND CALIBRATION PROCEDURES.

13-878. GENERAL.

13-879. The ultimate purpose of all microwave plumbing installations is to convey intelligence in the form of electromagnetic signals from one point to another. Different applications require different types of signals and equipment in order to obtain a given desired result. Many problems arise from

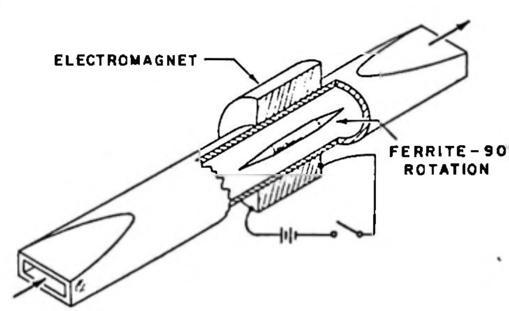


Figure 13-250. Faraday Rotation Waveguide Switch

the particular requirements of each micro-wave arrangement. These can, in general, be solved by the careful design of special components. Such components may be any-thing from a straight section of waveguide to a complicated structure such as an rf unit, which consists of a duplexer, tr and atr switch tubes, a local oscillator, a mixer, a beacon oscillator, and automatic frequency control provisions. Each component must be mechanically and electrically tested and, at times, carefully adjusted before you final-ly mount the component in its place in the equipment. This procedure calls for a great degree of flexibility in the test equipment and of the test procedures employed in order to achieve accurate and significant results in a number of different types of measure-ments. A better understanding of the prob-lems arising in connection with microwave measurements can be obtained by an analysis of a typical microwave arrangement, for example, 3-cm radar plumbing. Figure 13-251 shows such an arrangement, com-prising a magnetron, an rf unit, a direc-tional coupler, a transmission line with several bends and twists, a rotating joint, and an antenna. Each component has a definite function for which it was designed and tested. In general, we can say that a radar set functions by beaming a high level of rf energy in a definite direction and by receiving the faint echo reflected from any target intercepting this beam. The echo is so weak that it requires an enormous amount of amplification before it can be used for lo-cating a target. The intensity of the echo is proportional to the transmitted power; thus, the elimination of all losses in the transmis-sion line which connects the antenna to the

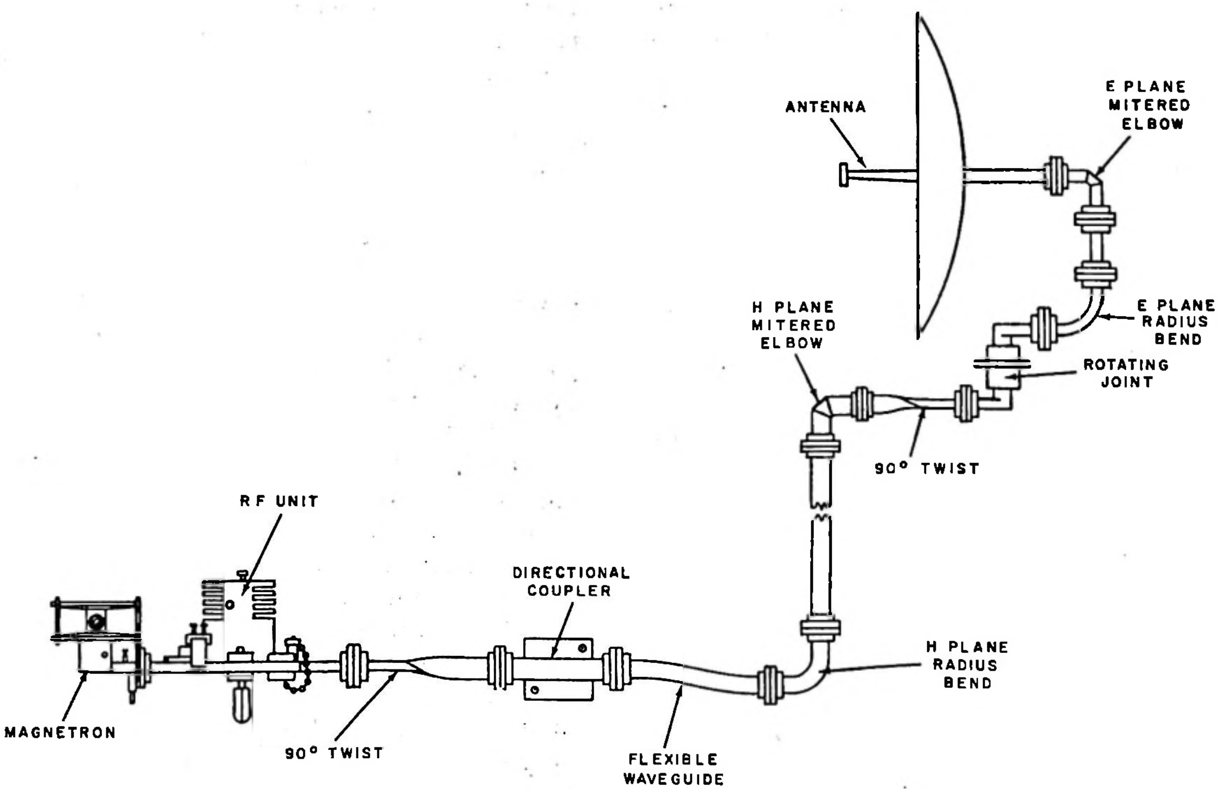


Figure 13-251. Typical Microwave Arrangement

generator and the receiver is of great importance. These losses, if not properly reduced, are doubly detrimental to the efficiency of the transmitting function in that they cut down the power available at the antenna for transmission and, at the same time, reduce the intensity of the originally weak return echo. Losses result from two principal causes: reflection and skin effect. The skin effect is distributed along the whole length of the transmission line, and can be reduced if you properly plate the interior of the waveguide and all other surfaces exposed to the high-frequency fields. Plating results in an attenuation on the order of a few hundredths of a decibel per foot. It is, therefore, almost negligible in the case of short runs of waveguide, but it may become quite appreciable in the case of long transmission lines. Thus, you should test long sections of waveguide for attenuation characteristics.

13-880. Reflections result from discontinuities and mismatch in the line. The amount of power lost as a consequence of reflections is usually much higher than that lost by attenuation; consequently, all components must be tested for reflections before they are assembled in the over-all arrangement. You should test the line, as a whole, after assembly. Usually, the total reflection is much less than the sum of all individual reflections, because these, by combining with different phases, tend to cancel out one another. If you know the reflection coefficient of each component, in terms of amplitude and phase, you can design the transmission line so that all reflections do cancel each other. The result is a reflectionless line, made with not quite reflectionless components. The importance of a very accurate measurement of the reflection coefficient (including phase and amplitude) is self-evident. Reflections must not only be small, but also constant. It is thus important for you to test, under actual operating conditions, all those components where a change in reflection might occur because of some change in shape or position (for example, the rotating joint shown in figure 13-251). The same applies to the flexible waveguide. Such components should be tested for reflection in a number of different configurations, selected as being representative of all the possible configurations that the component may assume. After you assemble the transmission line and put it into operation, it is desirable to provide a means for detecting any accidental change in the reflections in the line which may result from corrosion in the waveguides, obstructions in front of the antenna, water condensation, etc. It is also desirable to provide a means for monitoring the output power. Both of these functions are accomplished by the directional coupler, shown in figure 13-251. A directional coupler is a device which extracts a very small amount of energy traveling down the main waveguide and couples it to a secondary waveguide, where it can be separately measured. The most important feature of the directional coupler is the fact that energy traveling in only one direction is coupled out, while energy traveling in the opposite direction has almost no effect on the output of the secondary guide. Two important values must be known before using a directional coupler: the coupling, C (ratio of output power in the secondary guide to input power in the main guide in the preferred direction), and the directivity, D (ratio of output power for input in the preferred direction to output power for the same input in the opposite direction). Thus, measurements of coupling and directivity are essential in connection with all directional couplers, in addition to the usual measurements of the reflection coefficient. The function of the rf unit is more complicated. It must prevent the high-level transmitted pulse from reaching the sensitive detector crystal and damaging the crystal. It must also convey all of the weak, received pulse to the detector and prevent any percentage of it from deviating into the magnetron and being lost there. This is accomplished by means of gas discharge tubes and use of the geometry of the structures connecting them with the main guide. This involves several types of waveguide transformers, resonant sections, etc.

13-881. The rf unit must also provide a mixer effect, incorporating a local oscillator. The generated frequency is mixed with the arriving signal in the crystal detector, and the resulting output is fed to the i-f amplifier. Another oscillator is coupled to the main guide for beacon operation. The testing of this complex unit requires a means for measuring and comparing waveguide impedances, resonant effects, reactances, etc, as well as the usual attenuation, reflection, and coupling. Since all effects in waveguides generally change with frequency, the frequency must be accurately known while making all types of measurements. Every test setup must, therefore, include a means for measuring the frequency while the test is being made. The example listed above shows the large variety of microwave measurements necessary for properly testing the components of a complete arrangement, such as a radar plumbing arrangement. Special tests for experimental and research purposes may require even more complex measurements. The necessity of a comprehensive classification of all types of measurements is clear. Since all impedance and resonance effects can be determined with measurements of reflections and absorptions (i.e., in terms of wave parameters), a rigorous classification of all types of measurements can be made by observing that the characteristics of the most general electric oscillation depend on three parameters only. As a matter of fact, any generic alternating quantity, A, can be described by the simple equation:

$$A = A_0 \sin(2\pi ft + \phi)$$

where A_0 is the crest amplitude, f is the frequency, and ϕ is the phase angle. The most complicated oscillation can always be split up into a number of elementary oscillations of this form which may be considered separately. In this manner, it is obvious that the determination of A_0, f, and ϕ will give all possible information on the oscillation itself. A practical classification of measurements is as follows:

a. Measurements intended to determine the amplitude of the oscillation.

b. Measurements intended to determine the frequency of the oscillation.

c. Measurements intended to determine the phase of the oscillation.

13-882. The first group may be further subdivided into absolute amplitude measurements and relative amplitude measurements, depending upon whether the information required concerns the total energy present in the oscillation or a comparison of the amplitude of two waves with the same frequency, regardless of their absolute values. Absolute amplitude measurements are usually called power measurements. Relative amplitude measurements may be again subdivided into: attenuation measurements, intended to determine the ratio of the wave amplitude in one point of the equipment to the wave amplitude in another point of the equipment; and reflection measurements, intended to determine the ratio of the amplitude of the wave traveling in one direction, at some point of the equipment, to the amplitude of the wave traveling in the opposite direction at the same point.

13-883. Frequency measurements are always absolute, and phase measurements are always relative; you must define the reference plane in connection with phase measurements. Frequency measurements may be subdivided according to the type of wavemeter employed (either the absorption type or the transmission type). However, this subdivision refers to the physical structure of the equipment only, and not to the measured quantity. Figure 13-252 is a complete table of microwave measurements, showing some of the most important specific measurements currently made in the laboratories. All types of measurements described in connection with our example can now be classified in accordance with this table. Any other measurement can be fitted into the table in accordance with its relationship to the basic types illustrated.

13-884. MICROWAVE TEST EQUIPMENT.

13-885. The test equipment should furnish accurate results readily. The requirement of accuracy is sometimes in conflict with the requirements of ease and speed of procedure, and it is not always possible to reach a workable compromise. In some cases, high speed and simplicity of procedure are the most important requirements, as in the case of production testing. Here a large number of similar components are successively tested and, generally, the only information required is whether or not the components meet a given set of specifications. In other cases, accuracy is of the utmost importance, for example, in the calibration of test equipment components or in experimental projects. In these cases the number of pieces to be tested is usually small, and time and effort are of secondary importance. The test equipment involved and the techniques employed are accordingly different. In any case, it must be kept in mind that the precision of the most accurate instrument is no better than the precision of the measuring method employed. The best instrument may become worthless if improperly used; on the other hand, small procedural refinements may greatly improve its precision. When several instruments are concurrently used for one measurement, the accuracy that can be obtained is no higher than that which can be expected from the least accurate of the group of instruments employed, regardless of the precision of all the others. However, the best results can be obtained if the least accurate instrument is used in connection with equipment of the same quality level

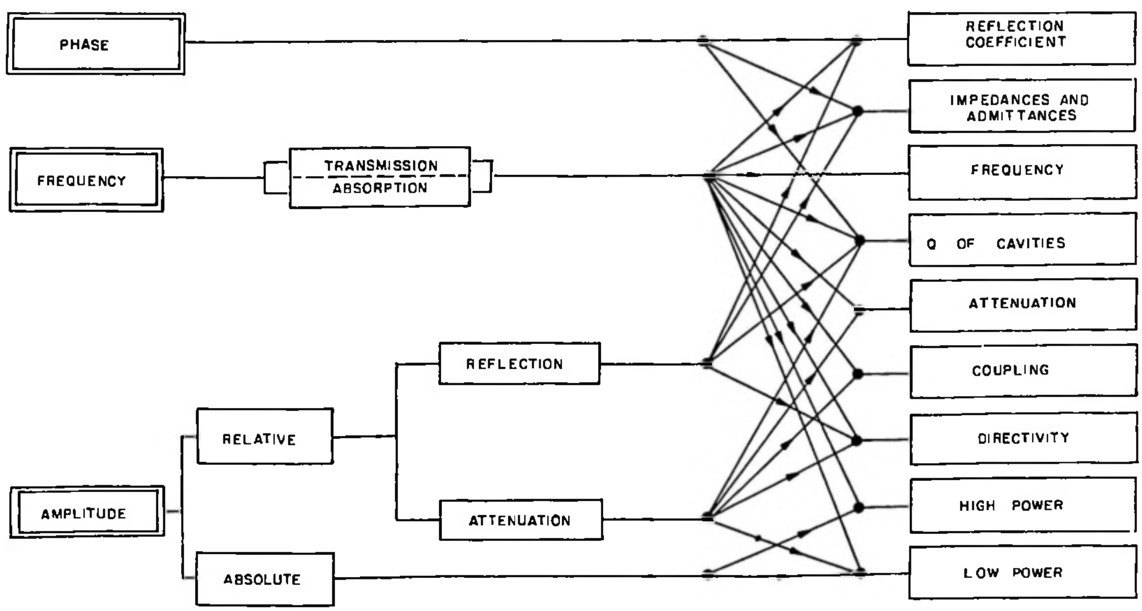


Figure 13-252. Microwave Measurements

(power supplies, oscillator tubes, amplifiers, etc). The test equipment required to perform each type of measurement will be described, together with the proper procedure, in the order indicated in figure 13-252.

13-886. PRINCIPLES OF REFLECTION COEFFICIENT MEASUREMENT.

13-887. The reflection coefficient is a vectorial quantity with absolute value equal to the ratio of the amplitude of the reflected wave to the amplitude of the direct wave. Since a wave does not ordinarily experience total reflection, the reflection coefficient is usually less than one in absolute value. You can define the phase angle of the reflection coefficient only by arbitrarily assuming a reference plane. Whenever possible, the reference plane is set to coincide with the plane where the reflection occurs. This, in turn, is not always possible because sometimes reflections may occur as a result of a gradual change in the properties of the transmission line or where the field distribution in the guide is so complicated that it is difficult to determine just where the reflection occurs. However, once a reference plane is established, the phase angle is defined as the phase difference between the reflected and direct waves in that plane. From this point on, the voltage gradient or electric field strength, E (in volts per meter), will be assumed to represent the amplitude of the waves.

13-888. Assume that you write the direct wave in a lossless line with the usual complex notation, as follows:

$$E_1 = E_0 \, e^{j\left(\omega t - \frac{2\pi}{\lambda_g} z\right)}$$

where E_0 is the crest amplitude, ω is the angular frequency ($\omega = 2\pi f$), λ_g is the guide wavelength, and the wave is propagating in the positive Z direction. You can write the reflected wave as follows:

$$E_2 = E_r \, e^{j\left(\omega t + \frac{2\pi}{\lambda_g} z + \phi\right)}$$

where E_r is the crest amplitude of the wave propagating in the negative Z direction. By definition, the reflection coefficient, T, is the ratio of E_2 to E_1; that is:

$$\Gamma = \frac{E_2}{E_1} = \frac{E_r}{E_0} \, e^{j\left(\frac{4\pi}{\lambda_g} z + \phi\right)}$$

Thus, it appears that Γ is a vector of constant amplitude which makes two complete turns in the complex plane for every full wavelength of displacement of the reference plane in the Z direction. The modulus of this vector is equal to the ratio of the crest values (E_r to E_0), and the argument reproduces itself every half wavelength of travel in the Z direction. At the origin (z = 0) the argument is equal to ϕ which is seen to be the phase angle as above defined.

13-889. The total amplitude where both the direct and reflected waves are present is given by the sum of their amplitudes; that is:

$$E_t = E_1 + E_2$$

If you substitute values, you will have the following:

$$E_t = E_0 \, e^{j\left(\omega t - \frac{2\pi}{\lambda_g} z\right)} + E_r \, e^{j\left(\omega t + \frac{2\pi}{\lambda_g} z + \phi\right)}$$

You can rewrite this equation as follows:

$$E_t = E_0 \left[e^{-j\frac{2\pi}{\lambda_g} z} + \frac{E_r}{E_0} e^{j\frac{2\pi}{\lambda_g} z + j\phi} \right] e^{j\omega t}$$

You can see that this a vector rotating with angular velocity ω, with the amplitude depending on position z and phase angle ϕ only. The maximum amplitude will occur where both components have the same argument, that is, when:

$$\frac{2\pi}{\lambda_g} z + \phi = -\frac{2\pi}{\lambda_g} z$$

or:

$$\frac{4\pi}{\lambda_g} z = -\phi \qquad (1)$$

The minimum amplitude will occur where the arguments differ by π, that is, when:

$$\frac{4\pi}{\lambda_g} z = \pm \pi - \phi \qquad (2)$$

The amplitude at the maximum point is:

$$E_{max} = E_0 (1 + |\Gamma|)$$

The amplitude at the minimum point is:

$$E_{min} = E_0 (1 - |\Gamma|)$$

The ratio of the maximum value to the minimum value is called the voltage standing wave ratio, and is given by:

$$Vswr = \frac{1 + |\Gamma|}{1 - |\Gamma|}$$

Conversely:

$$|\Gamma| = \frac{vswr - 1}{vswr + 1} \qquad (3)$$

Thus, the amplitude and phase angle of the reflection coefficient can be determined with the aid of equations (1), (2), and (3), by measuring the ratio of maximum field value to minimum field value and by determining the distance from the reference plane where such maxima and minima occur. Both measurements are readily made with the slotted line and probe, which provide a means of sampling the field intensity of the waves at different points along the waveguide and, at the same time, of accurately determining the position where the field is probed. Reference is specifically made to the X-band equipment with the understanding that the same procedure may be used on any band, simply by using components designed to perform equivalent functions at different frequencies. Specific differences will be separately indicated. The same slotted line and probe can be used both for induction testing and for high-precision measurements. The test setup and the test procedures will differ in the two cases because of the different requirements and specifications. Two types of test setups and procedures are recommended in order for you to operate the slotted line and probe at optimum efficiency by fully using its precision construction and design flexibility. The test setup recommended for production testing is illustrated in figure 13-253.

13-890. A signal generator feeds through rf cable to a coaxial-to-waveguide transformer. A variable flap attenuator is inserted immediately after the tube mount, for padding the generator, and is followed by the frequency meter. Next in line is the slotted line and probe, connected to the vswr indicator by a flexible coaxial cable. The component to be tested is inserted between the slotted line and probe and a moving load. The rectified signals proceeding from the crystal in the probe are amplified and displayed by the indicator, which is calibrated directly in vswr. Since an ac amplifier is much more reliable and stable than a dc amplifier, the crystal output must be modulated for proper amplification. This is done by modulating the output of the oscillator tube with a voltage square wave applied to the reflector electrode of the tube in the signal generator.

13-891. It is known that the power output and the frequency of a reflex klystron depend critically on the reflector voltage. A typical output characteristic is shown in figure 13-254. When you increase the reflector voltage, the tube enters successively into a number of modes of oscillation separated by gaps

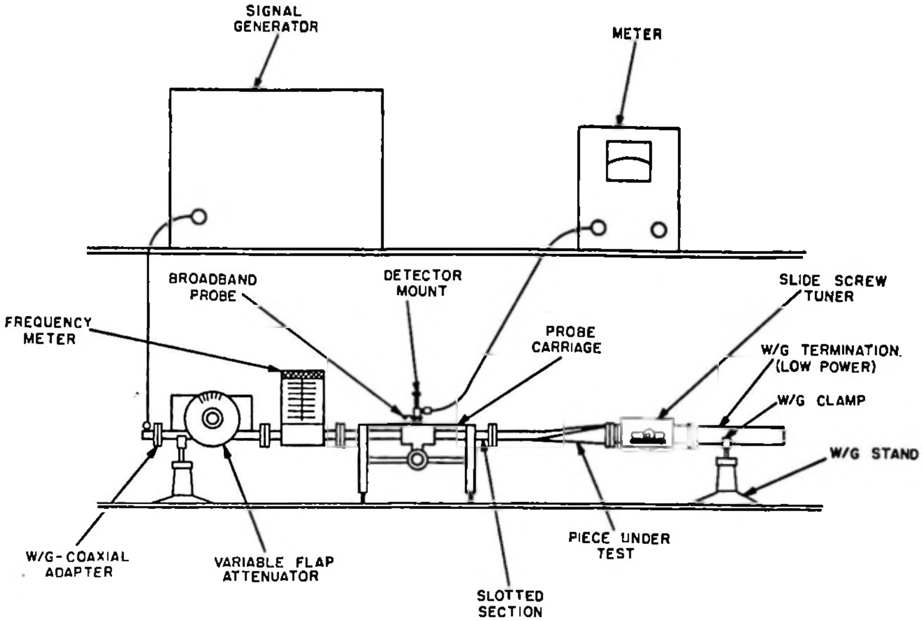


Figure 13-253. Waveguide Probe and Slotted Line Production Testing

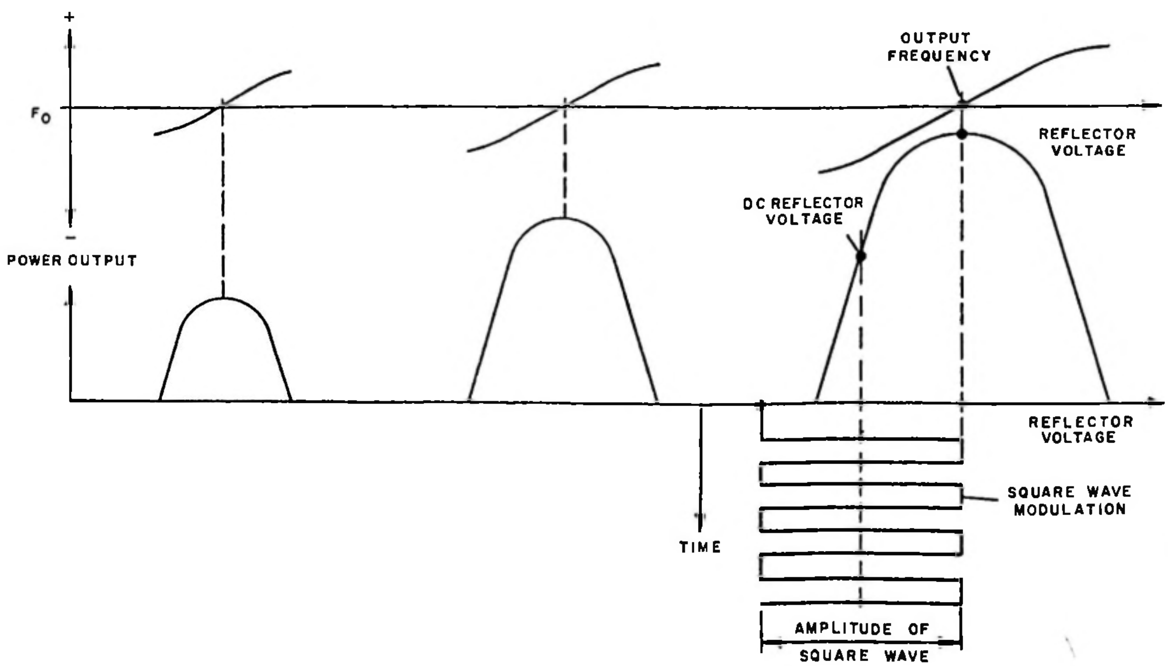


Figure 13-254. Reflector Voltage Versus Frequency and Power Output of Reflex Klystro:

of no output at all. The frequency varies through each mode from a minimum to a maximum around the medium value, corresponding approximately to the maximum power output of the tube for each mode. This explains the reason for the square-wave modulation. The voltages must be so adjusted that when a square wave is superimposed on the dc reflector voltage, the tube swings in and out of one mode, giving full output during one half-cycle and zero output during the other. At the same time, the frequency will correspond to the maximum output frequency, and all frequency modulation will be avoided. Frequency modulation is detrimental to the accuracy of the measurements because, in general, the reflection coefficient is sensitive to frequency changes.

13-892. You must tune the amplifier to the frequency of the square-wave modulation for maximum response. The optimum operating modulation conditions are then arranged by varying the reflector voltage and square-wave amplitude until the meter deflection is maximum. You will know when the tube is operating under the best modulation condition by observing whether a slight change in any control setting decreases the meter reading. This optimum operating condition corresponds also to the highest stability of the output because small changes in reflector voltage and square-wave amplitude, due to fluctuations of the line voltage or other accidental causes, will have no effect during the zero output half-cycle, and will have little effect during the full output half-cycle because of the rounded top of the mode characteristics. When you adjust the power supply, modulator, and amplifier, the probe is tuned to maximum output and the setup is ready for the measurement procedures. These are carried out by moving the detector carriage to a point where the meter reads maximum and then adjusting the amplifier gain until the meter reads exactly full scale. Without again touching the gain control, move the carriage to a point where the meter reads minimum and at this point note the meter reading. If the phase must be known, together with the amplitude of the reflection coefficient, note the positions of the carriage at maximum and minimum reading also. The meter face is calibrated to read the vswr directly. The reflection coefficient of everything that follows the slotted line is cumulatively determined by use of this method. This method, therefore, is useful for testing a single component, such as a termination which absorbs almost the total amount of the incident power. In the case of a transmission component which is not supposed to absorb any power, such as a twist, a bend, a corner, etc, a termination must be placed after it to absorb all the power conveyed by the component under test. This termination must give no reflection of its own; otherwise, the reflection of the termination will add vectorially to the reflection of the component under test, and a corresponding error will result. You must therefore tune the load to zero reflection. This is done by setting it directly against the slotted line and adjusting the position until no standing waves can be detected in the line.

13-893. The simplest method of tuning out the reflection of the load is as follows:

a. With the carriage of the slide-screw tuner in any position, start with the micrometer screw all the way out and insert it slowly, watching the meter until it just moves. This means that the tip is just protruding into the waveguide.

b. Read the vswr; then move the carriage by steps of about 1/8 inch and read the vswr each time.

c. Decrease the length of the steps until the carriage position for minimum vswr is found.

d. Adjust the tip depth until unity vswr is obtained. A minor change in carriage position may be required for very large probe penetrations. In such cases, steps c and d are repeated until the desired result is obtained.

13-894. The component to be tested is then inserted between the slotted line and the tuner and load, and the vswr reading is taken. When a large number of identical components are tested at the same frequency, the tuner and load should be handled with care, and occasionally checked against the standing wave detector. If any detuning appears after a series of tests, it should be corrected before you proceed with other tests. The test procedure with square-wave modulation and an audio-frequency amplifier can be summarized as follows:

a. Place the tuner and load against the standing wave detector.

b. Tune the amplifier to the square-wave modulation frequency.

c. Adjust, alternately, the reflector voltage and square-wave amplitude to maximize the meter reading.

d. Tune the standing wave detector probe for maximum response. The gain of the amplifier should be gradually decreased in steps 2 through 4, to avoid damage to the meter. The probe depth should be the minimum consistent with proper sensitivity.

e. Tune the termination as described.

f. Proceed with the testing.

13-895. The accuracy of the vswr that can be expected from this method is ± 0.01 for a range between 2.5 and 1.05, if you use reasonable values of stable line voltage and well regulated power supplies. The test setup recommended for high-precision testing is illustrated in part A of figure 13-255. Stability is one of the most important factors involved in the accuracy of the measurements. An effort must be made to eliminate as many of the potential sources of instability as possible. The first source of instability is the square-wave modulation, which also spreads out the output spectrum in a band of frequencies instead of providing a single frequency. The result of eliminating modulation is that an audio-frequency amplifier cannot be used for the measurements, and it becomes necessary to resort to a very sensitive galvanometer. Since no variable gain is provided at the input of the meter, the full-scale adjustment is made with the aid of a variable attenuator in the line. The amount of attenuation in the line should never be less than 20 db. This will prevent changes in the tube output, due to changes in the line impedance resulting from the standing-wave detector probe motion. Since the calibration curve of most variable attenuators is rather steep in the 20-db region, it is advisable to use two variable attenuators in series, the first one for coarse adjusting and the second, operated at small attenuation, as a vernier.

13-896. Connect the galvanometer to the standing wave detector through a resistive network designed to make the crystal work into its optimum resistance and to make the galvanometer see its critical damping resistance in the input circuit. A switch should also be provided to temporarily disconnect the crystal and occasionally check the zero setting of the galvanometer if there appears to be a tendency to drift. A movable load follows the component being tested. The operation of the movable load can be understood by observing that small reflection coefficients add vectorially (as long as second and third-order reflections can be neglected), and that a displacement along the line causes the reflected vector to rotate one complete turn for each half-wavelength of travel. The reflection coefficient, Γ, measured by the slotted line, is the sum of the reflection coefficient of the tested com-

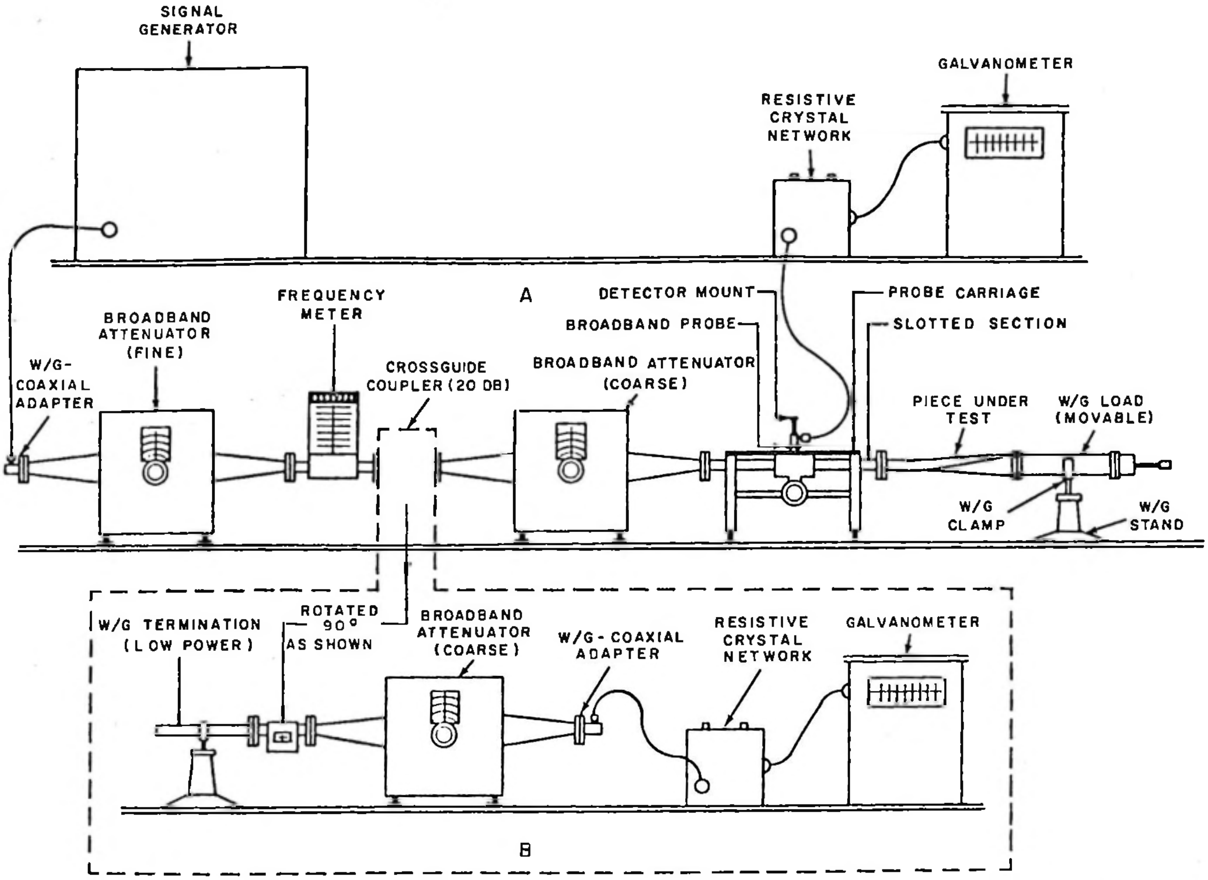


Figure 13-255. High-Precision Waveguide Testing

ponent and the reflection coefficient of the resistive strip.

13-897. You must consider the following two factors in this connection:

a. It is almost impossible to design and build a resistive strip with zero reflection coefficient over the band.

b. It is very difficult to measure the phase angle of very small reflection coefficients.

13-898. If the resistive strip has a small (but not negligible) reflection coefficient and it is moved along the guide, the total reflection coefficient measured by the slotted line will show a constant component (1) and a rotating component (2), which will make one full turn for each half-wavelength of travel of the resistive strip, as shown in part A of figure 13-256. If the strip is displaced exactly one-quarter of a wavelength in the waveguide, its reflection coefficient will rotate 180 degrees, so that in the second position (2), it will be exactly opposite the reflection coefficient of the strip in the first position (3), as shown in part B of the figure. Now if two measurements of Γ are taken with the strip in two positions $\lambda_{g/4}$ apart, as shown in part C, you can see that the reflection coefficient of the component is 1, as shown in part D, while 2 and 3 are the re-

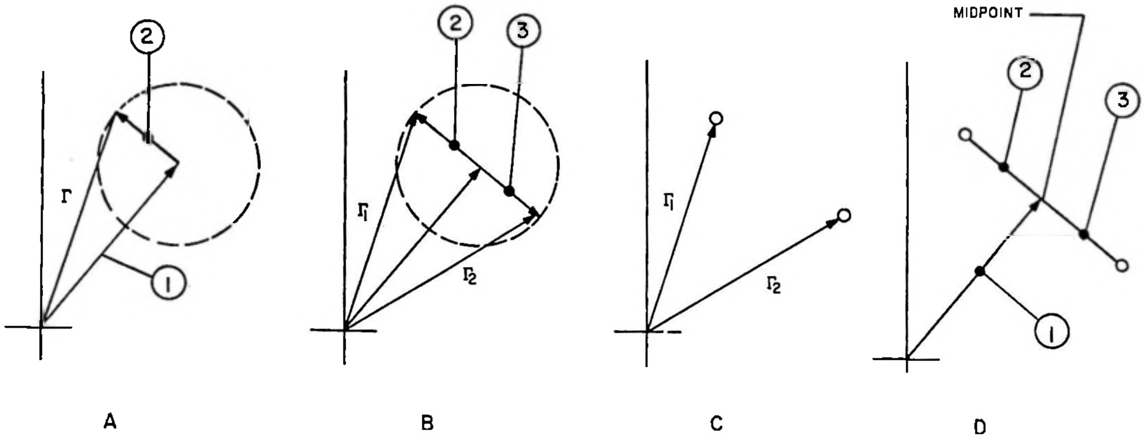


Figure 13-256. Components of Reflection Coefficient

flection coefficients of the strip in the two positions. Take each measurement of Γ at least three times, and plot the average value on polar paper, where the midpoint of the segment connecting the points representing Γ_1 and Γ_2 is taken to represent the reflection coefficient of the component in terms of amplitude and phase. At least three readings for each vector are required to decrease the probable error, resulting from accidental experimental errors, to an acceptable value. The three readings should be taken in different parts of the standing wave detector, using the following procedure:

a. Determine the positions of three successive minima.

b. Move the probe to a point equidistant from the first two minima.

c. Adjust the vernier variable attenuator for full-scale reading of the galvanometer.

d. Move the probe to the first minimum already found and noted. Do not determine the minimum again; simply move the carriage to the noted position.

e. Read the galvanometer at the first minimum.

f. Repeat all operations in connection with the other two minima.

13-899. Step b is required because a minimum is much sharper and better defined than a maximum, which is rounded and thus cannot be easily located. On the other hand, a maximum always occurs exactly halfway between two successive minima. The average of the three position readings should coincide with the central reading. The difference gives the order of experimental error involved in the determination of the phase angle. This difference should never be larger than two tenths of a millimeter in the X band, which is pertinent to our example. The average of the galvanometer readings gives the amplitude of the reflection coefficient. If the error spread of the three readings is too large, more readings should be taken. The difference between an individual reading and the average should never be larger than one-half percent of the total reading. When the first vector is determined, the strip is displaced by one quarter-wavelength in the guide, and the second vector is determined in a similar manner.

13-900. The complete determination of the reflection coefficient from the measured data requires the definition of a reference plane for the determination of the phase angle, and a calibration curve for transforming the galvanometer reading into the amplitude of the reflection coefficient. Both the reference plane and the calibration curve can be obtained by a simple calibration procedure. The calibration is made by tightly clamping a smooth metal plate against the output flange of the standing wave detector, thus presenting an almost perfect short exactly in the plane of the flange. In this case, the reflection coefficient is one, and the field distribution in the guide follows a sine law. The phase of the reflection coefficient is 180 degrees at the plane of the flange. Then the position of the first maximum is determined, and the line attenuator is adjusted until the galvanometer reads full scale. After this, the carriage is moved by steps of 1 millimeter from one end to the other of its travel, and the galvanometer readings versus carriage positions are recorded. The operation is repeated backward until the carriage comes to the original position. This procedure may be repeated again two or three times, to eliminate accidental errors. The resulting values are then compared with the readings of a table of sines, which gives the true value of the field at each point. The calibration curve for the system, comprising the crystal, the adapter, and the galvanometer, can then be constructed. Since no variable gain is inserted between the crystal and the galvanometer, the crystal always works at the same power level and always in the same region of its dynamic characteristic. The distance between two successive minima gives, directly, half-wavelengths in the slotted line, which is slightly different from half-wavelengths in the unslotted line; the position of any minimum is the required reference point for determination of the phase angle. You can obtain the phase angle by calculating the difference between the average position reading during the actual measurement and the closest reference point, and dividing this difference by a half-wavelength. The result is the fraction of 2π, or 360 degrees, corresponding to the phase angle. It should be noted that since the reference plane already has a phase angle of 180 degrees, a corresponding angle of 180 degrees should be added to or subtracted from the value thus determined to be consistent with the definition given for the phase angle. Of course, the phase angle determined by means of the described arrangement can also be used directly with a slight modification of the definition of phase angle given previously.

13-901. The calibration of the slotted line may take several hours, and it is very difficult to prevent oscillator output variations during this time. The instability of the oscillator over this long period is taken care of by the additional equipment shown in part B of figure 13-255. A small amount of power is extracted from the main line by the directional coupler and fed to a secondary line comprising a variable attenuator and a crystal mount. The output of the crystal is fed through an adapter to a second galvanometer. Both galvanometers are set to read full scale at the beginning of the calibration, and any variation in power output is monitored by the second galvanometer. The input power, as indicated by the second galvanometer, should be kept constant by varying the attenuation of the vernier attenuator in the main line. A single galvanometer may be used if the adapter is equipped with a selector switch connecting the meter to either of two separate inputs. The same arrangement illustrated in part B of figure 13-255 can be used during current measurements if, for some reason, the tube output is so unstable that it varies appreciably in the course of a single measurement. The calibration of the slotted line also indicates the amplitude of the instrumental error in connection with slight mechanical misalignments of the carriage, attenuation in the waveguide, etc. As a matter of fact, it is also possible to obtain, from a comparison of the measured values of the standing wave pattern with the theoretical sine law pattern, small corrections which are functions of the carriage position and which may be applied to the galvanometer reading in order to obtain a more accurate value. The calibration also gives an indication of the amount of probe error which can be eliminated. The calibration should be repeated for a series of frequencies throughout the band, and the results noted in the form of a graph or tables to be used with any test frequency. A graph showing the position of a reference point and the length of a half-wave in the detector versus frequency is most useful for its great ease of interpolation, and can logically become part of the calibration data for the standing wave detector. The degree of accuracy obtainable by the above described method is well below ± 0.001 of the reflection coefficient, and favorably compares with the accuracy of most instruments used in a general physics laboratory.

13-902. An alternative procedure for measuring the vswr is known as the <u>reflectometer method</u>. This method is useful in measuring the vswr over either a frequency band or at a single frequency. The method is fast, but not as accurate as the slotted line method. It has the further disadvantage of giving no location of the standing wave's position of minimum or other indication of phase angle. The reflectometer method measures, directly, by means of directional couplers, both the incident wave:

$$E_1 = E_0 \, e^{j(\omega t - \frac{2\pi}{\lambda_g} z)}$$

and the reflected wave:

$$E_2 = E_r \, e^{j(\omega t + \frac{2\pi}{\lambda_g} z)}$$

By contrast, the slotted line method compares the vectorial sum of these waves at different points along the line. Mathematically, when using the reflectometer method:

$$\text{Vswr} = \frac{E_{max}}{E_{min}}$$

where:

$$E_{max} = (E_0 + E_r) \, e^{j(\omega t + \frac{\phi}{2})}$$

$$E_{min} = (E_0 - E_r) \, e^{j(\omega t + \frac{\phi}{2} + \frac{\pi}{2})}$$

13-903. The test setup recommended is shown in figure 13-257. Adjust the signal source to a single frequency, or to cover an entire band. Connected to the attenuator is the incident wave directional coupler, which samples 20 db (10%) of the incident power. The power sampled by the incident directional coupler is detected by the first of a pair of matched detectors. The resulting 1-kc signal is presented to a ratio indicator for comparison with the reflected signal. Following the incident directional coupler is the reflected directional coupler. It samples 10 db (10%) of the reflected power, which, like the incident power, is detected and presented to the ratio indicator. The ratio indicator shows you the reflection coefficient, Γ, directly. However, in sweep frequency applications, the value of Γ generally changes as the frequency changes, resulting in a continually varying dial indication. To assist you in determining the value of Γ for any frequency of interest, you should use a long-persistence scope display. The vertical amplifier of the scope receives a dc signal from the ratio indicator which changes with the reflection coefficient, Γ, while the horizontal amplifier receives a voltage which changes with the oscillator frequency. Frequency vs Γ, or the vswr, can thus be read directly from the face of the scope. It is important that the test piece be terminated with a broad-band load whose vswr is well below the value to be measured.

13-904. PRINCIPLES OF IMPEDANCE AND ADMITTANCE MEASUREMENT.

13-905. The concept of impedance has a very definite meaning when applied to a two-

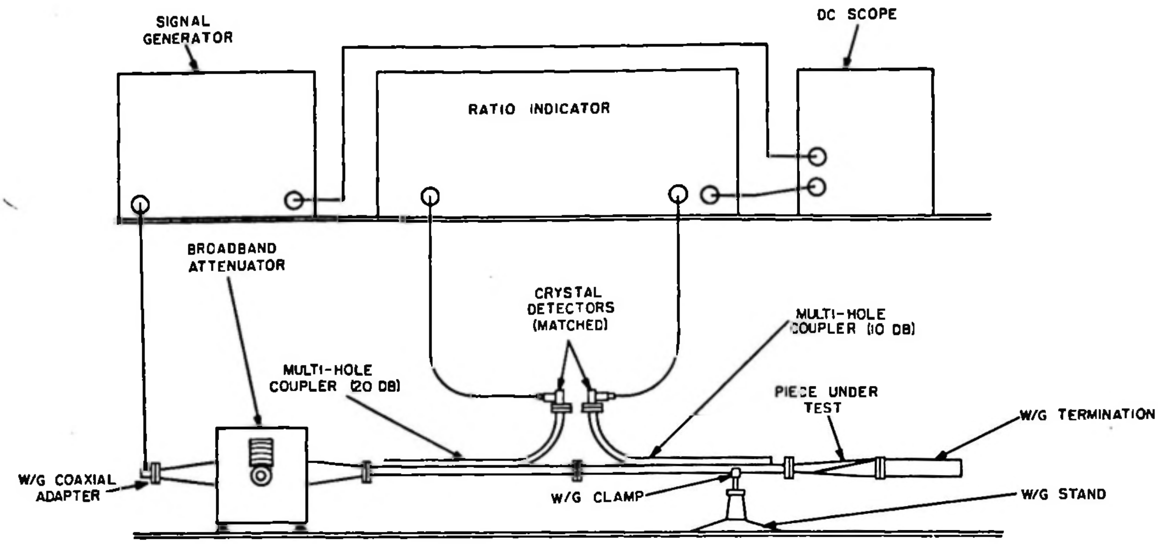


Figure 13-257. Reflectometer Arrangement for VSWR Measurement

conductor transmission line where voltages and currents can be easily measured at the input and output terminals or anywhere between the two conductors. The same definition of impedance as a ratio of voltage to current cannot be applied immediately to a waveguide because a waveguide has no distinct terminals from which voltages and currents may be measured. A more general definition of impedance, therefore, must be assumed in connection with waveguides which may be derived by considering the propagation of electromagnetic waves in a dielectric. The intrinsic impedance of a dielectric is defined as follows:

$$\eta = \sqrt{\frac{\mu}{\epsilon}}$$

where μ is the specific permeability of the dielectric and ϵ is its specific capacitivity, or dielectric constant. For air, $\mu = 1.26 \times 10^{-6}$ henries per meter, and $\epsilon = 8.85 \times 10^{-12}$ farads per meter. The intrinsic impedance of air, therefore, is 377 ohms. The theory of propagation of plane waves in the unbounded dielectric shows that η is also the ratio of the electric field intensity, E, in volts per meter to the magnetic field intensity, H, in ampere-turns per meter. E and H are both perpendicular to each other and to the direction of propagation. By analogy, we may define the wave impedance, Z_0, of the waveguide, as the ratio of transverse electric field intensity E to transverse magnetic field intensity H for a wave traveling in one direction and in a pure mode only. Z_0 depends upon the intrinsic impedance of the dielectric filling the guide, the mode of excitation, and the frequency. More specifically, the wave impedance for a TE mode is found to be:

$$Z_0(TE) = \frac{\eta}{\sqrt{1 - (f_c/f)^2}}$$

Also, the wave impedance for a TM mode is found to be:

$$Z_0(TM) = \eta\sqrt{1 - (f_c/f)^2}$$

where f_c is the cutoff frequency for the particular mode excited. The wave impedance, Z_0, has the dimensions of a resistance, being proportional to η , as long as the frequency is higher than the cutoff frequency but assumes the dimensions of a reactance (in imaginary ohms) when the frequency becomes lower than the cutoff frequency. According to this definition, a change in the dielectric properties of the medium filling the guide has the same effect on the wave propagation that the same change in the unlimited dielectric has on the propagation of plane waves through it. You can theoretically show that the reflection coefficient at a discontinuity where a dielectric with intrinsic impedance η_0 is substituted for by a medium with intrinsic impedance η is:

$$\left|\Gamma\right| = \left|\frac{S-1}{S+1}\right|$$

where S is the ratio of the wave impedance in the second section to the wave impedance in the first section. The phase angle of Γ is zero when S is larger than one, and π (or 180 degrees) when S is less than one.

13-906. More generally, you can show that the impedance, Z, defined as the ratio of transverse E to transverse H, is a complex quantity when a reflected wave is present, and that it formally obeys the following general equation, which is well known in two-conductor transmission line theory:

$$Z = Z_0 \frac{Z_1 + jZ_0 \tan\frac{2\pi}{\lambda_g} z}{Z_0 + jZ_1 \tan\frac{2\pi}{\lambda_g} z}$$

Up to this point Z_1 was the wave impedance of the second section of waveguide filled with dielectric intrinsic impedance η_1, that is, a purely resistive value. The reflection coefficient in this case was just what the conventional transmission line theory would predict in connection with a transmission line of characteristic impedance Z_0 terminated into a resistive load of resistance $R = Z_1$. A further extension of the definition of impedance in the waveguide may be readily made by observing that a purely resistive termination is a very particular case, and that, in general, waveguide components have reflection coefficients with phase angles not necessarily equal to 0 or π. Thus, the complex impedance of a termination, a component, etc, may be defined as equal to the impedance of the load that, placed in a conventional two-conductor transmission line, would present a reflection coefficient equal in phase and amplitude to the reflection coefficient actually measured in the waveguide.

13-907. As long as impedances are calculated according to this definition, they obey all formal laws well known in conventional transmission line theory. Impedance may be readily derived from reflection coefficient measurements with the aid of Smith's chart or other equivalent impedance charts. Measurements of impedances (and obviously measurements of admittances), therefore, require the same procedure as measurements of the reflection coefficient plus a transformation equivalent to the standard transformation in two-conductor transmission line theory.

13-908. PRINCIPLES OF FREQUENCY MEASUREMENT.

13-909. All types of microwave measurements require, directly or indirectly, determination of the frequency on which the measurement is made. A frequency meter (also called wave meter) is, therefore, a necessary part of every test setup. The frequency meter functions on the fundamental principle of resonance, and may be compared to the resonant LC circuit well known in lower-frequency techniques. However, the physical form that the resonator assumes is quite different from the conventional coils and capacitors known at the lower frequencies. The resonator is a section of waveguide, shorted at both ends, with metal plates; it resonates at the frequency which makes its physical length equal to any number of half-wavelengths at the particular mode which is excited. This attributes to each cavity an infinite number of resonant frequencies. The frequency, or rather the infinite set of frequencies, is varied by varying the physical dimensions of the cavity. For mechanical reasons, the resonator is usually in a cylindrical form, and the dimensions are varied by moving the bottom or top metal plate. The corresponding waveguide, therefore, is cylindrical, and may be excited in any of the modes pertaining to a cylindrical waveguide. The best suited modes are the TE_{0N} modes, in which no current flows between the cylindrical walls and the end plates. Greater ease of design and greater consistency of results are possible because no electric contact is required between moving parts. The TE_{01} mode is usually selected among all TE_{0N} modes because of the smaller dimensions of the cavity and the reduction in spurious frequency responses.

13-910. Resonance is indicated by a sharp change in the absorption or transmission properties of the cavity. The absorption type of wave meter is inserted on a branch of the line; when it resonates, it absorbs part of the energy traveling in the line into the branch. The line output, therefore, shows a marked dip when the wave meter resonates. The transmission type of wave meter is inserted in series with the line; when it resonates, it allows part of the energy to pass through and appear at the output end of the line. The output shows a marked peak when the wave meter resonates.

13-911. Frequency meters are designed to give a very sharp indication so that the measurement will be accurate. In other words, they are very frequency-sensitive around resonance, since even a very small change in frequency results in a large change in absorbed or transmitted power. For this reason, they must be well detuned after the frequency has been measured in the course of some other measurement. When the absorption type of wave meter is detuned, it does not absorb any significant amount of power; therefore, it can be inserted directly into the main line, as indicated in figures 13-253 and 13-255. The transmission type of wave meter cannot be used directly in the line because it would transmit no power when detuned, and, therefore, no measurement would be possible. On the other hand, it is not feasible to keep it resonating all the time. The solution is to insert it into an auxiliary line coupled to the main line, by means of a directional coupler, so that the power flow in the main line suffers almost no variation and a small amount of power is extracted for measuring the frequency. This small amount of power is either reflected or transmitted by the wave meter; when it is transmitted, a crystal at the output end of the wave meter indicates resonance. The procedure for a frequency measurement may be summarized as follows:

a. Slowly turn the adjusting head of the wave meter, watching the indicator (amplifier or galvanometer) until it shows a sharp dip. The dip is so sharp that it may pass unobserved if the adjusting head is turned too fast.

b. Tune very carefully until the meter reading is minimum.

c. Read the frequency indicated on the scale.

d. Detune the frequency meter by rotating the adjusting head several complete turns in either direction.

13-912. It is important to note that the absorption type of wave meter requires a certain amount of attenuation in the line between the generator and the frequency meter. If insufficient attenuation is present, the wave meter may "pull" the oscillator when it is near resonance, and cause a false peak in the indicator reading. Frequency measurements performed under such circumstances may result in considerable error.

13-913. PRINCIPLES OF RESONANT CAVITY MEASUREMENT.

13-914. Since you can use a cavity resonator as a tuned circuit in high-frequency circuits, it follows that you can use a resonator as a wavemeter. Although a cavity-resonator wavemeter uses various wave modes of oscillation, its operating theory is much the same as that of a coaxial line wavemeter shorted at both ends. A cylindrical cavity is perhaps the most practical type for use as a wavemeter because it is easier to machine accurately.

13-915. There are two main classes of wave modes that can be usefully employed in a cavity-resonator wavemeter: the $TM_{0,1,0}$ mode and the $TE_{0,1,1}$ mode. A typical wideband wavemeter which operates in the $TM_{0,1,0}$ mode is illustrated in part A of figure 13-258. A set of four wavemeters of this kind can be made to cover a frequency range from 200 to 10,000 mc, with an accuracy better than one part in ten thousand. The design of the wavemeter illustrated in part A of the figure is based on the very large change in characteristic impedance which occurs between the two sections of coaxial line. The use of the $TE_{0,1,1}$ mode as applied to wavemeters has several important advantages over the $TM_{0,1,0}$ mode; these advantages are summarized as follows:

a. The amount of energy dissipated on the cavity walls is extremely low; therefore, a high Q value results.

b. There is no radial current present on the end cavity walls.

c. The TE_0 form of wave composes only modes which have no angular variation of field strength, and thus does not introduce spurious points of resonance.

13-916. A high Q value is an essential feature of the resonance type of wavemeter, and a cavity excited in the $TE_{0,1,1}$ mode is superior in this respect to any other. With the absence of radial current flow in the TE_0 wave, there is no need to establish good electrical contact between the end walls and the cylindrical surface. In fact, the absence of such a contact will tend to suppress any other modes of oscillation in the cavity which might otherwise arise. If one of the end plates is made slightly smaller in diameter than the cylinder, and arranged to move axially over a limited range, the resonant frequency of the cavity may be varied and the cavity used as a wavemeter. Wavemeters excited in the $TE_{0,1,1}$ mode are illustrated in parts B and C of figure 13-258. Coupling to the wavemeter, shown in part B of the figure, is provided by means of small loops similar to those in the coaxial type of wavemeter. An alternative coupling method, which has the advantage of avoiding spurious wave-mode excitations, is illustrated in part C of the figure. Cavity-resonator wavemeters are usually silver-plated on the inside to improve their Q value; for accurate work the enclosure should be hermetically sealed or equipped with a dehydrator such as silica gel.

13-917. The calibration of a cavity wavemeter is theoretically calculable from the dimensions of the cavity. However, such dimensions are usually not known with a sufficient degree of accuracy because the effective cavity size changes with temperature. A copper cavity may change in resonant wavelength about three parts in ten thousand between summer and winter, and such a change is by no means negligible at frequencies above 3000 mc. A cavity-resonator wavemeter, therefore, should have the calibration temperature recorded

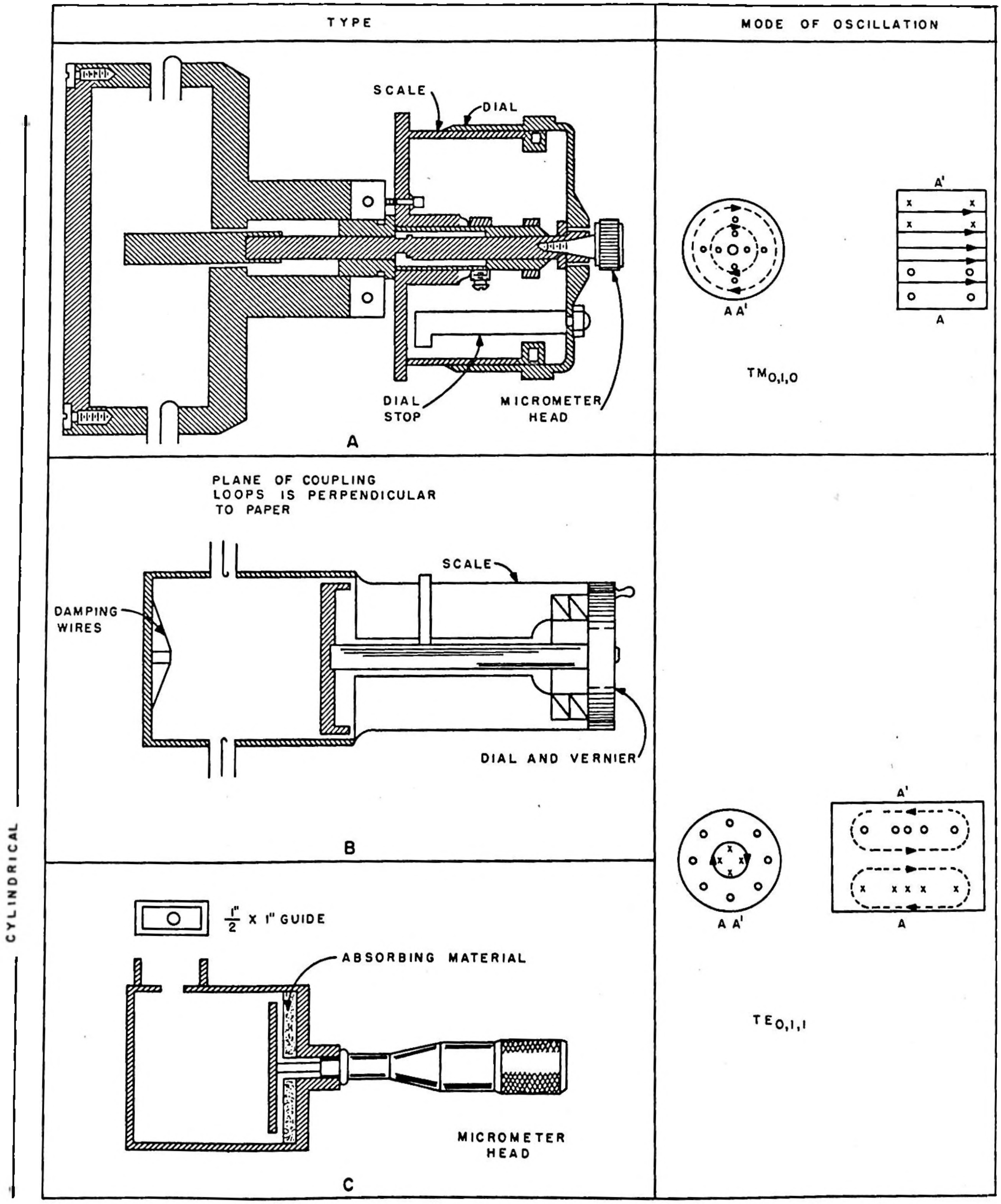


Figure 13-258. Cylindrical Wavemeters

on the instrument; if the working temperature differs greatly from that value during operation, you must make appropriate allowances.

13-918. You can calibrate cavity-resonator wavemeters by either of the two methods listed below:

a. By use of a heterodyne frequency meter.

b. By use of a coaxial-line wavemeter.

13-919. The first method is perhaps the more accurate system of calibration. Power from an rf oscillator is fed into a matched waveguide, which is equipped with short probes to couple the cavity and the heterodyne meter to the guide. You must vary and measure the oscillator frequency with the heterodyne frequency meter while taking dial readings of the cavity wavemeter. You should then prepare a calibration chart to include the date, temperature, and humidity at the time of calibration. In the heterodyne method of calibration, you must decrease the unknown frequency to a predetermined value by heterodyne action, and then amplify and measure the signal at this frequency. The unknown frequency is determined from the heterodyne frequency that must be used in order to change the unknown frequency to the known fixed frequency.

13-920. Cavity calibration may be made as described above by substituting a coaxial-line wavemeter or another cavity of known calibration in the place of the heterodyne frequency meter. Although the accuracy obtained by use of these alternative standards is considerably lower than that obtained with the heterodyne meter, the calibration procedure can be carried out much faster. The accuracy of a wavemeter depends upon the accuracy with which its scale is divided, and also upon the loaded Q value. For example, if an accuracy of ± 0.001 cm is desired, the scale must be fine enough to read at least three decimal places, and the sensitivity of the wavemeter must be great enough to determine when the tuning is 0.001 cm or less from resonance. In other words, the frequency response curve of the wavemeter must drop to its "half power" points within 0.001 cm of the resonant wavelength. The necessary loaded Q value of a wavemeter may be found by the following formula:

$$Q_L = \frac{\lambda}{\Delta \lambda}$$

where:

Q_L = loaded Q value

λ = operating wavelength in centimeters

$\Delta \lambda$ = desired accuracy in centimeters

Therefore, at 10 cm, for an accuracy of ± 0.001 cm, the loaded Q value must be

$$10/2\ (0.001) = 5000$$

13-921. PRINCIPLES OF ATTENUATION MEASUREMENT.

13-922. Attenuation is the ratio of the power available at the output end of a dissipative component to the power fed to the input end. Attenuation is currently measured in decibels (db), defined as the logarithm to the base 10 of this ratio, multiplied by 10; that is:

$$\text{Att}_{(db)} = 10 \log_{10} \frac{P_2}{P_1} = -10 \log_{10} \frac{P_1}{P_2}$$

13-923. If and when the impedance of the output waveguide is equal to the impedance of the input line, the ratio of the power is equal to the square of the ratio of the electric (and magnetic) fields. In this case you can write the attenuation in db as follows:

$$Att_{(db)} = 20 \log_{10}\frac{E2}{E1} = 20 \log_{10}\frac{H2}{H1} =$$

$$-20 \log_{10}\frac{E1}{E2} = -20 \log_{10}\frac{H_1}{H_2}$$

13-924. Under the same condition, the attenuation can also be measured in nepers, defined as the logarithm to the base e of the ratio of electric (or magnetic) fields; that is:

$$Att_{(n)} = \ln\frac{E2}{E1} = \ln\frac{H2}{H1}$$

In the case of a lossy line, the propagation equation becomes:

$$E = E_0 \, e^{j\omega t - (j\beta + \alpha)z}$$

You can see that α is the negative of the attenuation in nepers per unit length (meter).

13-925. You can measure attenuation by using a substitution method, that is, by measuring what known amount of attenuation in the line results in the same power reduction as the insertion of the tested component in the line. A variable attenuator can be used for this purpose; it is inserted between the component and a crystal mount, as shown in part A of figure 13-259. The attenuator is adjusted until the meter contains an arbitrary midscale reading, which is recorded. Then the component is taken away and the attenuator is set directly against the input waveguide. The attenuation is increased until the meter repeats the same reading again. The difference between the attenuation corresponding to the two attenuator settings gives the attenuation of the tested component. This method can be varied in several ways according to the type of component to be tested and the test equipment available. For example, instead of reading the difference in attenuation with one variable attenuator, you can use two attenuators, adjusting the power with one and leaving the other set to zero. When the component is removed, read the attenuation directly on the second attenuator. This variation may be preferable because the calibration curve of a variable attenuator is generally less steep and more accurate near the beginning than at higher attenuation values. You can apply another variation, using the variable gain of the amplifier to set the meter at the reference reading instead of using a separate attenuator. This method is less preferable because the crystal may be exposed to a high power level in the course of the measurement; consequently, its characteristic may deviate from the square law which it should closely approximate. Another important variation will be described in connection with coupling measurements. Table 13-7 gives the correspondence between decibels, nepers, power ratios, and voltage ratios, assuming that the input and output impedances are equal. It should be noted that the number measuring the attenuation in db is negative, and that a positive number measures the reciprocal of the attenuation, that is, gain (the ratio of output power to input power). However, the minus sign in front of the db value is often omitted when no confusion is possible, as is done in table 13-7.

13-926. PRINCIPLES OF COUPLING MEASUREMENT.

13-927. Coupling is simply a special case falling under the more general category of attenuation measurements, being the ratio of the power in one part of a particular structure to the power in another part. More particularly, coupling is the ratio of the power output of an auxiliary guide to the power input of a main guide. This definition is used in measuring the coupling coefficient of directional couplers.

13-928. A directional coupler is a device which extracts a small amount of power traveling down the main waveguide and couples it into an auxiliary waveguide. The main

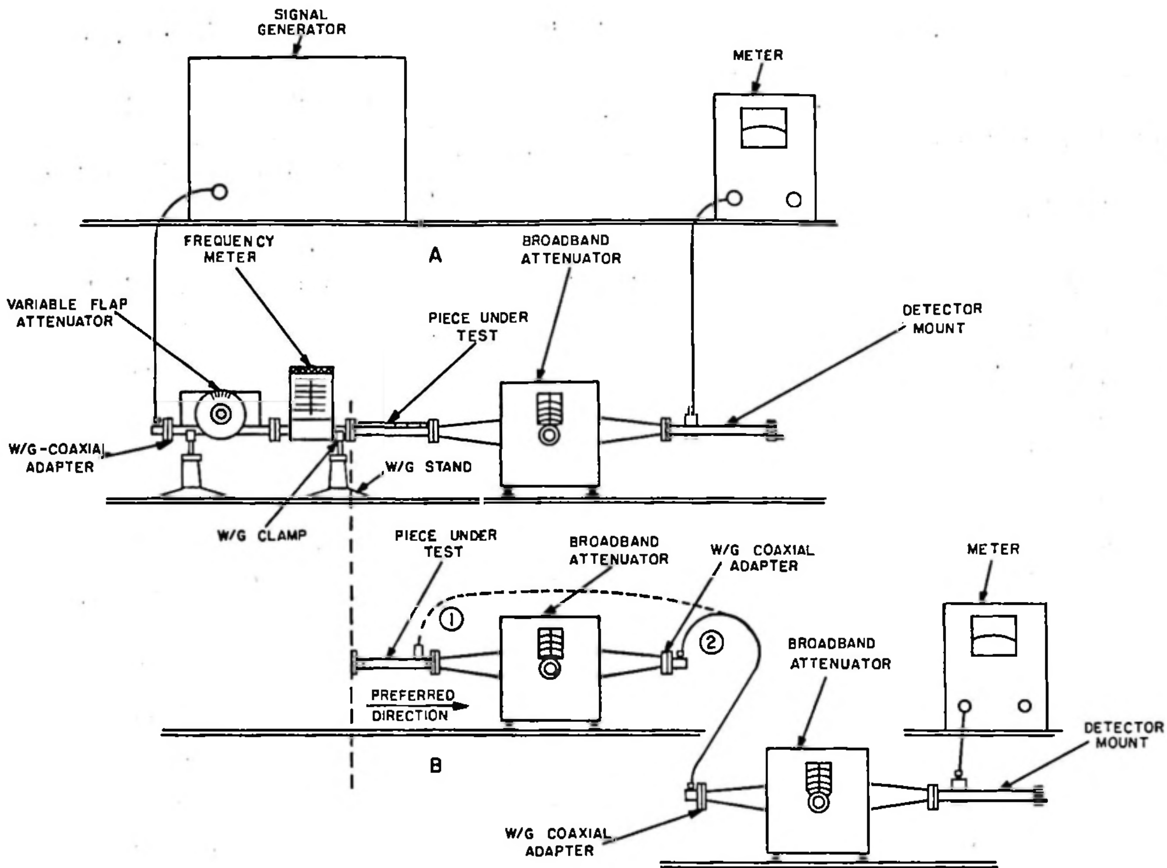


Figure 13-259. Attenuation Measurement Using the Substitution Method

feature of the directional coupler is the fact that a percentage of the power traveling in one direction appears only at the output of the auxiliary guide, while almost no part of the power traveling in the opposite direction is coupled out. The measurement of the coupling is performed as a regular attenuation measurement, except that the crystal mount is first connected to the auxiliary guide output and the meter is set to the reference reading. The directional coupler is then removed and the attenuation measured as usual. The main guide output is connected to a matched load during the first part of the operation. A variation of this method must be employed when the auxiliary guide output is taken through a coaxial connection. In this case, a coaxial cable feeds the output to a transformer, as shown in part B of figure 13-259, which is connected to the variable attenuator and the crystal mount (connection 1 in the figure). After setting the reference reading on the meter by means of this variable attenuator, the directional coupler can be removed and the connection made to another transformer, following the measuring variable attenuator (connection 2 in the figure), which is connected in place of the removed directional coupler. During the first part of the operation, this variable attenuator may be used as a matched load, as shown in the figure, or a separate load can

Table 13-7. Correspondence Between Decibels, Nepers, Power, and Voltage Ratios

DECIBEL	NEPERS	POWER RATIO (DB POSITIVE)	POWER RATIO (DB NEGATIVE)	VOLTAGE RATIO (DB POSITIVE)	VOLTAGE RATIO (DB NEGATIVE)
0.1	0.011513	1.0233	0.97723	1.0116	0.98852
0.2	0.023026	1.0471	0.95502	1.0233	0.97723
0.3	0.034539	1.0715	0.93327	1.0315	0.96946
0.4	0.046052	1.0965	0.91199	1.0471	0.95502
0.5	0.057565	1.1220	0.89127	1.0593	0.94402
0.6	0.069077	1.1482	0.87093	1.0715	0.93327
0.7	0.080590	1.1749	0.85114	1.0839	0.92259
0.8	0.092103	1.2023	0.83174	1.0965	0.91199
0.9	0.10362	1.2303	0.81281	1.1092	0.90155
1.0	0.11513	1.2589	0.79434	1.1220	0.89127
1.2	0.13815	1.3183	0.75855	1.1482	0.87093
1.4	0.16118	1.3804	0.72443	1.1749	0.85114
1.6	0.18421	1.4454	0.69183	1.2023	0.83174
1.8	0.20723	1.5136	0.66068	1.2303	0.81281
2.0	0.23026	1.5849	0.63095	1.2589	0.79434
2.2	0.25328	1.6595	0.60259	1.2882	0.77628
2.4	0.27631	1.7328	0.57710	1.3183	0.75855
2.6	0.29934	1.8198	0.54951	1.3490	0.74129
2.8	0.32236	1.9055	0.52480	1.3804	0.72443
3.0	0.34539	1.9953	0.50118	1.4125	0.70796
3.5	0.40295	2.2387	0.44669	1.4962	0.66836
4.0	0.46052	2.5119	0.39811	1.5849	0.63095
4.5	0.51808	2.8184	0.35481	1.6788	0.59566
5.0	0.57565	3.1623	0.31623	1.7783	0.56233
5.5	0.63321	3.5480	0.28185	1.8836	0.53086
6.0	0.69077	3.9811	0.25119	1.9953	0.50118
7.0	0.80590	5.0119	0.19953	2.2387	0.44669
8.0	0.92103	6.3096	0.15849	2.5119	0.39811
9.0	1.0362	7.9433	0.12589	2.8184	0.35481
10.0	1.1513	10.0000	0.10000	3.1623	0.31623
11.0	1.2664	12.589	0.079434	3.5480	0.28185

Table 13-7. Correspondence Between Decibels, Nepers, Power, and Voltage Ratios (Cont)

DECIBELS	NEPERS	POWER RATIO (DB POSITIVE)	POWER RATIO (DB NEGATIVE)	VOLTAGE RATIO (DB POSITIVE)	VOLTAGE RATIO (DB NEGATIVE)
12.0	1.3815	15.849	0.063095	3.9811	0.25119
13.0	1.4967	19.953	0.050118	4.4668	0.22387
14.0	1.6118	25.119	0.039811	5.0119	0.19953
15.0	1.7269	31.623	0.031623	5.6234	0.17783
16.0	1.8421	39.811	0.025119	6.3096	0.15849
17.0	1.9572	50.119	0.019953	7.0795	0.14125
18.0	2.0723	63.096	0.015849	7.9433	0.12589
19.0	2.1875	79.433	0.012589	8.9125	0.11220
20.0	2.3026	100.00	0.010000	10.0000	0.10000
22.0	2.5328	158.49	0.0063095	12.589	0.079434
24.0	2.7631	251.19	0.0039811	15.849	0.063095
26.0	2.9934	398.11	0.0025119	19.953	0.050118
28.0	3.2236	630.96	0.0015849	25.119	0.039811
30.0	3.4539	1000.0	0.0010000	31.623	0.031623
32.0	3.6841	1584.9	0.00063095	39.811	0.025119
34.0	3.9144	2511.9	0.00039811	50.119	0.019953
36.0	4.1446	3981.1	0.00025119	63.096	0.015849
38.0	4.3749	6309.6	0.00015849	79.433	0.012589
40.0	4.6052	10^4	0.00010000	100.000	0.010000
42.0	4.8354	$10^4 \times 1.5849$	$10^{-5} \times 6.3095$	125.89	0.0079434
44.0	5.0657	$10^4 \times 2.5119$	$10^{-5} \times 3.9811$	158.49	0.0063095
46.0	5.2959	$10^4 \times 3.9811$	$10^{-5} \times 2.5119$	199.53	0.0050118
48.0	5.5262	$10^4 \times 6.3096$	$10^{-5} \times 1.5849$	251.19	0.0039811
50.0	5.7565	10^5	10^{-5}	316.23	0.0031623
52.0	5.9867	$10^5 \times 1.5849$	$10^{-6} \times 6.3095$	398.11	0.0025119
54.0	6.2170	$10^5 \times 2.5119$	$10^{-6} \times 3.9811$	501.19	0.0019953
56.0	6.4472	$10^5 \times 3.9811$	$10^{-6} \times 2.5119$	630.96	0.0015849
58.0	6.6775	$10^5 \times 6.3096$	$10^{-6} \times 1.5849$	794.33	0.0012589
60.0	6.9077	10^6	10^{-6}	1,000.00	0.0010000
70.0	8.0590	10^7	10^{-7}	3,162.3	0.00031623
80.0	9.2103	10^8	10^{-8}	10,000.0	0.00010000
90.0	10.362	10^9	10^{-9}	31,623.0	0.000031623
100.0	11.513	10^{10}	10^{-10}	100,000.0	0.000010000

be used to terminate the directional coupler output. In either case, the reflections from the termination should be small.

13-929. When the coupling is small (less than approximately —20 db), very little power is coupled into the auxiliary guide, and the difference between the input power and the output power is negligibly small. If the vswr of the main guide is small, it is proper to assume that the output power of the main guide is equal to the input power. In this case, the measurement can be made without removing the directional coupler by measuring the power ratio between the output of the auxiliary guide and the output of the main guide. This variation saves time in production testing, where, in general, the tolerances in specifications are wider than the probable experimental error involved in this method. The probable experimental error can be estimated from the reflection coefficient of the main line (power lost by reflection), the coupling (power lost by coupling in the auxiliary line), and the calibration tolerances of the variable attenuator. However, this time-saving procedure is not recommended for precision testing.

13-930. PRINCIPLES OF DIRECTIVITY MEASUREMENT.

13-931. The directivity is the ratio of power coupled out when the main guide is fed in the preferred direction, to the power coupled out when the main guide is fed with the same amount of power in the opposite direction. The first method that presents itself for measuring the directivity is to measure the coupling in both directions and determine the power ratio (or the difference in decibels). It is very important that the main line termination be perfectly matched when the coupling is measured in the opposite direction, because any reflection from the termination enters the directional coupler in the preferred direction and experiences a much tighter coupling than the direct power which enters the directional coupler in the opposite direction. Since the output of the auxiliary line depends on the power coupled from both directions and is maximum when the two waves add in phase (it is minimum when the two waves add with opposite phase), the resulting power output can be anywhere between the two limits, depending upon the phase of the reflection coefficient of the termination. Now if C is the coupling and Γ is the amplitude of the reflection coefficient of the termination, the power output due to the wave traveling in the preferred direction (i.e., from the termination) is $C\,\Gamma$ for unit power input. The power coupled from the wave traveling in the opposite direction (i.e., from the direct wave) is, by definition, $\frac{C}{D}$, where D is the directivity. The total power output, therefore, will be contained between the limits:

$$C\left(\Gamma - \frac{1}{D}\right) \leq W \leq C\left(\Gamma + \frac{1}{D}\right)$$

13-932. The power output will be zero when, and only when, the reflection coefficient is equal to the reciprocal of the directivity and the two waves have opposite phase. This suggests another method for measuring the directivity, by means of a termination whose reflection coefficient can be varied at will in amplitude and phase. Figure 13-260 illustrates the test setup.

13-933. Insert the directional coupler into the line with the preferred direction corresponding to the direction of propagation of the wave reflected from the termination. The reflection coefficient of the termination is varied in phase and amplitude by means of the slide-screw tuner until the output from the auxiliary line is reduced to zero (connection 1). The directional coupler is then removed and the reflection coefficient of the termination measured with the standing wave detector (connection 2). The directivity, D, can be transformed in decibels as follows:

$$D_{(db)} = -20 \log_{10} \Gamma$$

13-934. PRINCIPLES OF MICROWAVE POWER MEASUREMENT.

13-935. GENERAL. High-frequency rf power measurements are generally classified into two groups: high-level measurements and low-level measurements. Several devices are not available for measuring high-level rf power, some of which are outlined below.

a. Continuous-flow calorimeter.

b. Johnson power measurer.

c. Low-level power measurer equipped with an attenuator.

13-936. CONTINUOUS-FLOW CALORIMETER. The continuous-flow calorimeter is probably the simplest device to use and the easiest to build. Figure 13-261 illustrates the construction of such a device for use with a coaxial transmission line. A section of coaxial transmission line is short-circuited at the far end and closed at the other by means of a thick dielectric bead, which is usually made of glass or micalex. The length of the device should be such that the distance from the bead to the short circuit is several wavelengths; this is in order to insure complete absorption of the rf energy. The length of the dielectric bead should be a quarter-wavelength of the operating frequency, and the dielectric constant of the bead is chosen in accordance with the following equation:

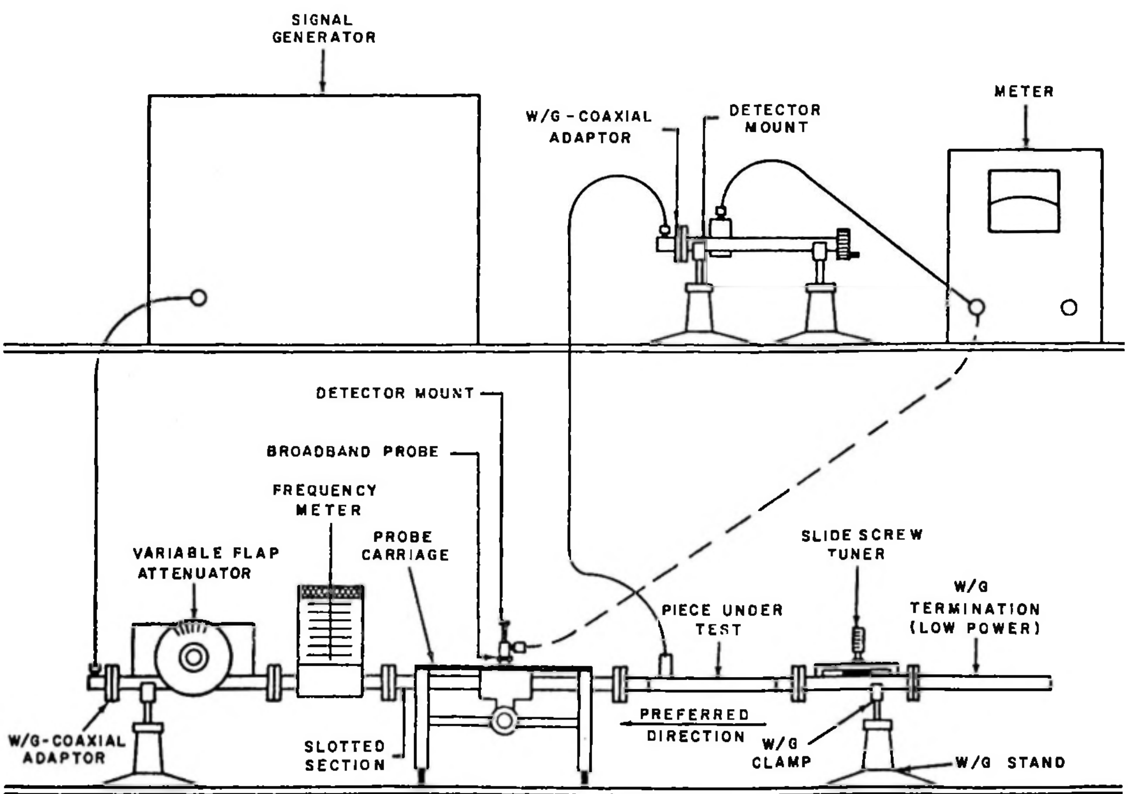


Figure 13-260. Directivity Measurement Test Setup

$$Z_d = \sqrt{Z_{air} \times Z_{H_2O}}$$

where:

Z_d = impedance of dielectric-filled line

Z_{air} = impedance of air-filled line

Z_{H_2O} = impedance of water-filled line

13-937. Water flows continuously through the chamber, and is heated by the rf energy. Thermocouple probes are placed in the water inlet and outlet to measure the rise in temperature of the water flowing through the cavity. To find the rf power, it is necessary to measure the rate of flow and the change in temperature of the water. The rf power may then be found by the following formula:

$$W = 4.19\ VT \qquad (5)$$

where:

W = rf power in watts

V = volume of water flowing through cavity per unit time

T = temperature change of the water

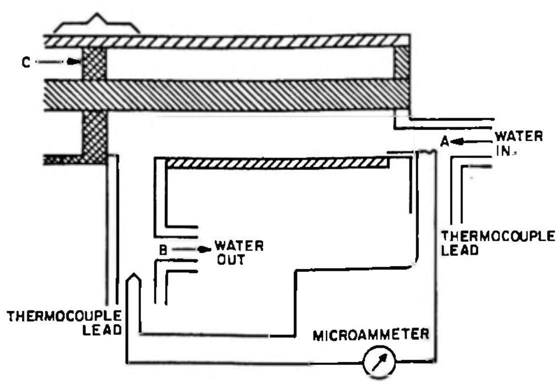


Figure 13-261. Coaxial Water Load

In practice, T is not measured directly; it reflects the current in microamperes flowing through the thermocouple. The difference in temperature of the two thermocouple junctions is a linear function of the current. Therefore, it is necessary to know the calibration of the thermocouples and the meter. The constant may be obtained by placing the thermocouple junctions in water baths which differ in temperature by T_0 and noting the deflection, d, of the meter used. Formula 5 may then be rewritten as follows:

$$W = 4.19 \frac{T_0}{d}\ VD \qquad (6)$$

where:

$\frac{T_0}{d}$ = calibration constant for components used

d = deflection of meter during power measurement

Since the expression $4.19\frac{T_0}{d}$ in formula 6 is a constant for a given set of components, the rf power may then be found by the formula:

$$W = AVD \qquad (7)$$

where:

$A = 4.19\ T_0/d$

Thus, once the constant A is known, the rf power absorbed in the termination can be determined by multiplying the constant by the rate of water flow, V, and the meter deflection, d. There are two advantages in measuring rf power by this method, as follows:

a. The measurement is absolute.

b. The rise in temperature can be made small, thereby limiting the energy lost by radiation and conduction from the transmission line. However, in order to maintain a constant rate of water flow, it is necessary to use some device to supply water at a constant pressure. The only serious disadvantage of the method is that such a device must completely terminate a line, and, in so doing, absorb 100 percent of the rf power. A continuous-flow calorimeter, therefore, cannot be used as a power monitor except by the use of a mechanical switching arrangement.

13-938. A waveguide calorimeter is electrically the same as a coaxial-line calorimeter, and is matched to the guide by a quarter-wavelength transformer. The transformer consists of a section of guide whose electrical length is one quarter-wavelength of the operating frequency, and is filled with a material whose dielectric constant is the geometric mean of the dielectric constants of air and water.

13-939. JOHNSON WATTMETER. A type of rf power measuring device which may be used as a constant monitor is the Johnson wattmeter. This type of meter requires only 1 percent of the rf energy in the transmission line as compared to the 100 percent needed for the constant-flow calorimeter. Figure 13-262 illustrates a type of Johnson wattmeter which is designed for use with a waveguide transmission line. A small longitudinal section of the waveguide wall is removed and replaced with a piece of high-resistance material. A sheet of constantan 0.015 inch thick can be used. At high rf power levels, enough heat is dissipated in the constantan to produce an appreciable temperature change. Since the conductivity of the waveguide is greater than that of the constantan, there will be a measurable difference in the temperature of the two materials. This difference in temperature is used to give an indication of the rf power level. The difference in temperature is measured by a platinum resistance thermometer which consists of 200 ohms of 0.002-inch platinum wire wound on the constantan section and another 200 ohms wound on the guide. The two windings serve as two arms of a Wheatstone bridge, as shown in figure 13-262.

13-940. If the bridge is initially balanced, any change in temperature of one of the coils with respect to the other will unbalance the bridge. The unbalanced condition is caused by the change in resistance of the platinum wire around the section of constantan with temperature. The unbalanced current in the bridge is directly proportional to the power in the line, and a straight-line calibration curve can be drawn by measuring only one point with an accuracy of 5 percent. Since there is a time delay between turn-on of the power and the thermal equilibrium of the measuring device, there is a time interval of between one and three minutes after turn-on before readings can be taken.

13-941. LOW-LEVEL POWER MEASURER USING AN ATTENUATOR. Low-level power measurements cannot generally be carried out by the methods described above because the temperature rise becomes too small for accurate determination. However, three common detectors may be used to advantage for such measurements: crystals, bolometers, and thermistors. The crystal detects by virtue of its rectifying property, while the bolometer and thermistor detect through a resistance change induced by the heating effect of absorbed rf power. The thermistor consists of a resistance element made from certain metal oxides which have a high negative temperature coefficient; that is, the resistance of the thermistor decreases with increasing temperature. Its sensitivity is approximately ten times that of a bolometer, and it has both mechanical ruggedness and good overload characteristics. Therefore, for accurate work, the thermistor is the most desirable of the three detectors mentioned.

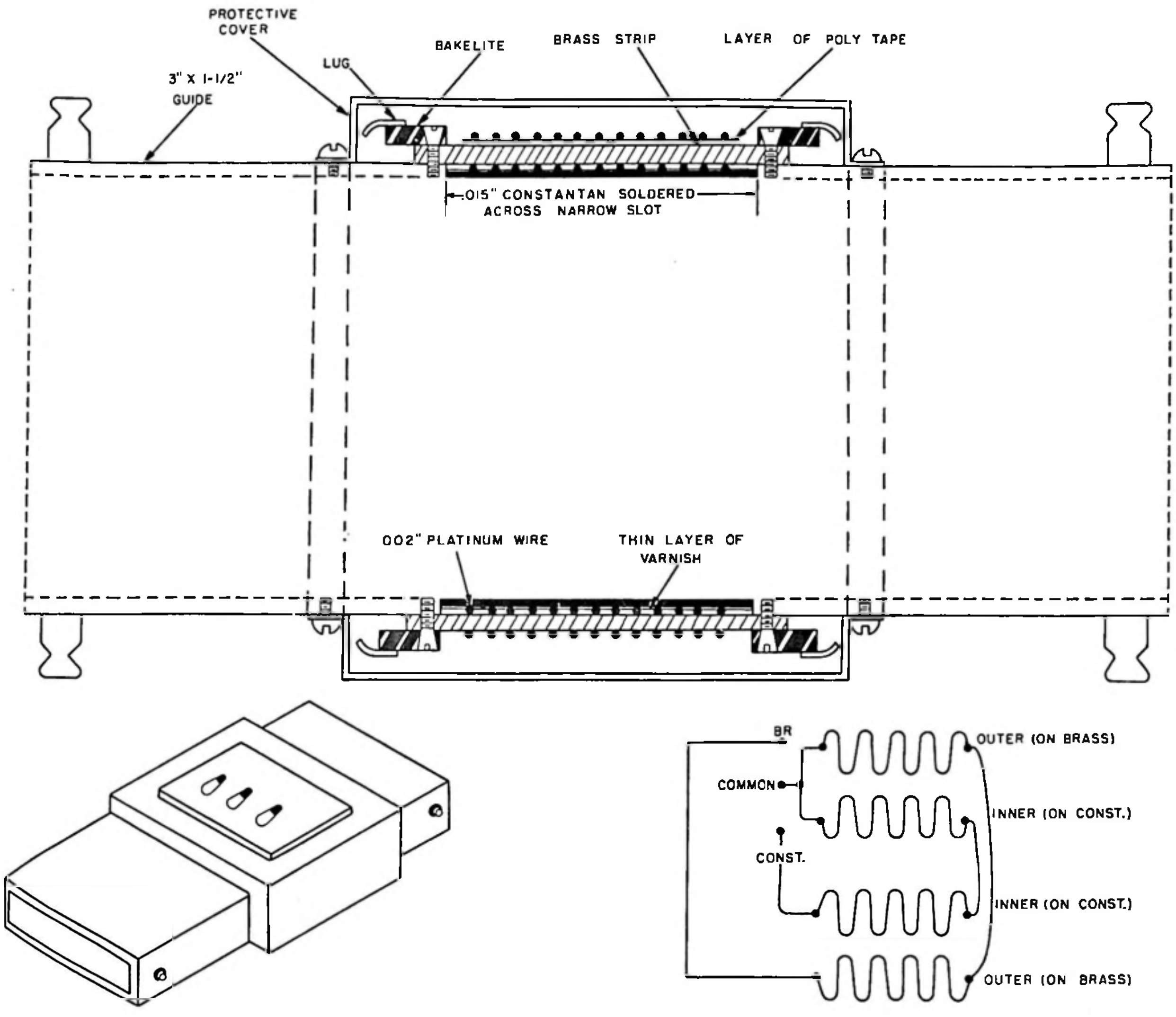


Figure 13-262. Johnson Wattmeter

13-942. Precise measurements of small rf power levels are generally made by use of a balanced bridge. The bridge is first balanced with a known amount of direct current, after which an unknown amount of rf power is applied to the detector, unbalancing the bridge again. Balance is restored by removing some of the direct current from the detector, and the change in dc power is used to calculate the magnitude of the rf power added. Most bridge circuits are so arranged that the difference in dc power between the two balanced conditions is exactly equal to the magnitude of the rf power.

13-943. An unbalanced bridge may also be used to measure low-level rf power; in fact, it is easier and faster to use than the balanced bridge. Such a bridge is initially balanced on direct current, and the deflection of the bridge meter is noted when the rf power is added. The meter scale is previously calibrated in rf power, to give a direct-reading rf microwattmeter. For instance, with a thermistor detector and a low-resistance microammeter, it is possible to make full-scale deflection represent 1 milliwatt with a calibration that is appreciably linear.

SECTION XII

LOADS AND BALANCING DEVICES

13-944. GENERAL.

13-945. The load for a transmission line may be any device capable of dissipating rf power. When lines are used for transmitting purposes, the most common type of load is an antenna. When a transmission line is connected between an antenna and a receiver, the receiver input circuit, not the antenna, is the load since the power taken from a passing wave is delivered to the receiver.

13-946. Whatever the application, only the conditions existing at the load determine the standing-wave ratio on the line. If the load is purely resistive and equal in value to the characteristic impedance of the line (Z_0), there will be no standing waves. If the load is not purely resistive and/or is not equal to the line Z_0, there will be standing waves. Adjustments made at the input end of the line cannot change the swr, nor is the swr affected by changing the line length. Only in a few special cases is the load inherently of the proper value to match a practical transmission line. In all other cases, it is necessary either to operate with a mismatch and accept the swr that results, or to take steps to bring about a proper match between the line and the load by means of transformers or similar devices. Impedance-matching transformers may take a variety of physical forms, depending on the circumstances.

13-947. Note that if the swr is to be made as low as possible, the load at the point of connection to a transmission line must be purely resistive. In general, this requires that the load be tuned to resonance. If the load itself is not resonant at the operating frequency, the tuning can sometimes be accomplished by the matching method.

13-948. BALUNS.

13-949. GENERAL.

13-950. The devices most commonly used for coupling between an antenna and its feed, and between transmitter output circuits and their respective lines, are called baluns. A balun performs a two-fold function: it provides impedance-matching capabilities while it also maintains a balanced condition from the transmitter output to the coaxial line, and from the coaxial line to the antenna. The baluns used between the antenna and transmission line are generally of the linear type, consisting of transmission-line sections. Those generally used between the transmitter and the transmission line are of the inductively coupled matching network type, described later in this section. In cases where a fixed impedance ratio can be tolerated, a device called a coil balun may be used.

13-951. LINEAR BALUNS.

13-952. When an inherently balanced antenna is fed at the center through a coaxial line, as indicated in part A of figure 13-263, the balanced condition is upset. On the side connected to the shield, a current can flow down over the outside of the coaxial line. The re-

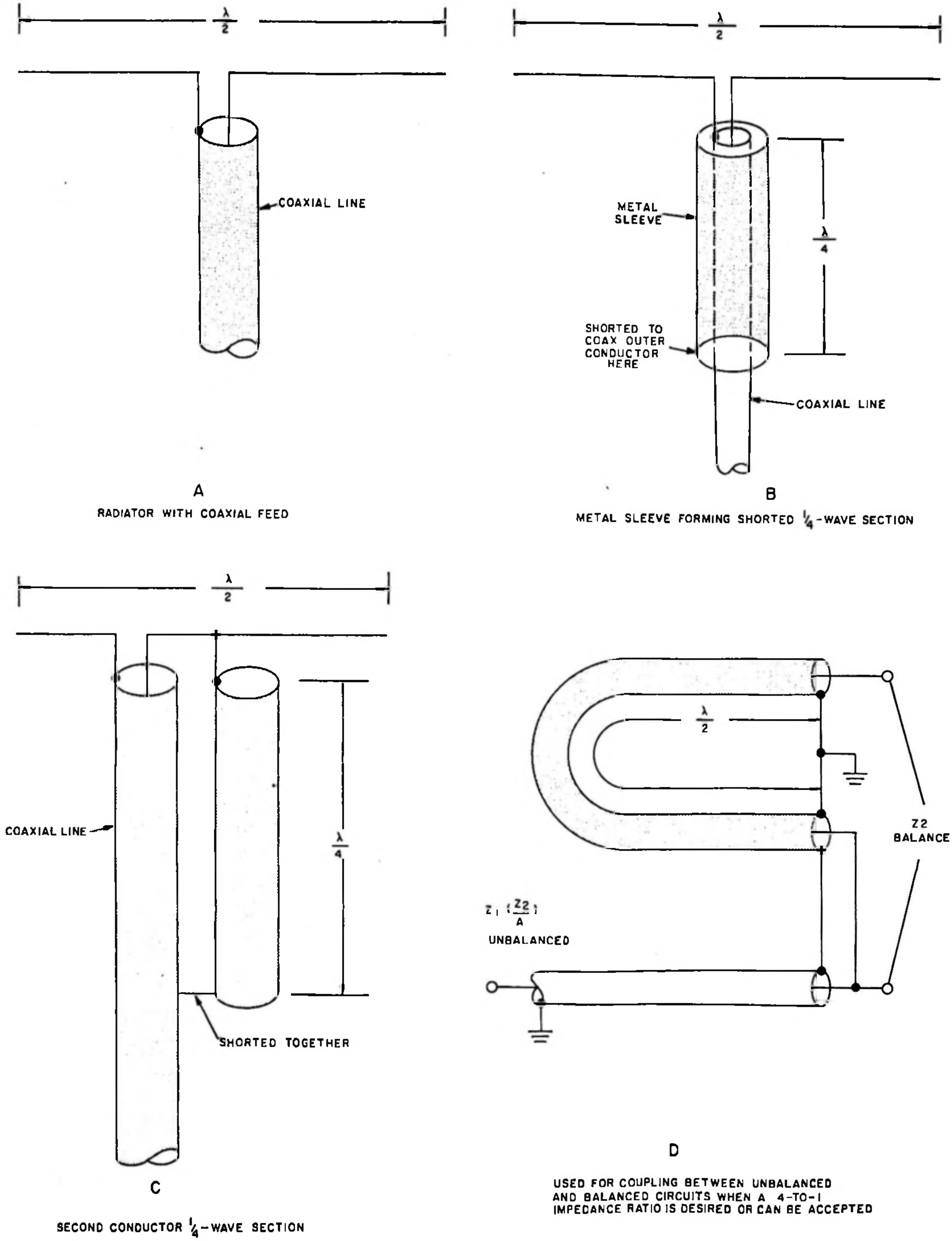


Figure 13-263. Coaxial-Fed Radiator and Typical Balun Configurations

sultant fields produced cannot be canceled by the fields from the inner conductor, because the fields inside the line cannot escape through the shield provided by the outer conductor. Hence, the current flowing on the outside of the line produces undesired radiation.

13-953. Part B shows a balun arrangement which uses a sleeve over the transmission line to form a shorted quarter-wave line section with the shield, or outer conductor, of the coaxial line. The impedance looking into such a section is very high, so that the end of the outer conductor of the coaxial line is effectively insulated from the part of the line below the sleeve. The length is an electrical quarter wave, and may be physically shorter if the insulation between the sleeve and the coaxial line is other than air. This arrangement has no effect on the impedance relationships between the antenna and the coaxial line.

13-954. Another method that provides an equivalent effect is shown in part C. Since the voltages at the antenna terminals are equal and opposite (with reference to ground), equal and opposite currents flow on the surfaces of the line and secondary conductor. Below the shorting point of the two conductors, these currents combine to cancel each other. The balancing section appears as an open circuit to the antenna, since it is a quarter-wave parallel-conductor line shorted at the far end, and thus has no effect on the normal operation of the antenna. However, the length is not essential to the line-balancing function of the device, and baluns of this type are sometimes made shorter than a quarter wavelength in order to provide the shunt-inductive reactance required on certain types of matching equipments.

13-955. Part D shows a third type of balun, in which equal and opposite voltages, balanced to ground, are taken from the inner conductors of the main transmission line and the half-wave phasing section. Since the voltages at the unbalanced end are in parallel, there is a 4-to-1 step-down ratio of impedance from the balanced side to the unbalanced side. This arrangement is useful, for example, for coupling between a balanced 300-ohm line and a 75-ohm coaxial line.

13-956. COIL BALUNS.

13-957. The principal advantage of using coil baluns is that they permit operation over a wide band, covering a frequency range of about 10 to 1 (3 to 30 mc, for example) without the need for any adjustment. This is possible because coil baluns use no resonant circuits, which limit the frequency range of other commonly used coupling devices. However, since this type of balun has a fixed impedance ratio (4 to 1), its use is restricted to equipment which can tolerate this ratio over the frequency range used and still maintain the swr within the desired limits.

13-958. Coil baluns are based on the principles of a linear transmission line, as shown in part A of figure 13-264. Two transmission lines of equal length having a characteristic impedance, Z_0, are connected in series at one end and in parallel at the other. At the series-connected end, the lines are balanced to ground and will match an impedance equal to $2Z_0$. At the parallel-connected end, the lines will match an impedance of $Z_0/2$. One end may be grounded at the parallel-connected end if the length of the two lines (considering each line as a single wire) is such that the balanced end is effectively decoupled from the parallel-connected end. This requires a length that is an odd multiple of a quarter wavelength. The impedance transformation from the series-connected end to the parallel-connected end is 4 to 1.

13-959. A definite line length is necessary only for decoupling purposes. As long as there is adequate decoupling, the arrange-

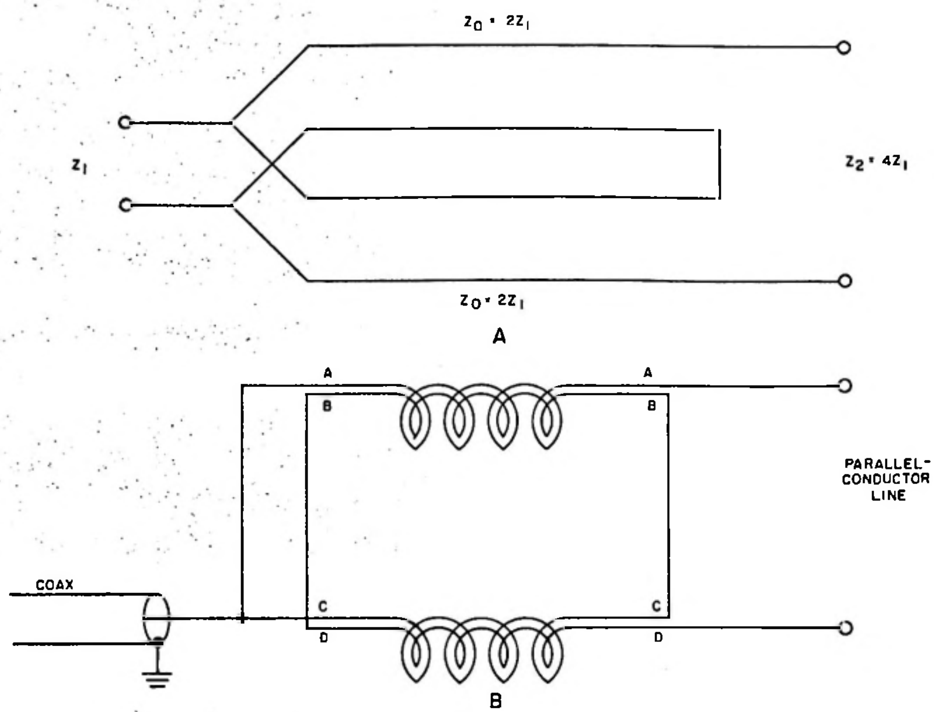


Figure 13-264. Coil Baluns

ment will act as a 4-to-1 impedance transformer regardless of the line length. If each line is wound into a coil, as shown in part B of figure 13-264, the resultant inductances will act as choke coils and will tend to isolate the series-connected end from any ground connection that may be placed on the parallel-connected end. Balun coils made in this manner will operate over a wide frequency range since the choke inductance is not critical. The lowest usable frequency is the lowest frequency at which the coils are effective in isolating one end from the other. The length of line in each coil should be about equal to a quarter wavelength at the lowest frequency to be used.

13-960. The principal use of such coils is in coupling a 300-ohm balanced line to a 75-ohm coaxial line. This requires that the characteristic impedance of the lines forming the coils be 150 ohms. A balun of the type, when matched, is simply a fixed-ratio transformer. It cannot compensate for inaccurate matching elsewhere in the equipment. For example, with a 300-ohm line on the balanced end, a 75-ohm coaxial cable will not be matched unless the 300-ohm line is actually terminated in a 300-ohm load.

13-961. NON-RADIATING LOADS.

13-962. GENERAL.

13-963. Typical examples of non-radiating loads for a transmission line are the grid circuit of a power amplifier, the input circuit of a receiver, and another transmission line. The last case includes the antenna coupler, which is a misnomer because it is actually a device for coupling a transmission line to a transmitter or receiver.

13-964. COUPLING TO A POWER AMPLIFIER GRID CIRCUIT.

13-965. GENERAL. When the grid of a power amplifier is to be fed from a low-impedance transmission line, as in a radio transmitter, an impedance step-down is necessary, since the grid input resistance is a matter of a few thousand ohms. This can be done by the use of a tank circuit as an impedance-transforming device in the grid circuit of the amplifier, as shown in figure 13-265. This coupling arrangement may be considered either as simply a means of obtaining mutual inductance between the two tank coils or as a low-impedance transmission line. If the line is longer than a small fraction of a wavelength, and if an swr bridge is available, the line is more easily handled by adjusting it as a matched transmission line.

13-966. INDUCTIVE LINK COUPLING WITH FLAT LINE. In adjusting this type of line, the object is to make the swr on the line as low as possible over as wide a band of frequencies as possible so that power can be transferred over this range without retuning. As far as the amplifier grid circuit is concerned, the controlling factors are the Q of the tuned grid circuit, L_2C_2, the inductance of the coupling coil, L_4, and the degree of coupling between L_2 and L_4. An swr indicator is essential.

13-967. Assuming that the coupling is adjustable, start with a trial position of L_4 with respect to L_2, and adjust C_2 for the lowest swr. Then change the coupling slightly and repeat. Continue in this manner until the swr is as low as possible; if the circuit constants are in the right region, it should not be difficult to obtain an swr of 1

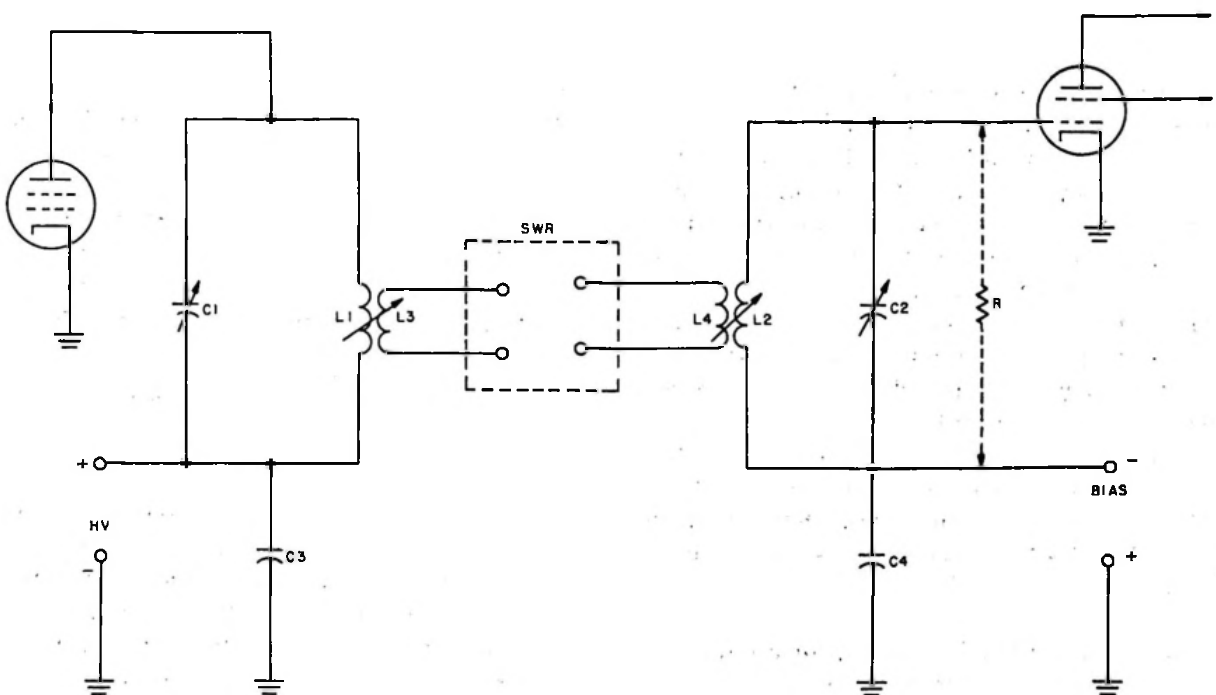


Figure 13-265. Coupling Excitation to the Grid of an RF Amplifier by Means of a Low-Impedance Coaxial Cable

to 1. The Q of the grid circuit should be at least 10. Maximum coupling, for a given degree of physical coupling, will occur when the inductance of L4 is such that its reactance at the operating frequency is equal to the characteristic impedance of the link line.

13-968. Once the swr has been reduced to 1 to 1, the frequency should be shifted over the band so that the variation of the swr can be observed without changing C_2 or the coupling between L_2 and L_4. If the swr rises rapidly on either side of the original frequency, the circuit can be made "flatter" by reducing the Q of the tuned circuit. This may be done by decreasing C_2 and correspondingly increasing L_2 to maintain resonance. The tuned circuit must then be readjusted for minimum swr. It is possible to set up the equipment so that the swr will not exceed 1.5 to 1 over the entire band. If the coupling between L_2 and L_4 is not adjustable, the same results can be obtained by varying the L/C ratio of the tuned circuit, that is, varying its Q.

13-969. When a resistance-type swr bridge is used, it is not possible to feed full power through the line when making adjustments. In such a case, the operating conditions in the amplifier grid circuit can be simulated by using a carbon resistor (1/2 or 1 watt size) of the same value as the calculated grid impedance, connected as indicated by the arrows in the figure. In this case the amplifier must be operated cold, without filament or heater power. The adjustment is the same as described above, but the driver power is reduced to a value suitable for operating the swr bridge.

13-970. When the grid coupling equipment has been adjusted so that the swr is close to 1 to 1 over the desired frequency range, the power fed into the link line will be delivered to the grid circuit with minimum loss.

13-971. LINK FEED WITH UNMATCHED LINE. When the equipment is to be treated without regard to transmission line effects, the link line must not offer any appreciable reactance at the operating frequency. Any appreciable reactance will, in effect, reduce the coupling, making it impossible to transfer sufficient power from the driver to the amplifier grid circuit. Coaxial cables especially have considerable capacitance even for short lengths and it may be desirable to use a spaced line, such as twin-lead, if the radiation can be tolerated. The reactance of the line can be nullified only if the link is made resonant. This may require a change in the number of turns in the link coils, a change in the length of the line, or the insertion of a tunable capacitance. Since the swr on the link line may be quite high, the line losses increase because of the greater current, the voltage increase may be sufficient to cause a breakdown in the cable insulation, and the added tuned circuit makes adjustment more critical with relatively small changes in frequency.

13-972. These troubles may not be encountered if the link line is kept very short for the highest frequency. Adjusting the coupling in such an arrangement must necessarily be largely a matter of trial and error. If the line is short enough to have a neglibible reactance, the coupling between the two tank circuits can be increased, within limits, by adding turns to the link coils or by coupling the link coils more tightly to the tank coils, if possible. If it is impossible to change either of these, a variable capacitor may be connected in series with, or in parallel with, the link coil at the driver side of the line, depending upon which connection is most effective. If coaxial line is used, the capacitor should be connected in series with the inner conductor. If the line is long enough to have considerable reactance, the variable capacitor is used to resonate the entire link circuit. The size of the link coils and the length of the line, as well as the size of the capacitor, will affect the resonant frequency, and it may take an adjustment of all three before the capacitor will show a pronounced effect on the coupling. When the coupling equipment has been made resonant, the coupling may be adjusted by varying the link capacitor.

13-973. COUPLING TO A RECEIVER.

13-974. A good match between an antenna and its transmission line does not guarantee a low standing-wave ratio on the line when the antenna equipment is used for receiving. The swr is determined by what the line sees at the receiver input terminals. For a minimum swr the receiver input circuit must be matched to the line. The most desirable condition exists when the receiver is matched to the line Z_0 and the line in turn is matched to the antenna. This will result in maximum power transfer from the antenna to the receiver, with the least loss in the transmission line.

13-975. Figure 13-266 shows the circuit diagram of an antenna coupler commonly used for receiving. Actually, this circuit provides impedance matching between the receiver input and the antenna feed line, and not the antenna itself. The values of capacitance and inductance vary depending upon the operating frequency of the equipment and the impedance of the line. Normally, the coupler is adjusted for optimum coupling or maximum image rejection, but if the coupler is detuned, it can be used as an auxiliary gain control to reduce the overloading effects of strong signals. This circuit is basically the same as the coupling circuits used to couple a transmitter to the line, and a more detailed description of its function and adjustment can be found in subsequent paragraphs in this section.

13-976. COUPLING A TRANSMITTER TO THE LINE.

13-977. GENERAL.

13-978. The type of coupling needed to transfer power adequately from the final rf amplifier to the transmission line depends almost entirely on the input impedance of the line. This input impedance is determined by the standing-wave ratio and the line length. The simplest case is that where the line is terminated in its characteristic im-

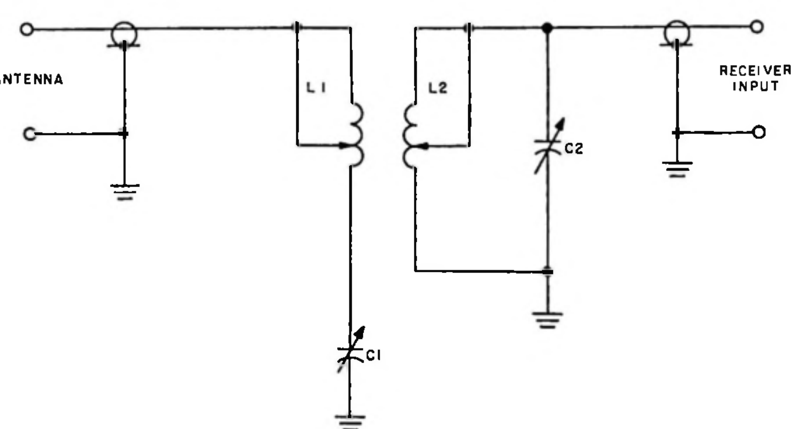


Figure 13-266. Antenna Coupler Circuit Diagram

pedance so that the swr is 1 to 1 and the input impedance is equal to Z_0 regardless of line length. Coupling methods that will deliver power into a flat line are readily designed. For all practical purposes, the line can be considered flat if the swr is no greater than about 1.5 to 1. A coupling method designed to work into a pure resistance equal to the line Z_0 will have enough leeway to handle the small variations in input impedance that will occur when the line length is changed.

13-979. Current practice in transmitter design is to provide an output circuit that will work into such a line, usually a coaxial line of 50 to 75 ohms characteristic impedance. If the input impedance of the transmission line that is to be connected to the transmitter differs appreciably from the value of impedance into which the transmitter output circuit is designed to operate, an impedance-matching network must be inserted between the transmitter output circuit and the line input terminals.

13-980. IMPEDANCE-MATCHING CIRCUITS FOR PARALLEL-CONDUCTOR LINES.

13-981. The input impedance of a line that is operating with a high standing-wave ratio can vary over quite wide limits. The simplest type of circuit that will match a range of impedances of 50 to 75 ohms is a parallel-tuned circuit resonant at the operating frequency. Such a circuit is shown in figure 13-267. Ordinarily such a circuit will be connected to a short length of coaxial line or link by inductive coupling, as shown. The other end of the cable is attached to the output terminals of the transmitter. The cable may be of any convenient length if the impedance that it sees at the matching circuit is equal to its own characteristic impedance.

13-982. The constants of tuned circuit C_1L_1 are not particularly critical. The principal requirement is that the circuit be capable of being tuned to the operating frequency. The construction of L_1 must be such that it can be tapped at least every turn. L_2 must be tightly coupled to L_1, and the inductance of L_2 should be approximately the value that gives a reactance equal to the Z_0 of the connecting line at the frequency in use. An average reactance of about 60 ohms will suffice for either 50- or 75-ohm coaxial cable. Initial set up of the coupling arrangement can be accomplished with the use of a simple resistance-type swr bridge, requiring only that the transmitter output power be reduced to a low enough value to prevent damage to the bridge.

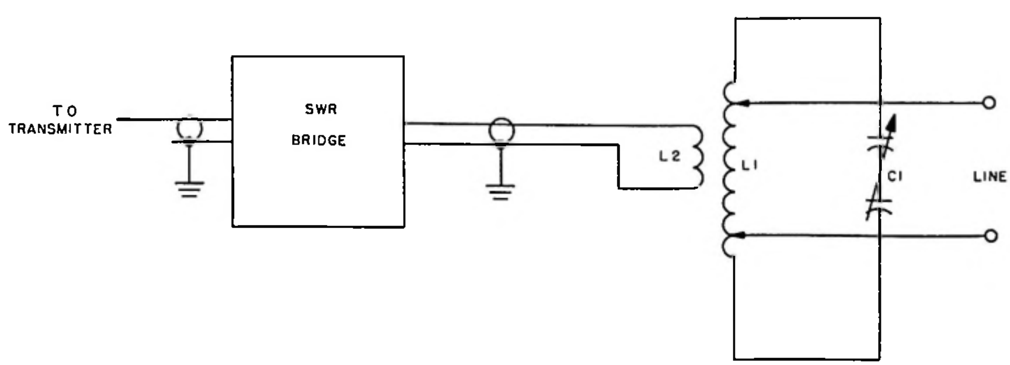


Figure 13-267. Matching Circuit Used with Parallel-Conductor Transmission Lines

13-983. To adjust the circuit, take a trial position of the line taps on L_1, selecting taps that are equidistant from the center of the coil, and adjust C_1 for minimum swr as indicated by the bridge. If the swr is not close to 1 to 1, try new tap positions and adjust C_1 again, continuing this procedure until the swr is practically 1 to 1. The setting of C_1 and the tap positions may then be logged for future reference. At this point, check the link swr over the frequency range normally used, without changing the setting of C_1. If the swr remains below 1.5 to 1, no readjustment is required. If the swr exceeds 1.5 to 1 at any part of the band, it is advisable to note as many settings of C_1 as necessary to maintain the swr below 1.5 to 1. Changes in the link swr are caused chiefly by changes in the swr of the main transmission line; relatively little change is due to the coupling circuit itself. If the antenna used is broad band enough, a single setting of C_1 at mid-frequency will suffice.

13-984. Figure 13-268 shows how the matching adjustment may be facilitated, by using a variable capacitor (C_2) in series with the matching circuit coupling coil. This additional adjustment makes the tap setting on L_1 much less critical, since varying C_2 has the effect of varying the coupling between the two circuits. For optimum control of coupling, L_2 should be considerably larger than it is when C_2 is not used; it should have about twice the reactance value discussed in the preceding paragraphs. The reactance of C_2, when set for maximum capacitance, should be the same as that of L_2 at the operating frequency. L_1 and C_1 are the same value whether or not you use C_2. The method of adjustment is the same, except that for each trial tap position, C_1 and C_2 are alternately adjusted, a little at a time, until the swr is brought to its lowest possible value. In general, an adjustment should be sought which keeps C_2 at the largest possible value, since this broadens the frequency response of the circuit.

13-985. UNTUNED COUPLING. A simple coil can be used for coupling to a transmission line having a high standing-wave ratio, provided that the line length is adjusted so that there is a current loop near the point where it connects to the pick-up coil. For a given separation between the pick-up coil and the amplifier tank coil, the coupling will be maximum if the line is cut to a length such that the input impedance is just sufficiently capacitive to cancel the inductive reactance of the pick-up coil. The higher the swr on the line, the easier it is to load the amplifier with loose coupling between the two coils. The sharper the antenna tuning and the higher the swr on the line, the more difficult it is to operate over a band without progressively changing the line length.

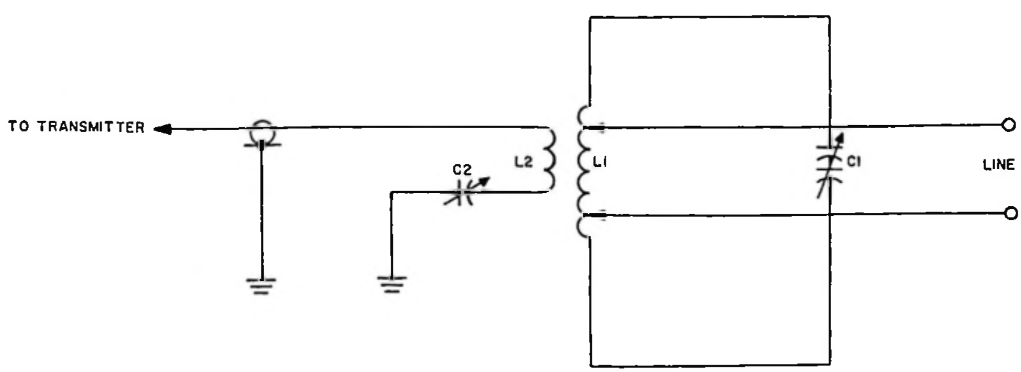


Figure 13-268. Using a Series Capacitor To Control Coupling

13-986. SERIES AND PARALLEL TUNING. Lines classified as tuned or resonant have a length equal to a multiple of one-quarter wavelength. They are characterized by having either a very high or a very low input impedance depending on whether the multiple is odd or even. Also, the impedances of such lines are essentially resistive. Under these conditions the circuits shown in figure 13-269 will work satisfactorily. Their advantage over the circuits shown in figures 13-267 and 13-268 is that taps are not required on the matching circuit coil, L_1. Series tuning is used when a current loop occurs at or near the input end of the line; i.e., when the input impedance is low. Parallel tuning is used when a voltage loop occurs at or near the input end of the line; i.e., when the input impedance is high.

13-987. In series tuning, the circuit formed by L_1, C_1, and C_2 with the line terminals short-circuited should tune to the operating frequency. C_1 and C_2 should be maintained at equal capacitance. In parallel tuning, the circuit formed by L_1 and C_1 should tune to resonance with the line disconnected. The L/C ratio on either circuit depends on the transmission line Z_0 and the standing-wave ratio. With series tuning, a high L/C ratio must be used if the swr is relatively low and the transmission line Z_0 is also low. With either series or parallel tuning, the L/C ratio is less critical when the swr is high. The coupling coil, L_2, should have a reactance equal to the Z_0 of the coaxial line. The coupling between L_1 and L_2 should be continuously adjustable. A balanced capacitor is used in the parallel arrangement in preference to a single unit. An alternative method of maintaining balance is to use two single-ended capacitors in parallel, but with the frame of one connected to one side of the line and the frame of the other connected to the other side of the line. The same two capacitors may then be switched in series when series tuning is desired. As an alternative to adjustable the coupling between L_1 and L_2, fixed coupling with a variable capacitor connected in series with L_2 may be used, as shown in figure 13-268.

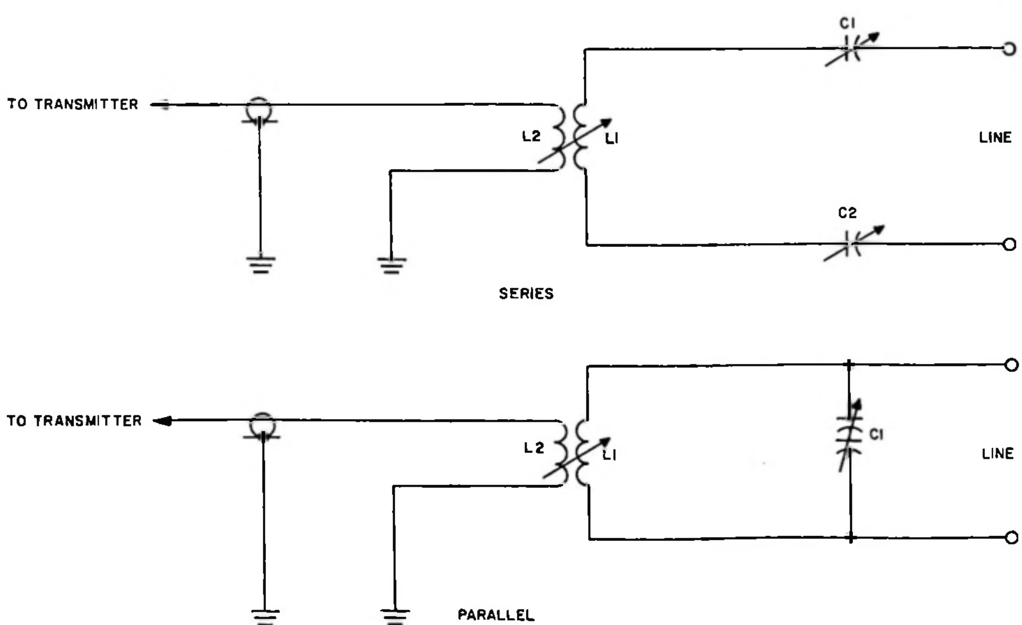


Figure 13-269. Link-Coupled Series and Parallel Tuning

13-988. These circuits should be set up and adjusted in the same way as the tapped matching circuit shown in figure 13-267. That is, an swr bridge is used to indicate the impedance match, which is obtained by alternately adjusting C_1 and the coupling between L_1 and L_2 until the bridge shows a null.

13-989. ADJUSTMENT WITHOUT SWR BRIDGE. Use of the swr bridge with the circuits previously described is an effective way of arriving at optimum adjustments. However, if a bridge is not available, the transmitter can usually be made to take the proper load by a trial method of adjustment. In the case of figure 13-267, take a trial position of the taps fairly close to the center of L_1. With loose coupling between L_1 and L_2 (either by adjustment of the mutual inductance or the series capacitor, C_2) and with the amplifier plate tank circuit tuned to resonance as indicated by the plate-current dip, vary C_1 until the plate current rises to a peak. This peak will be less than the normal loaded plate current. Then increase the coupling between L_1 and L_2, readjust C_1 for maximum plate current, and readjust the amplifier tank circuit for a current dip. Continue until the amplifier is fully loaded at the plate-current dip, increasing the coupling between the transmitter tank and the coaxial line, if necessary, to obtain full loading. Then spread the taps on L_1 a little farther apart and repeat the procedure. The object is to use the widest spread between taps that will permit proper loading of the transmitter. The procedure for series and parallel tuning is similar, except that there are no taps to adjust.

13-990. Although this trial method will generally lead to adequate transmitter loading, the adjustments are seldom optimum from the standpoint of a low swr in the coaxial link. A high swr may lead to excessive power dissipation in the link, with overheating as a result. Also, the loading may change more rapidly with small frequency changes than would be the case with a matching circuit adjusted for optimum performance with the aid of an swr bridge.

13-991. LINES OF RANDOM LENGTH. Series or parallel tuning will always perform satisfactorily with lines having a high standing-wave ratio as long as either a current or voltage loop appears at the input end of the transmission line. This condition will exist when the antenna is resonant and the line length is a multiple of a quarter wavelength. However, it is not always possible to couple satisfactorily when lines of intermediate length are used. This is because at some lengths the input impedance of the line has a relatively large reactive component, and the resistive component is too large to be connected in series with a tuned circuit and too small to be connected in parallel.

13-992. The coupling arrangement shown in figure 13-267 is capable of handling the resistive component of the input impedance regardless of the swr on the transmission line. Consequently, it can generally be used wherever series or parallel tuning would normally be called for, simply by setting the taps properly on the coil. Within limits, the same circuit is capable of compensating for the reactive component of the input impedance; this means that a 1-to-1 swr would be obtained at a different setting of C_1 than it would be if the line were purely resistive. However, C_1 may not have enough range to give satisfactory compensation, particularly when the input impedance is principally reactive. When such conditions exist, it is necessary to provide additional methods for canceling the reactive component of the input impedance. If the line impedance appears inductive, a suitable capacitance in parallel will resonate the circuit. If the line impedance appears capacitive, a suitable inductance in parallel will resonate the circuit. The resistive impedance that remains can easily be matched to the coaxial link by means of the circuit shown in figure 13-267.

13-993. The practical application of this principle is shown in parts A and B of figure 13-270, where L and C are the reactances required to cancel the line reactance, L being used for a line having capacitive reactance, and C for a line having inductive reactance. The amount of inductance or capacitance required can be determined by using a resistive-type swr bridge in the coaxial link. First, disconnect the main transmission line from L_1, and connect a noninductive 1-watt resistor with about the same resistance as the Z_0 of the line. Adjust the coil taps and C_1 for a 1-to-1 standing-wave ratio, as described earlier. This determines the proper setting of C_1 for a purely resistive load. Then remove the resistor and connect the line, again adjusting the taps and C_1 to obtain a minimum swr. Compare the new setting of C_1 with the original setting. If the capacitance has increased, the line reactance is inductive, and a capacitor must be connected as shown in part B of the figure. If the capacitance of C_1 has decreased from the original setting, the line reactance is capacitive, and an inductor must be connected as shown in part A of the figure. No specific values for L and C can be recommended, since they vary widely with Z_0, line length, and swr. Their values are usually comparable to those used in the regular compensating circuits at the same frequency. Using this procedure makes it possible to couple practically any length of line to a transmitter, even when the swr is quite high.

13-994. MATCHING TO COAXIAL LINES.

13-995. GENERAL. Coaxial transmission lines are generally operated at a low enough swr so that no special matching circuits are needed. The line may be connected directly to the equipment output terminals. The output circuit of a transmitter should be capable of handling variations in the swr that are acceptable from the standpoint of line losses. However, there are some cases where it is

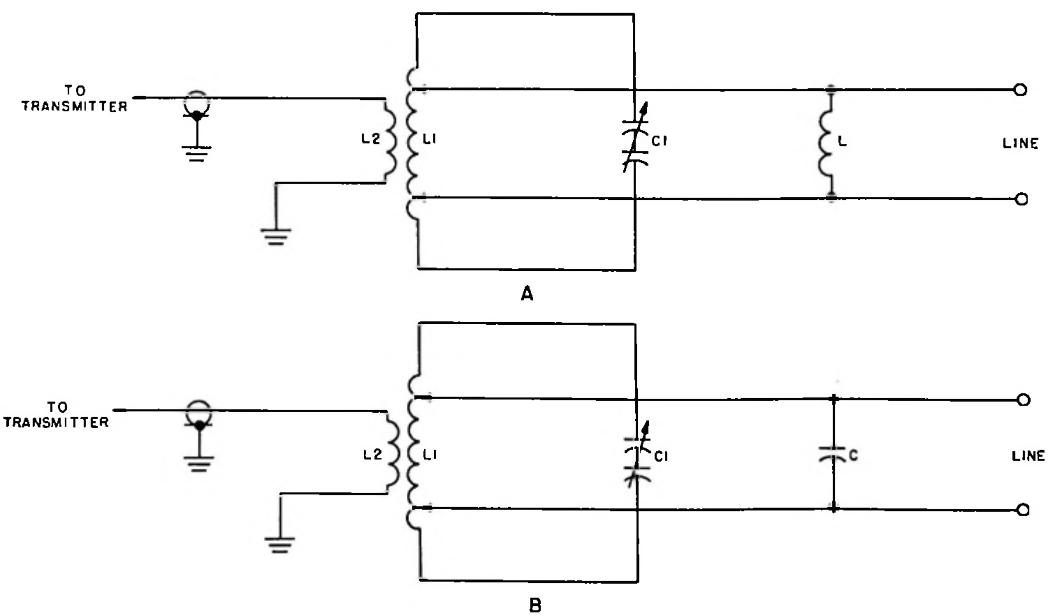


Figure 13-270. Reactance Cancellation on Random-Length Lines Having a High Standing-Wave Ratio

necessary to provide some frequency selectivity between the transmitter and the antenna to prevent undesired radiation of harmonics. A matching circuit of the same general type previously discussed can provide a considerable degree of selectivity, in addition to impedance-matching capabilities. The difference in circuit arrangement is simply that the secondary, or output side, need not be balanced to ground.

13-996. Figure 13-271 shows a typical circuit. Except for the fact that there is only one coil tap, the design and adjustments are the same as described for figure 13-267. Also, the series capacitor, C_2, shown in figure 13-268 may be used with this circuit for fine variation of the effective coupling

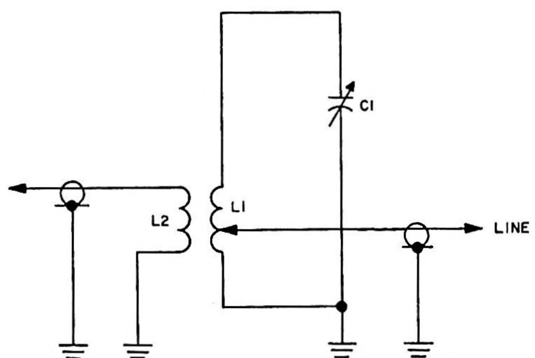


Figure 13-271. Inductively Coupled Matching Circuit for Coupling Between Coaxial Lines

between L_1 and L_2. The constants for circuit L_1C_1 are not critical. Any convenient values that will tune to the operating frequency may be used. The Q of the circuit, and hence the selectivity, is controlled principally by the position of the line tap. As the tap is moved farther up the coil, the Q and selectivity decrease.

13-997. HALF-WAVE FILTERS FOR HARMONIC SUPPRESSION. If impedance matching is not a consideration (i.e., the transmission line to the antenna is operating at a low swr), but harmonic suppression is desirable, the circuit of figure 13-272 may be used. This is a half-wave filter circuit with its input impedance equal to the impedance which the filter sees at its output. For example, if the line input impedance is a pure resistance of 50 ohms, the impedance at the filter output terminals is also 50 ohms. The characteristic impedance of the filter can be any value without altering its performance with respect to the input and output impedances. However, if broad-band operation is desirable, the filter characteristic impedance should equal the Z_0 of the line.

13-998. To insure effective operation, the two sections of the filter should be shielded from each other, as indicated by the dotted line, and the entire filter should be con-

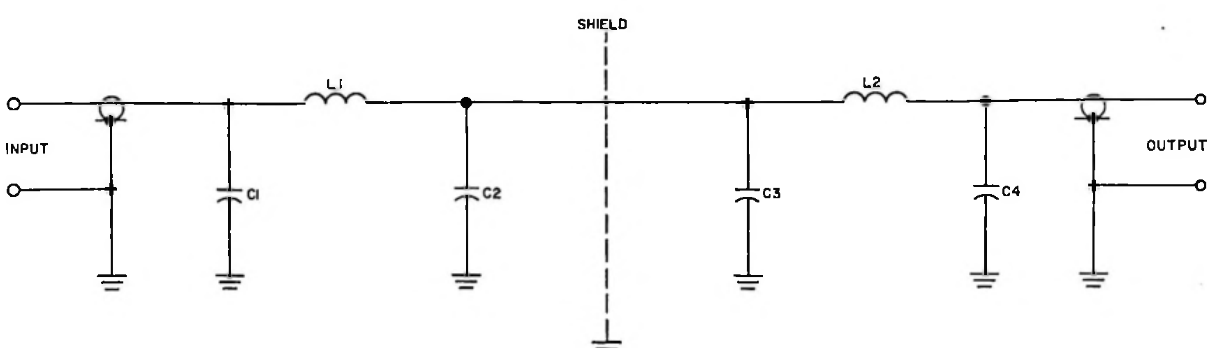


Figure 13-272. Half-Wave Filter for Harmonic Suppression

structed in a shielded enclosure. Component values are determined by the frequency of the particular equipment in use. The attenuation provided by a filter of this type is about 30 db at the second harmonic and even greater at higher harmonics, up to the point where the self-resonance that occurs in inductors at high frequencies limits the attenuation. The attenuation value is not important at harmonics below the fourth.

SECTION XIII

ALTERNATE TECHNIQUES FOR MEASURING REFLECTION COEFFICIENTS AND IMPEDANCES

13-999. WAVEGUIDE TECHNIQUES

13-1000. GENERAL.

13-1001. The waveguide vswr detector is the basic measuring tool for determining reflection coefficients and impedances in waveguide apparatus. However, there are times when alternative techniques, using simpler apparatus, permit a faster deduction of data. Techniques which are not commonly used with coaxial arrangements but which are especially adapted to waveguide components are discussed in the following text.

13-1002. PHASE SHIFTER AND FIXED PROBE.

13-1003. GENERAL. When waveguides operate at very short wavelengths, slot radiation difficulties can be reduced by associating a fixed probe with a phase shifter. The arrangement shown in figure 13-273 can be regarded as a standing-wave detector, where the electrical length of the arrangement between the probe and the unknown impedance is variable. Instead of mechanically moving the probe, a phase shifter varies the electrical length of the arrangement. The effect is exactly as if the probe were mechanically moved along the line. As the phase shift is varied, the probe voltage will be found to go through its high and low peaks. From this data the standing-wave ratio can be determined.

13-1004. INTERPRETATION. Introducing an additional phase shift of ϕ radians between the probe and the load has the same effect as moving the probe a distance (d_1) farther from the load, as derived from the equation:

$$d_1 = \frac{\phi}{2\pi}\lambda_g$$

where λ_g is the guide wavelength. Any convenient type of phase shifter may be used. Typical arrangements suitable for waveguide apparatus are the wave-squeezer, movable-vane, and rotary phase shifters. In the arrangement shown in the figure, the magnitude of the standing-wave ratio existing in the guide can be determined without knowing the actual amount of phase shift introduced. Simply observe the maximum and minimum values of probe response as the phase shifter is varied, and then divide the minimum voltage into the maximum voltage to obtain the standing-wave ratio (S). Once the standing-wave ratio has been determined, you can also calculate the reflection coefficient by using the formula:

$$|\rho| = \frac{S-1}{S+1}$$

where $|\rho|$ is the magnitude of the reflection coefficient, and S is the standing-wave ratio.

13-1005. PHASE-SHIFTER CALIBRATION. If the phase angle of the reflection coefficient

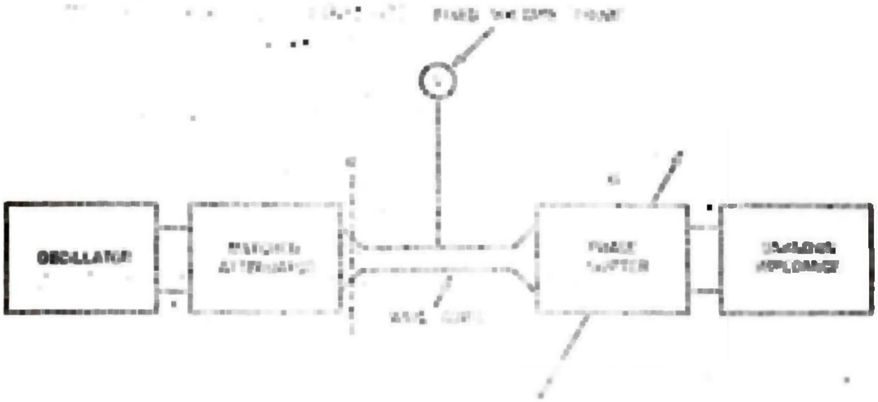


Figure 13-273. Phase Shifter and Fixed Probe Arrangement for Measuring an Unknown Impedance

is desired, or if the magnitude and phase of the load impedance are to be determined, it is necessary that there be available a calibration accurately giving the phase shift as a function of the setting of the phase-shifter control. To calibrate the phase shifter in this arrangement, you must first replace the load impedance with a short circuit. Under these conditions the relation between the voltage induced in the probe and the adjustable phase shift ϕ introduced by the phase shifter is:

$$\text{Voltage induced in probe} = A\left[\sin(\phi + a)\right]$$

where A is a proportionality constant determined by the design of the probe, and a is the total electrical phase shift existing between the probe position and the load point when $\phi = 0$. The value of a is determined by the distance in guide wavelengths from probe to load, augmented by any insertion phase shift introduced by the phase shifter when the phase-shifter control dial reads $\phi = 0$. A calibration of the phase shifter is then obtained by observing the relative voltage induced in the probe as a function of the setting of the phase-shifter dial. Settings that make the induced voltage a minimum then correspond to $\phi + a = n\pi$, where n can be any even integer (including zero), while the settings that make the induced voltage a maximum correspond to $\phi + a = m\pi/2$, where m is an odd integer. The phase shift corresponding to intermediate settings of the phase shifter can then be determined by introducing the observed variations in probe response into the equation, as follows:

$$\text{Voltage induced in probe} = A\left[\sin(\phi + a)\right]$$

Thus, when the probe voltage is 70.7 percent of the maximum voltage, ϕ has a value that differs in phase by 45 degrees from the value corresponding to a minimum response.

13-1006. ATTENUATOR PAD. In measuring standing-wave ratios with a phase shifter and fixed probe, it is necessary that a matched impedance be seen looking from the probe along the waveguide toward the oscillator. This can be accomplished by the use of an attenuation pad with an impedance that matches that of the guide. The amplitude of the incident wave is made independent of the presence or absence of a reflected wave. This is accomplished when the attenuator absorbs the wave reflected by the load impedance. The attenuator also isolates the oscillator from the measuring arrangement, thus preventing the oscillator frequency from changing as variations in the phase-shifter setting alter the input impedance at point a in the figure.

13-1007. HYBRID JUNCTION OR MAGIC-TEE

13-1008. FUNCTIONAL DESCRIPTION. The magnitude of the standing-wave ratio or the magnitude of the reflection coefficient can be determined by the "Magic-Tee" or hybrid junction method shown in figure 13-274. A wave entering waveguide A cannot pass directly to output waveguide B, but rather splits into two equal parts that travel down the side arms, C and D. Reflected waves in arms C and D reaching the junction will, however, enter arm B, producing a resultant wave in B that is the difference between the reflected wave in arm C and the reflected wave in arm D. If arms C and D are of equal length and are terminated identically, no output will appear in B. However, since the terminating impedances are not identical, some output will appear in B.

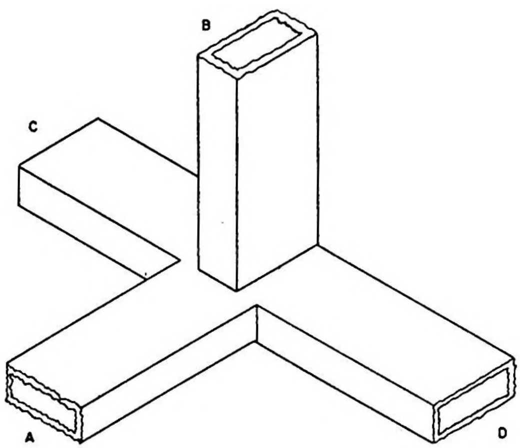


Figure 13-274. Magic-Tee Configuration

13-1009. A block diagram of the magic-tee configuration is shown in figure 13-275. Here a reference termination Z_o is connected at D, while the unknown impedance, Z_x, is used as the termination for arm C. Arms C and D are further made identical in every respect, particularly length. The oscillator supplies power to arm A, through an isolating attenuator, while the arm B output is observed by a crystal detector, preferably operated on the square-law part of its characteristic. A bolometer may be used in place of the crystal detector. Zero output will be observed in arm B when the unknown impedance, Z_x, is the same as the known reference impedance, Z_o. Thus, if Z_o has a value that provides a nonreflecting termination for arm D, then zero output indicates that Z_x is a nonreflecting termination for arm C. However, if an output is present in arm B when D is terminated in a nonreflecting termination, then the magnitude of the reflection coefficient introduced by Z_x is directly proportional to the magnitude of this output. The relationship involved can be readily obtained experimentally from the fact that when Z_x is replaced by a short circuit, the output magnitude present in B corresponds to a reflection coefficient of 1.0.

13-1010. USE WITH AN OSCILLOSCOPE. The hybrid junction arrangement shown in figure 13-275 is particularly useful where impedances must be compared over a wide frequency range, because the magic-tee is not frequency-sensitive. If the output of the amplifier in the figure is applied to the vertical input of an oscilloscope, and a sweep voltage from an oscillator is applied to the horizontal input, you can observe the presentation of the reflections existing on a waveguide arrangement as a function of frequency. The carrier frequency of the exciting oscillator is swept electronically by a voltage that also provides the horizontal deflection of the oscilloscope display. Thus, the horizontal

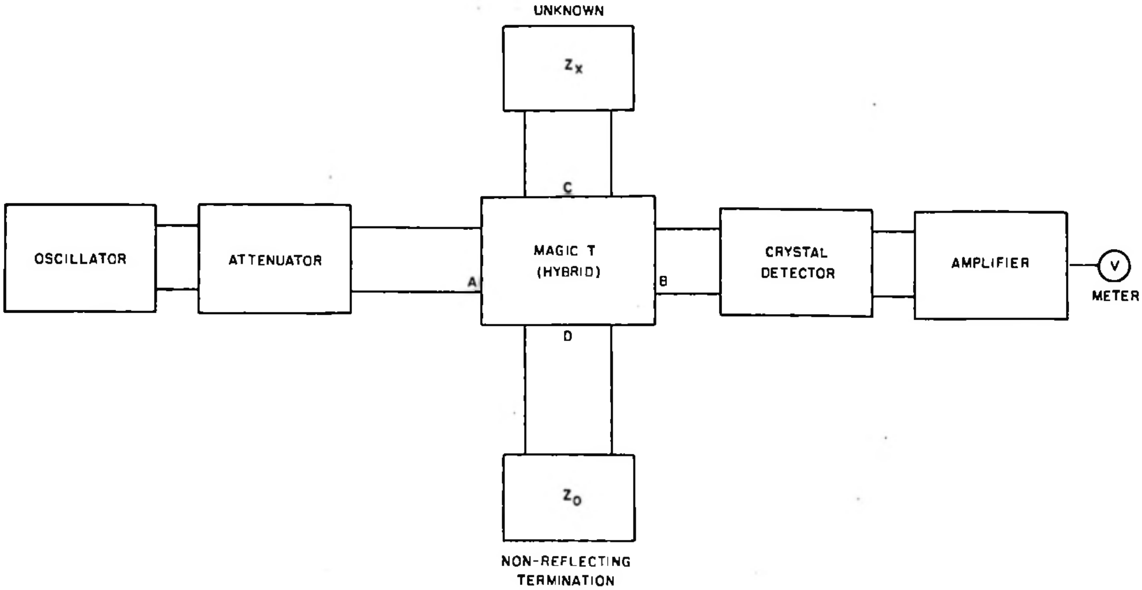


Figure 13-275. Magic-Tee Arrangement for Making Impedance or Reflection-Coefficient Measurements

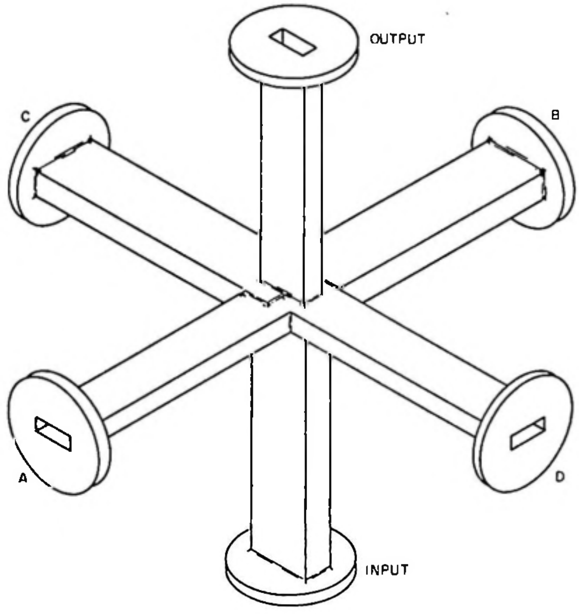
position on the presentation represents the frequency, while the vertical deflection represents the reflection coefficient. A scale of reference for interpreting vertical deflections can be obtained by placing a short circuit across arm C in place of Z_X, thus producing a reflection coefficient of 1.0 for all frequencies.

13-1011. The hybrid-junction technique requires relatively simple equipment; therefore, this technique is convenient to use. It will precisely compare two terminations that are almost identical; also, it will accurately determine very small values of reflection coefficients when a good nonreflecting termination is available.

13-1012. EARLY-TYPE IMPEDANCE BRIDGE.

13-1013. GENERAL. The six-arm waveguide structure shown in part A of figure 13-276 is a true microwave equivalent of a Wheatstone bridge. The resulting relation among the admittances of the various arms is exactly that of a Wheatstone bridge with shunting susceptances across each pair of terminals. With this bridge it is possible to measure practically any impedance to approximately the same accuracy as with a standing-wave detector. One of the important features of the six-arm waveguide structure is that the standard impedances required are three variable reactances (movable shorting plungers) and a Z_O termination. The data are obtained in the form of three lengths—the positions of the movable shorts. Because this device is the equivalent of a Wheatstone bridge, it can also be used as a four-terminal lattice section in filter design or in any other related application requiring the microwave equivalent of a lattice section.

13-1014. Any comparison method of measuring impedances at microwave frequencies



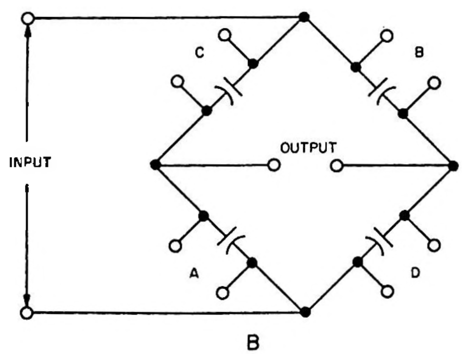


Figure 13-276. Six-Arm Impedance Bridge with Equivalent Circuit

requires an impedance standard and some sort of device which can function as a bridge.

$$AB = CD \quad \text{or} \quad B = \frac{C}{A} D$$

In a true Wheatstone bridge, A, B, C, and D are four impedances, three of which are known and the other unknown. The magic-tee or hybrid junction, previously discussed, is not a bridge, because it is necessary to have an adjustable impedance which will exactly equal the unknown impedance, and this requires both a variable resistance and a variable reactance. When microwave frequencies are involved, a variable resistance is difficult to provide. However, if a true bridge is used, it is possible to match practically any impedance by the use of a fixed resistance only, plus variable reactances. The bridge shown is made by joining six pieces of rectangular waveguide at a symmetrical junction. Four of the waveguides represent the arms of the bridge, and the two remaining waveguides represent the input and output. Opposite pairs of arms (A, B and C, D) lie on common axes, but are displaced 90 electrical degrees with respect to each other. A bridge of this type was constructed and tested. It was made of 3 x 1-1/2-inch waveguide, which is the standard size used in the 10-cm region.

13-1015. BASIC PRINCIPLES OF OPERATION. Assume that B is the unknown impedance. If A and C are variable reactances consisting of shorted, variable-length sections of transmission line, the ratio C/A can be adjusted to a desired value. D consists of a perfectly terminated line in series with a variable reactance, which is another shorted-section line. By adjusting the shorted-section line, D can be made to have any arbitrary phase angle and can be adjusted to have the phase angle of the unknown, B. The ratio C/A can now be adjusted to provide the correct magnitude. The impedance of the unknown, B, can then be determined from the three physical lengths of stubs A, C, and D.

13-1016. Once properly calibrated, the bridge is simple to use, and measurements can be performed rapidly. The unknown impedance is calculated in terms of three physical lengths—the distances of the shorts from their correct reference planes. This is a simple standard, and is independent of detector calibration. The accuracy of the measurement is equivalent to that of a good swr detector. The use of the three physical lengths to specify the unknown impedance can be facilitated with the aid of formulas and charts. With conventional broad-banding techniques it is possible to make the device cover quite a broad band, so that the fringing susceptance is almost zero over a considerable range of frequency.

13-1017. As mentioned previously, this device can be used as a four-terminal lattice section in microwave filter design or in similar applications. Generally, microwave filter structures have been the equivalents of ladder sections. However, a lattice section offers greater flexibility in design, because the image impedances and transfer constant can be separately adjusted. Also, physical four-terminal networks can be synthesized by means of lattice sections, while this is not always true for ladder sections.

13-1018. SWEPT-FREQUENCY IMPEDANCE INDICATOR.

13-1019. GENERAL. The swept-frequency impedance indicator, shown in figure 13-277, was designed to provide a method of determining the magnitude and phase of an unknown impedance by presentation on an oscilloscope. This is accomplished at a number of closely spaced frequencies over a 12-percent frequency range centered in the 3-cm band. The impedance indicator contains a signal generator, a wave sampler with load, and a voltage-ratio computer. The output of the voltage-ratio computer supplies the vertical deflection voltage for a cathode-ray tube, while the signal generator supplies the horizontal deflection voltage for the tube.

13-1020. OPERATION. A variable-frequency signal generator with a reasonably constant output applies a signal to a four-terminal waveguide circuit, called a wave sampler. The wave sampler is equivalent

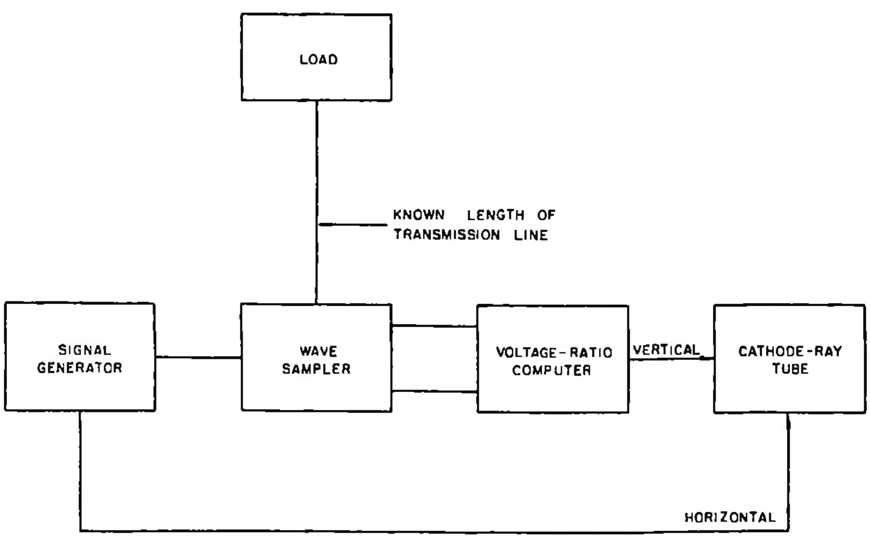


Figure 13-277. Swept-Frequency Impedance Indicator

to a section of waveguide having on it two identical, uncoupled short probes one-quarter wavelength apart at all frequencies. A known length of transmission line receives the signal from the wave sampler and applies the signal to the load containing the unknown impedance. Two voltages are developed in suitable detectors; these voltages fed, after amplification, to a continuous computer, which has an output proportional to the ratio of these voltages. The output voltage is applied to the vertical deflection plates of a cathode-ray tube. The voltage on the horizontal plates of the tube is synchronized with the frequency of the signal generator.

13-1021. WAVE SAMPLER. The wave sampler shown in figure 13-278 is fundamental to the accuracy and bandwidth of this arrangement. The signal generator is applied to terminal A, and the load is connected to terminal B. When terminals F and D are well matched, signals proportional to $A^2 + B^2 - ZAB \cos \phi$ and $A^2 + B^2 + 2AB$

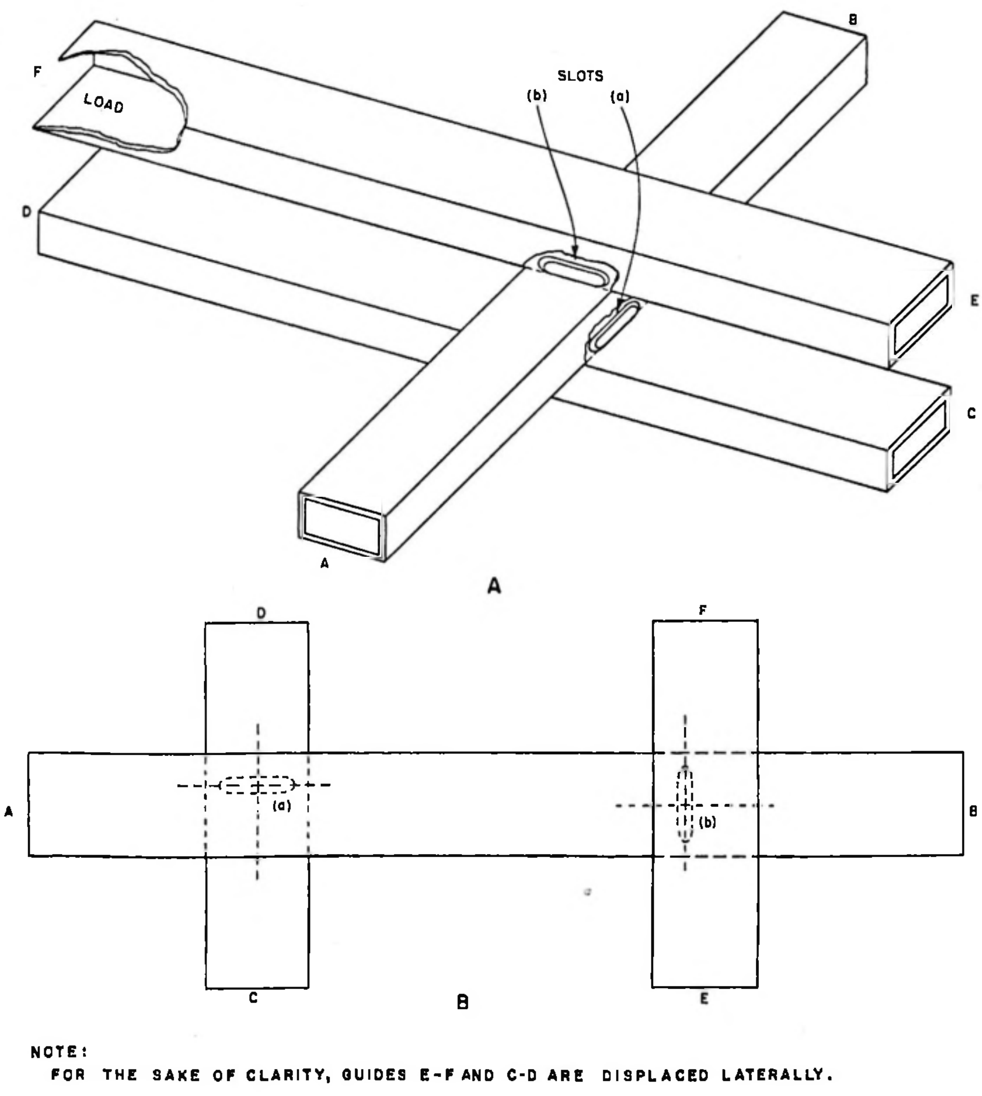


Figure 13-278. Wave Sampler

cos ϕ are obtained from terminals E and C. This is accomplished by cutting slots (a) and (b) in the main guide so that their center falls perpendicular to the axis of the guide. Slot (b) is excited by the total longitudinal current flow in the main guide (A to B), while slot (a) is excited by the total transverse current flowing in the main guide (C to D). The excitation of slot (b) is proportional to the vector difference of the incident and reflected voltages, whereas the excitation of slot (a) is proportional to the vector sum of the incident and reflected voltages. The principles by which the impedance of a termination is determined with this arrangement are identical to those employed in slotted-line measurements. The use of the wave sampler and the rather wide frequency variations combine to make any mechanical motion unnecessary. Consequently, the speed with which the data may be presented on the cathode-ray tube is limited by the time constants of the electronic circuits and the characteristics of the frequency-modulated signal generator.

13-1022. VOLTAGE-RATIO COMPUTER. The voltage-ratio computer is a two-channel amplifier arranged so that an automatic gain control, simultaneously acting on both channels, holds the output of one of the channels constant. The output of the other channel gives the ratio of the input signals. Identical tube types are used in both channels to minimize the effects of tube unbalance. The numerator channel operates at 1000 cps, and the denominator channel operates at 3000 cps. These frequencies are obtained directly from the repetition rate of the signal generator. The output of the voltage-ratio computer can be calculated from the following formula:

$$\frac{a^2 + b^2 + 2ab \cos \phi}{a^2 + b^2 - 2ab \cos \phi}$$

The result is applied to the vertical deflection plates of the oscilloscope.

13-1023. Tune the signal generator across the desired band, and record the maximum and minimum values of the computer output as indicated by a voltmeter, together with the frequencies at which they occur. Normalization of these values to the values obtained with a matched load at the same frequencies then allows you to determine the standing-wave ratio over the frequency band. The phase of the load impedance is determined from the electrical length of the wave sampler and the frequencies at which the maximum and minimum voltage amplitude values occur.

13-1024. It is fundamental to an understanding of the operation of this device to know that the phase (ϕ) of the reflected voltage relative to the incident voltage (measured at the wave sampler) is a function of the following factors:

a. The frequency.

b. The length of line separating the load from the slots of the wave sampler.

c. The phase of the impedance of the load.

13-1025. At 3 cm, for a transmission line whose length approximates 3 feet, ϕ changes sufficiently with λ for reasonably broadband loads, so that the impedance (Z) representing the ratio (R) takes on alternately the values $(a + b)^2 / (a - b)^2$ and $(a - b)^2 / (a + b)^2$ about once for every 1-percent change in wavelength. These are true values of the square of the standing-wave ratio and its reciprocal, respectively. All other values of Z lie between these extremes. True values of standing-wave ratios are given only at those wavelengths which correspond to maxima and minima of the trace. When the phase of the reflected wave at the wave sampler and the length of the transmission are known, you can determine the phase and magnitude of the impedance of the load at those wavelengths.

13-1026. CIRCULAR-POLARIZATION COUPLER.

13-1027. A broad-band impedance measurement that is particularly suitable for use in automatic or manual impedance-measurement apparatus is the circular-polarization coupler. This device consists of an appropriate length (Y) of circular waveguide attached at right angles to the broad wall of a rectangular waveguide, and coupled to it by three apertures. High accuracy over approximately a 20 percent bandwidth is maintained. By rotating a probe one turn in the circular waveguide, the probe can sample a field variation exactly equivalent to that in one guide wavelength of the rectangular waveguide. This impedance-measurement instrument has the ability to reproduce a standing wave in the rectangular waveguide as a field-strength variation around the circumference of the circular waveguide. The varying field has a maximum-to-minimum ratio equal to the swr in the rectangular waveguide. One rotation in the circular guide is exactly equivalent to one rectangular-waveguide wavelength at all frequencies in the operating band. This means that the angular position of the minimum value in the circular waveguide varies linearly with the position of the minimum in the rectangular waveguide, with the circular variation in mechanical degrees being equivalent to the rectangular variation in electrical degrees.

13-1028. The standing wave may be displayed on an oscilloscope by operating the probe at a constant speed, or the swr and minimum position may be determined by operating the probe manually. The coupler can also be used to indicate or directly record the impedance as the frequency is swept over the 20-percent operating range of the instrument. The latter method, however, requires a more complicated arrangement. By comparison with slotted-wave impedance meters, the circular-polarization coupler has the operational convenience of rotary motion with maxima and minima always separated by exactly a quarter of a turn. The rotary motion and the constant equivalent wavelength make it more easily adaptable to automatic instrumentation. In comparison with reflectometers, the circular coupler offers readily available phase data and does not require a well-matched termination. The only important disadvantage is the limitation of the bandwidth to approximately 20 percent within the tolerance factor (ρ) near units.

13-1029. In part A of figure 13-279, the two z slots couple the horizontal depth (H_x) fields in the two waveguides, while the x slot couples the respective horizontal width (H_z) fields. Hence, these slots excite TE_{11}-mode waves in the circular waveguide, which are orthogonal in space relationship. If a pure traveling wave were in the rectangular waveguide, H_x and H_z would be 90 degrees out of phase; therefore, the two TE_{11} waves would also be 90 degrees out of phase. These facts, plus an additional stipulation that the amplitudes of the TE_{11} waves must be equal, are conditions for the excitation of a circularly polarized wave. The equality of the amplitudes may be obtained by properly positioning and dimensioning the slots.

13-1030. A TE_{10} wave generating in the +z direction will produce a right-hand circularly polarized wave (clockwise rotation when viewed from the rear). Alternately, a TE_{10} wave generating in the −z direction will produce a left-hand or counterclockwise circularly polarized wave. When the rectangular waveguide is connected to a mismatched impedance, waves in both the +z and −z direction are present; hence, both right- and left-hand circularly polarized components exist in the circular waveguide. The resultant of the combination of these

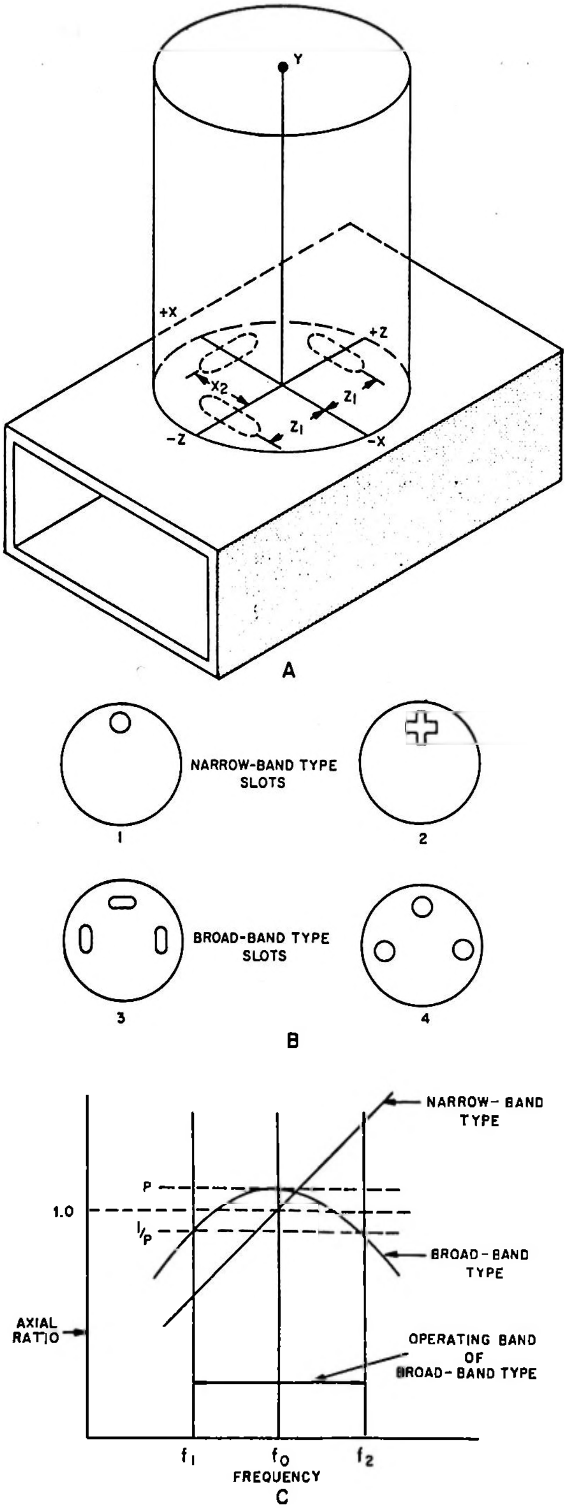


Figure 13-279. Circular-Polarization Coupler with Apertures

waves is an elliptically polarized wave whose axial ratio is equal to the ratio of the sum and difference of the component amplitudes, which, in turn, are proportional to the respective TE_{10} amplitudes in the rectangular guide. Therefore, it follows that the axial ratio is equal to the swr in the rectangular waveguide. Also, the angular position of the major or minor axis of the polarization ellipse will vary linearly with the angle of the reflection coefficient of the load; a 360-degree change in the latter produces a 180-degree change in the former. In other words, the minor axis of the ellipse will vary linearly with the minimum position in the rectangular guide; a one guide-wavelength change in the latter corresponds to a 360-degree change in the former. It is evident, then, that all information concerning swr, impedance, and reflection coefficient of a load connected to the rectangular waveguide may be obtained by means of a probe rotated in the transverse plane of the circular waveguide.

13-1031. The first two types of apertures in part B of the figure are narrow-band designs yielding an axial ratio proportional to λg (wavelength in the rectangular waveguide). The last two types of apertures yield a compensated axial-ratio response by means of a proper spacing of apertures along the z axis. Part C of the figure illustrates the proportionality of the axial ratio developed in the following formula:

$$\frac{\cos (z\ \pi z l / \lambda g)}{\lambda g}$$

This formula provides a very close approximation to a unity axial ratio over a wide frequency range. In the case of unity swr the axial ratio may be held within plus or minus 2 percent of unity over approximately a 20-percent frequency band. An elliptically polarized wave has a characteristic in which the major and minor axes are fixed in direction in all transverse planes. Therefore, the minimum-signal position of the rotatable probe is independent of the particular transverse plane in which the probe is located. However, the probe should be far enough from the apertures to insure that all higher modes excited by the apertures have decayed to negligible amplitudes. An adequate distance is normally three times the waveguide diameter. By measuring the angular position of the minimum from an axis parallel to the x axis (part A of figure 13-279), the angle of the reflection coefficient at $z = 0$ is equal to 2ϕ.

13-1032. The end of the circular waveguide need not have a matched load. However, the load must be symmetrical with respect to the probe. This may be understood by representing the elliptical wave by means of two orthogonal, linearly polarized components, one parallel to the probe and the other at right angles to it. If the parallel component sees a mismatch, repetitive reflections will occur; this will result in a certain degree of increase or decrease of the picked-up signal, depending on the degree of mismatch and the length (y) of the circular waveguide. Since the end wall containing the small apertures is virtually a perfect short circuit to a wave returning at any angle, the relative change in signal due to the reflections will be independent of the probe angle, and the axial ratio and minimum position indicated by the probe will not be affected. The perpendicular component to the probe will be reflected repeatedly until it is dissipated. If the load were not symmetrical with respect to the longitudinal plane through the probe, cross coupling between the components might occur; thus errors might be introduced.

13-1033. METHODS USED ON COAXIAL LINES.

13-1034. BYRNE BRIDGE.

13-1035. GENERAL. The Byrne bridge device, shown in figure 13-280, is a null-type impedance meter that provides a direct-reading in either impedance or admittance. The oscillator applies power through a coaxial line to the unknown impedance. At a point on this line as close as is feasible

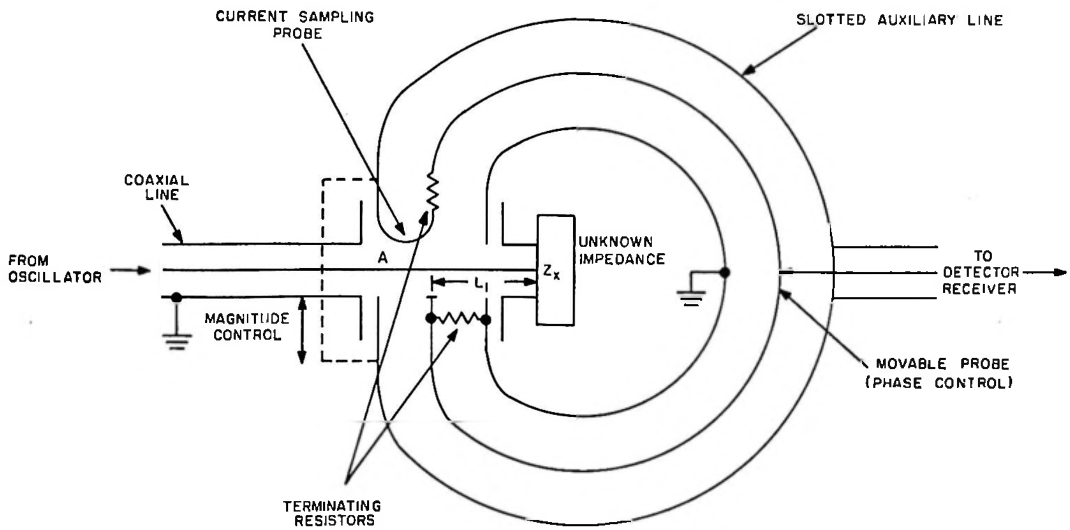


Figure 13-280. Byrne Bridge

to the load, two sampling elements are introduced; these elements are responsive, respectively, to the voltage and the current flowing into the load. The element that is responsive to voltage is a capacitive probe, and the element that is responsive to current is an inductance loop probe. The amount of pickup by each sampling probe can be adjusted by moving the probe with respect to the coaxial line. In practice, the two probes are joined together so that when one probe moves in toward the center conductor of the line, the other moves away from it. A slotted auxiliary line connects the two sampling probes. The voltage induced in each sampling probe initiates a wave which starts from the probe and travels around the auxiliary line, where it is entirely absorbed in the terminating resistance at the opposite end. The movable probe in the auxiliary line measures the net voltage in the line resulting from the super-position of the two waves traveling in opposite directions. The relative magnitudes of the two waves induced in the auxiliary line depend both upon the vector ratio of voltage to current in the unknown impedance and upon the relative positions of the sampling probes.

13-1036. OPERATION. To determine the unknown impedance, Z_X, set the detector output to zero by positioning the movable probe with one hand and varying the sampling probe with the other hand. Under these conditions, the two waves induced on the auxiliary line are equal in magnitude and are in phase opposition at the position of the movable probe. The setting of the sampling probe is determined by the ratio of voltage to current flowing into the unknown impedance, i.e., by the magnitude of the impedance to be determined, and is independent of the frequency. You can also calibrate this dial to read impedance directly, and the phase angle of the unknown impedance can be determined by the position of the movable probe corresponding to the null condition. If the load is resistive, the null position will be midway between the two sampling probes. The phase angle of the impedance being measured will be proportional to the displacement from the reference position corresponding to a resistive load. The phase angle being measured will also be proportional to frequency. Thus, the indicator providing the position of the movable probe can be calibrated directly in phase angle at some convenient frequency. The impedance measured corresponds to the voltage and currents existing at the sampling probes (point A toward Z_X).

13-1037. The Byrne bridge is equivalent to an ac bridge, and gives the vector value of the unknown impedance. Because of the critical tolerances incorporated during the manufacturing process, this device has good accuracy over a wide frequency range. The ease of operation, as contrasted with that of the swr detector, is very significant.

13-1038. NODE-SHIFT METHOD.

13-1039. GENERAL. When a standing-wave detector constructed from a length of large-diameter rigid coaxial line is to be used in measurements on small-diameter flexible coaxial cable, as shown in part B of figure 13-281, the junction between these two types of line should contain as few reflections as possible. The discontinuity capacitance will cause some reflection at the junction, even when the two lines have the same characteristic impedance. Small reflections are difficult to determine by the usual method of standing-wave measurement. These reflections could be evaluated by ordinary standing-wave measurement techniques, by terminating line 2 in its characteristic impedance so that there are no reflections in this line, and then measuring the standing-wave ratio present on line 1. However, the standing-wave ratio will be close to unity if the junction is well designed. The exact value will be difficult to determine, because mechanical irregularities in the traveling-wave probe may introduce variations in the voltage

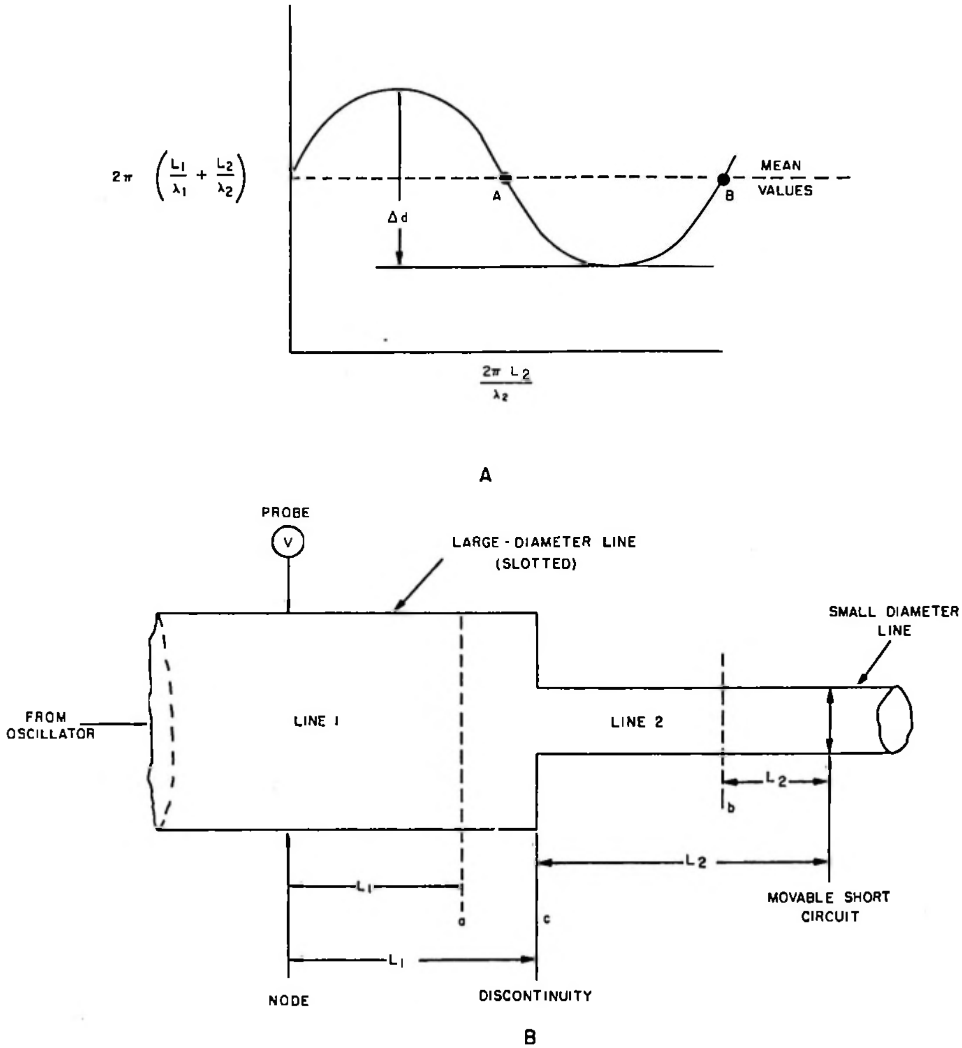


Figure 13-281. Node-Shift Method

readings that will be as great as those due to the maxima and minima of the standing-wave ratio.

13-1040. APPLICATION. Even though mechanical irregularities may cause fluctuations in the magnitude of the probe voltage, the position of the probe where a voltage node or minimum occurs is not influenced. The following procedure describes how to use the node-shift method:

a. Short-circuit line 2 as shown in the figure.

b. Determine the position of the voltage node in line 1 by moving the probe to the left and to the right of the voltage minimum value until a small but appreciable voltage reading of equal amount is obtained on either side of the minimum voltage point. An average of these two positions is the correct location of the node.

c. If a small superimposed reflection appears, the position of the node in line 1 will vary sinusoidally, as shown in part A of the figure. The magnitude of the reflection coefficient at the junction can be determined from the following equation:

$$|\rho| = \sin \frac{\Delta d}{2} \approx \frac{\Delta d}{2}$$

where Δd = total variation in the quantity $2\pi\left[\frac{L1}{\lambda 1} + \frac{L2}{\lambda 2}\right]$

$\lambda 1$ = line wavelength for line 1.

$\lambda 2$ = line wavelength for line 2.

L1 = position of a minimum on line 1 measured with respect to a convenient reference point, such as a or c in part B of the figure.

L2 = position of short circuit on line 2 measured with respect to a convenient reference point, such as b or c in part B of the figure.

The corresponding standing-wave ratio is:

$$S \approx 1 + \Delta d$$

d. If no reflection is produced at the junction, the position of the node in line 1 would move a distance exactly equal to the movement of the position of the short circuit in line 2.

13-1041. The magnitude of the reflection coefficient can be determined without an exact knowledge of the location of the discontinuity, since Δd can be determined when L1 and L2 are measured from arbitrary fixed positions such as points a and b in part B of the figure. Thus, the presence and magnitude of a discontinuity of unknown origin can be detected and evaluated without a precise knowledge of its location.

13-1042. The phase angle of the reflection coefficient at the junction can be determined with the aid of data corresponding to points A and B of part A in the figure. The appropriate data can be calculated from the following:

$$\phi = \frac{2\pi}{\lambda 1} L1_a = \frac{2\pi}{\lambda 1} L1_b - \frac{\pi}{2}$$

where

$L1_a$ = distance L1 from discontinuity junction c to the node point in line 1 corresponding to point A in part A of the figure.

$L1_b$ = distance L1 from discontinuity junction c to the node point in line 2 corresponding to point B in part A of the figure.

Unlike the magnitude of the reflection coefficient, the phase angle, ϕ, of the reflection can be determined only if the location of c (in part B of the figure) of the irregularity producing the reflection is known. This technique has been described in terms of coaxial transmission lines, but the method is completely adaptable to two-wire transmission lines and to waveguides.

13-1043. PROBE METHODS.

13-1044. The standing-wave ratio or reflection coefficient, from which the load impedance can be determined, can be calculated from the output of fixed capacitance probes in various arrangements. One arrangement consists of three probes spaced along the line at eighth-wavelength separations. The magnitude and phase of the impedance terminating the line can be computed from the radio-frequency voltages observed at the three probes. An alternative arrangement uses four probes, and yields two voltages, respectively proportional to the real and imaginary part of the reflection coefficient. The probes are used in pairs, each pair having a quarter-wavelength separation, and one pair being an eighth-wavelength nearer the load than the other. A crystal detector, with a [illegible] characteristic, rectifies the radio-frequency voltage at each probe. From one pair of probes the difference of the two rectified outputs of the detectors is applied to the vertical-deflection plates of an oscilloscope. Similarly, the corresponding output of the other pair is applied to the horizontal-deflection plates. The resulting display of the oscilloscope presents both the magnitude and phase of the reflection coefficient. This method is also convenient for rapid measurements on resonant cavities and other terminations.

13-1045. Both of these methods for determining the magnitude and phase of the reflection coefficient are essentially single-frequency methods. They cannot be used over a frequency band exceeding 10 percent of the frequency at which the probe spacing is exactly an eighth-wavelength.

13-1046. TRANSMISSION-LINE IMPEDANCE BRIDGE.

13-1047. When it is sufficient to compare an unknown impedance (in both magnitude and phase) with a known impedance at frequencies so high that ordinary bridges or null networks cannot be used, it is possible to resort to the hybrid or magic-tee junctions previously discussed. Another method which can be used at high frequencies, when it is desirable to detect small differences between an impedance under study and a known adjustable impedance, is the use of a bridge involving transmission-line links. An example of a transmission-line bridge is shown in figure 13-282. Voltages exactly equal in magnitude and opposite in phase are applied to coaxial lines A and B, respectively, while the two impedances Z_s and Z_x that are to be compared are connected to coaxial side arms E and G. If the arrangement is perfectly symmetrical electrically, then $Z_g = Z_x$ and there will be no output at arm F. Electrical symmetry can be checked by making $Z_g = Z_x$. This is accomplished by [illegible] whether or not the output at F is zero. If it is not zero, the balance of the input transformer may be adjusted to give a null-output condition, after which Z_g and Z_x may be connected in place and compared.

13-1048. COMPARATORS.

13-1049. GENERAL. Intermediate between bridges and directional couplers are devices classified as comparators. In bridges, the currents in the unknown and reference arms balance directly; in directional couplers, the current is sampled in the unknown arm. Comparators include both a junction where unknown and reference currents add, and a sampling device for comparing a small fraction of these currents. Two types of comparators for coaxial-line service are discussed in the following text.

13-1050. WOODWARD COMPARATOR. A single rotating loop is coupled to the arms of the tee through slots cut parallel to the H-field, as shown in part A of figure 13-283. The transverse slots are made to eliminate capacitive coupling. The operation of this device is similar to that of a slotted line. Set the loop so that the generator input will provide an output deflection of $I_d = 1$. The reading Id at the 0-degree position is then numerically equal to the reflection coefficient, ρ. The measurement of impedance or admittance with a Woodward comparator is more complicated than with a slotted line, and it is less accurate. Therefore, this type of comparator is of limited value in general line measurements.

13-1051. ADMITTANCE COMPARATOR. Another type of comparator, called admittance comparator, is the null-type device

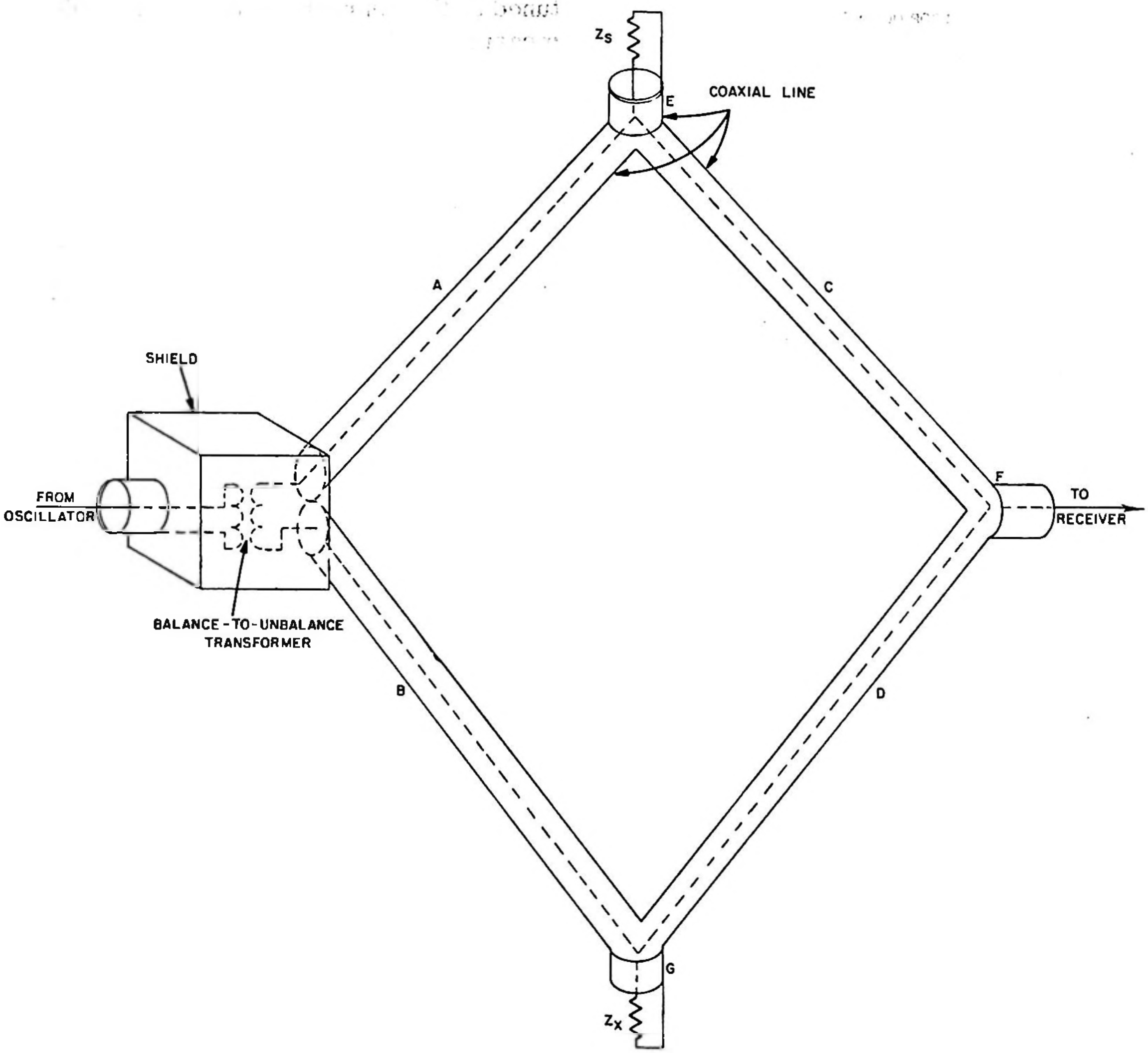
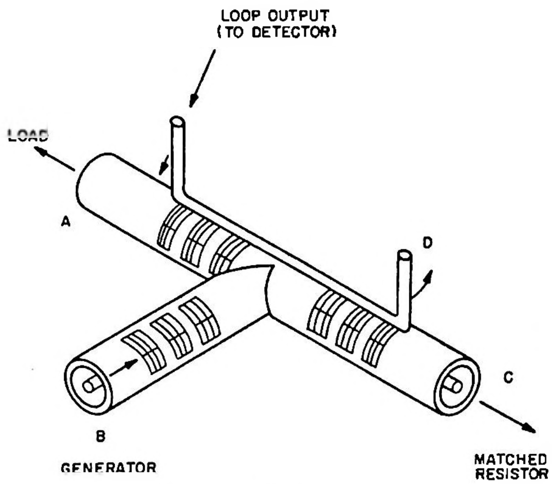
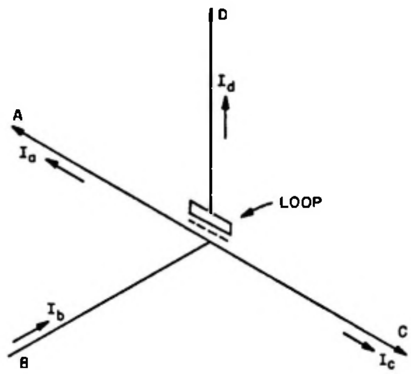


Figure 13-282. Simple Transmission-Line Impedance Bridge

shown in figure 13-284. The power from the oscillator is applied through the coaxial line (A). It then branches into three coaxial lines, one of which (E) is terminated by the unknown impedance; this line is kept as short as possible so that its input impedance at the junction point is almost equal to the unknown impedance. The second branch (B), or susceptance branch, is arranged to offer a susceptance corresponding to 50 ohms inductive reactance at the junction of the three lines. This is accomplished by giving the susceptance branch a characteristic impedance of 50 ohms, and then adjusting the length of the branch to be exactly an eighth-wavelength at the frequency being used. The third, or conductance, branch (D) also has a 50-ohm characteristic impedance, but is terminated by a 50-ohm resistance so that this branch always offers an input impedance of 50 ohms.



A



B

Figure 13-283. Woodward Comparator

13-1052. The same voltage is supplied to each of the three branches (E, B, and D) from the oscillator. This causes branch currents which at the junction have relative magnitudes and phases as determined by the input admittances of the three branches. These currents are sampled at each branch by coupling loops which can be rotated to vary the amount of current induced. The three loops are connected, in parallel, to the detector, which can be a radio receiver tuned to the appropriate frequency. The measurement is made by rotating the three loops until the three currents add to zero, giving a null in the receiver output.

13-1053. OPERATION. The principles involved in the operation of the admittance comparator can be understood from the following procedure:

a. Assume first that the loop in the unknown impedance arm (E) is kept fixed at the position of maximum coupling. Normally, this is when it is parallel to the center conductor.

b. Rotate the conductance loop (D) (varying the amount of in-phase current available from the conductance standard) to balance the in-phase current for the unknown loop (E).

c. Similarly, rotate the susceptance loop (B) (varying the amount of reactive current available) to balance out the reactive current induced in the unknown loop (E). By rotating the loop 180 degrees, the reactive current can be made either positive or negative. Thus, for a given setting of the unknown loop (E), the conductance and susceptance components of the unknown impedance are determined by the settings of the conductance and susceptance loops, respectively, when you have a null.

d. Directly calibrate the controls for these two loops, in mhos, on the assumption that the unknown loop (E) is at a standard reference position.

e. Rotate the unknown loop (E) from its standard reference position. This changes the magnitude of the induced current that must be balanced out by the other two loops. Hence, the control for the unknown loop (E) may be calibrated to provide a multiplying factor.

13-1054. The calibration of the admittance comparator is independent of frequency to

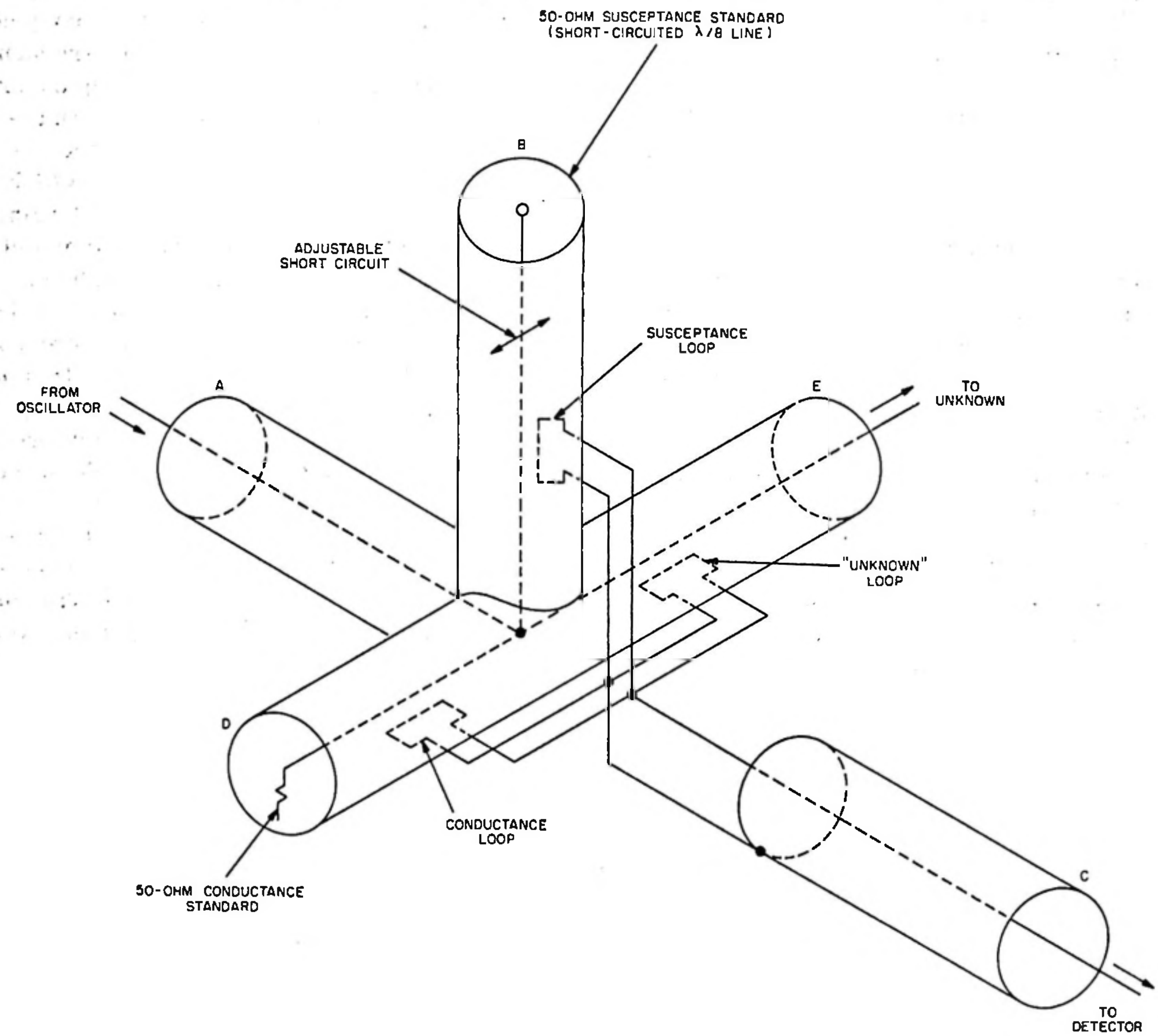


Figure 13-284. Admittance Comparator

the extent that the distances from the junction to each of the three loops are small as compared with a wavelength. This sets an upper limit to the useful frequency range. At low frequencies the only limiting factor is the length of the susceptance stub, which becomes impractically long if the frequency is quite low.

13-1055. A commercial form of the admittance comparator has the T junction and coupling loop contained within a cube about 2 inches on a side, and has accuracy on the order of 5 percent up to a frequency of about 1000 mc when measuring admittances on the order of 0.02 mho. This device is designed to have a low-frequency limit of 70 mc, and has a working range of 0.004 to 0.1 mho (250 to 10 ohms) with an accuracy on the order of 10 percent or less.

13-1056. SWEPT-FREQUENCY METHOD.

13-1057. Combining a frequency-swept source with a delay line can give you a flexible method for measuring low values of swr. The resulting reflection-coefficient indicator offers moderate accuracy and a convenient cathode-ray-tube display. For a source, a triode modulated by a vibrating capacitor or a pair of klystrons (one modulated by a sawtooth wave) is practicable. Connect the load to be measured, as shown in figure 13-285, through an appropriate length of low-loss cable. The cable should provide a delay of several tenths of a microsecond. Connect a crystal detector in parallel with the source. This will detect the beat note between the outgoing and reflected signals. The beat frequency is the result of the frequency sweep rate with respect to time in conjunction with the delay time (T). The frequency resolution is provided by the sweep excursion per cycle of beat frequency. An attendant loss will eventually mask the reflections at the far end of the line, which will limit the maximum delay-line length. Delays up to 1 microsecond are commonly used, corresponding to cable lengths of several hundred feet. For example, a 30-megacycle sweep excursion will correspond to 30 cycles per sweep, or a resolution of 1 megacycle per cycle of beat frequency. The height of the envelope on the crt presentation corresponds to the magnitude of the reflection coefficient, and the horizontal or x-axis (abscissa) corresponds to the frequency. The phase of the reflection coefficient may be estimated from the shift in the pattern when a standard replaces the unknown termination.

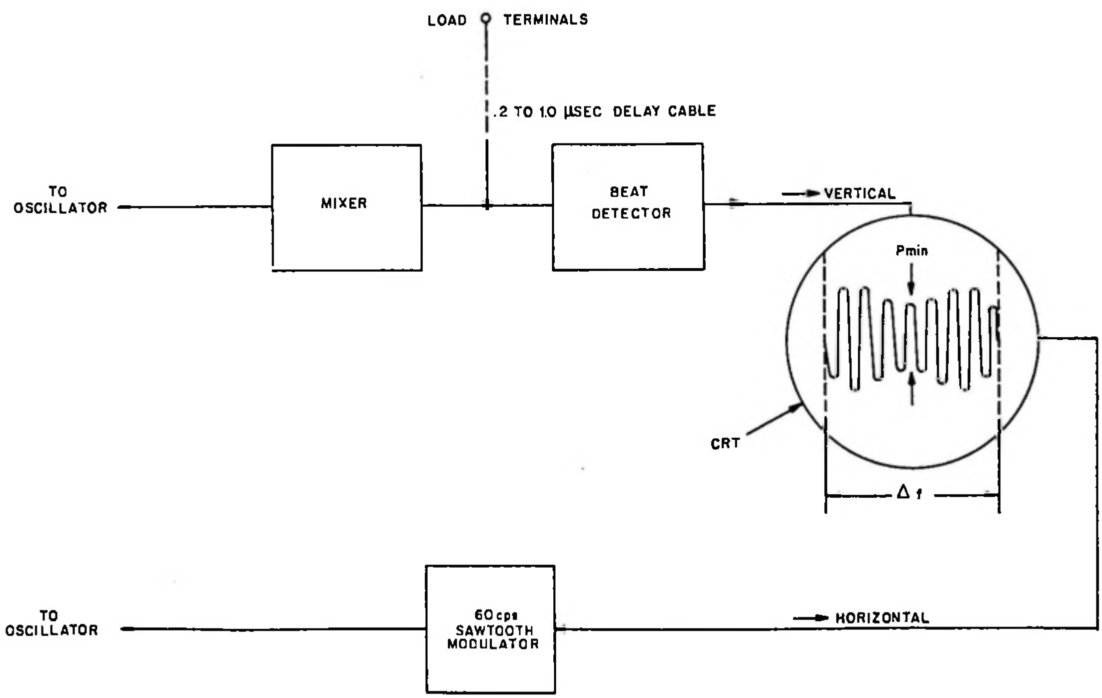


Figure 13-285. Frequency-Scanning Reflection Meter

13-1058. REFLECTED-PULSE TESTING.

13-1059. In most cases, the reflected-pulse method of testing transmission lines can eliminate the need for bridge-type measuring equipment. This type of check can be quickly performed with highly accurate results, and has an advantage over the bridge type in that it can locate several line defects at one time. The testing of transmission lines with reflected pulses is a method ideally suited to the needs of most radar stations.

13-1060. Each type of transmission line has its own characteristic or surge impedance, referred to as Z_0. As shown in part A of figure 13-286, L is the series inductance of the line and C is the distributed capacitance between the two conductors. This same transmission line can be reconstructed (part B) by using lumped values of inductance and capacitance. By the addition of identical sections, the length of the line can be effectively increased without a change in Z_0, since the L/C ratio remains constant. The only change is the time required for energy to travel from the generator end of the line to the load end, as computed from the formula $T = N\,2\pi\sqrt{LC}$, where T is the time required for energy to travel the length of the line, N is the number of line sections, and LC is the product of one section. From this you can see that the time for an electrical impulse to travel from one end of the line to the other depends not only on the length of the line, but also on the time required to change the inductance and capacitance of the line. The charging time accounts for the fact that energy travels along a line at a velocity less than the speed of light. The velocity factor is the ratio of line velocity to the speed of light, and is normally expressed as a percentage. Velocity factors for common lines are shown in table 13-8.

13-1061. When a transmission line is permitted to discharge into an impedance other

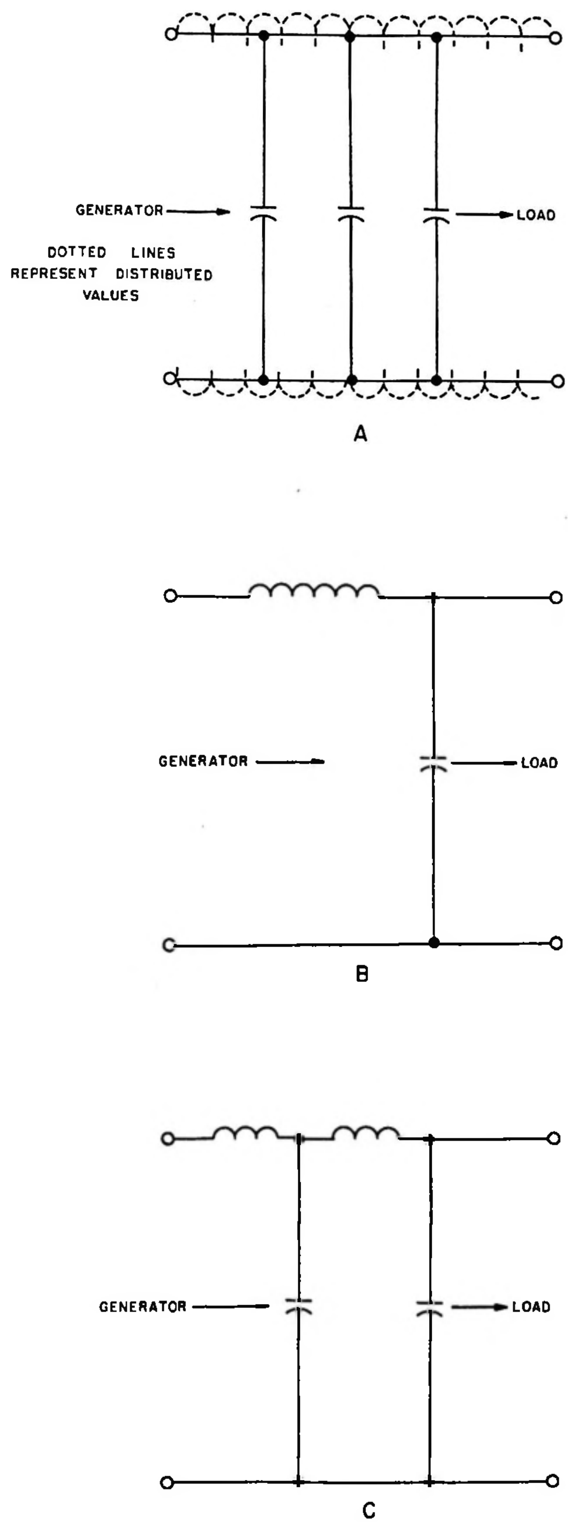


Figure 13-286. Distributed LC on Two-Wire Line

Table 13-8. Common Transmission Line Data

TYPE	Z_O (ohms)	VELOCITY FACTOR	ATTENUATION (db/100 ft) AT 3.5 MC
RG-8/U	53	.66	.28
RG-58/U	53	.66	.53
RG-11/U	75	.66	.27
Twin Lead	300	.84	.37
Zip Cord	110	.60	.96

than its own Z_O, the energy not dissipated in the impedance is reflected back into the line. The only time a line will produce no reflections is when it is terminated in an impedance equal to Z_O. A line of this type will have all of its energy absorbed by the impedance, and is said to be matched. When rf energy is traveling in an unmatched line, the reflections will result in standing waves, the magnitude of the reflected waves depending on the amount of mismatch. All energy will be reflected with either an open or a shorted line, and the polarity (or phase) of the reflection will depend on the type of termination used at the load end. Applying a positive pulse to an open line will cause a positive reflection with an amplitude depending only upon the line loss. Conversely, applying a positive pulse to a shorted line will cause the reflection of a negative pulse. Therefore, in reflected-pulse tests for line defects, the polarity of the return is indicative of the type of defect.

13-1062. An oscilloscope with a minimum sweep length of 0.05 microsecond, and provision for obtaining an 0.18-microsecond test pulse can be used to measure the elapsed time between the initial pulse and its reflection. An accurately calibrated time base is also necessary. The scope is internally synchronized, and the trigger output is applied to the vertical input jack of the scope as well as to the line under test. Attenuation encountered in testing extremely long lines will necessitate the use of an external trigger, such as a radar modulator trigger, to increase the amplitude of reflection. In free space or atmosphere, electrical energy will travel 483.7 feet and return in 1 microsecond. For example, assume that 5000 feet of RG-11/U cable is used with a 50-volt positive trigger fired into the line, which is opened at the receiving end. If the line is in good condition, a positive reflection 2.23 volts in amplitude will be seen 15.7 microseconds after the main pulse. A simplified formula for converting round trip time (in microseconds) to feet is: Distance in feet to reflection = μsec x velocity factor x 483.7. In the case of RG-11/U, since the pulse must actually travel 10,000 feet, the reflection will be 27 db down from the main pulse. Therefore, measuring the return-pulse amplitude will supply information relative to the attenuation of the line. Assume that a negative pulse was observed on the scope 9.6 microseconds after the occurrence of the trigger pulse. The fact that the return pulse is negative indicates that the line is shorted, and the point at which the short is located can be determined. Since the velocity factor for RG-11/U (from table 13-8) is 0.66, energy will travel along the cable 320 feet and return in 1 microsecond. Multiplying this by 9.6, the number of microseconds observed on the scope gives a total distance of approximately 3075 feet. Thus, the short is located approximately 3075 feet from the end of the cable to which the scope is connected. This check is rapid, accurate, and complete. The scope time base can be considered to represent the transmission line itself. Any poor connections or shorts will show on the time base, and the distance to the defect can be ascertained by one simple multiplication.

SECTION XIV

TRANSMISSION LINE PRESSURIZATION

13-1063. COAXIAL CABLE PURGING AND PRESSURIZING.

13-1064. GENERAL.

13-1065. A coaxial transmission line is most efficient when the interior of the line is dry. Moisture increases the attenuation and reduces the maximum operating voltage of the line. Hence, the coaxial cable must be purged of any moisture which may have entered, and then pressurized to prevent moisture from entering during operation. Purging should be accomplished immediately upon completion of cable installation. Purging should also be performed whenever a cable has been opened for repairs or when the resistance between the inner and outer conductors measures less than 100,000 megohms per 1000 feet of cable. After purging, pressurizing the coaxial cable transmission line to about 10 psi with dry air will keep the interior of the line dry.

13-1066. Dry air is air which has had sufficient moisture removed from it that the remaining moisture will not condense onto the inner walls of the cable over the cable's range of temperature and pressure variations. The temperature at which the water vapor in the air (in the cable) condenses is known as the dew point. Explained very briefly, as the temperature of the air in the cable is decreased, it is capable of holding a smaller quantity of water vapor. Finally, a temperature is reached where the air is saturated; that is, it cannot hold any more water vapor. If the temperature is lowered below this point, some of the water vapor condenses as water on the interior walls of the cable. If the dew point is below freezing, the vapor condenses into ice crystals.

13-1067. The air within a cable must be sufficiently dry to maintain its dew point below the minimum temperature to which the cable will be subjected. This value varies depending on the latitude, altitude, and time of year. Also, the temperature of the cable and of the air inside varies through out the day for a fixed graphical location. Since the volume of air inside the cable is constant, the pressure will vary with a change in temperature. Table 13-9 gives the pressure changes which are caused by a 100-degree Fahrenheit change in temperature, from 10 degrees to 110 degrees.

Table 13-9. Temperature-Pressure Variations

TEMPERATURE (°F)	PRESSURE (psi)
10	7.6
30	8.4
50	9.5
60	10.0
90	11.5
110	12.4

13-1068. Peak power ratings of coaxial cables can be increased by maintaining higher pressures in the dry air dielectric. Figure 13-287 shows the increase in peak power percentage obtained with an increase of pressure to approximately 3 atmospheres.

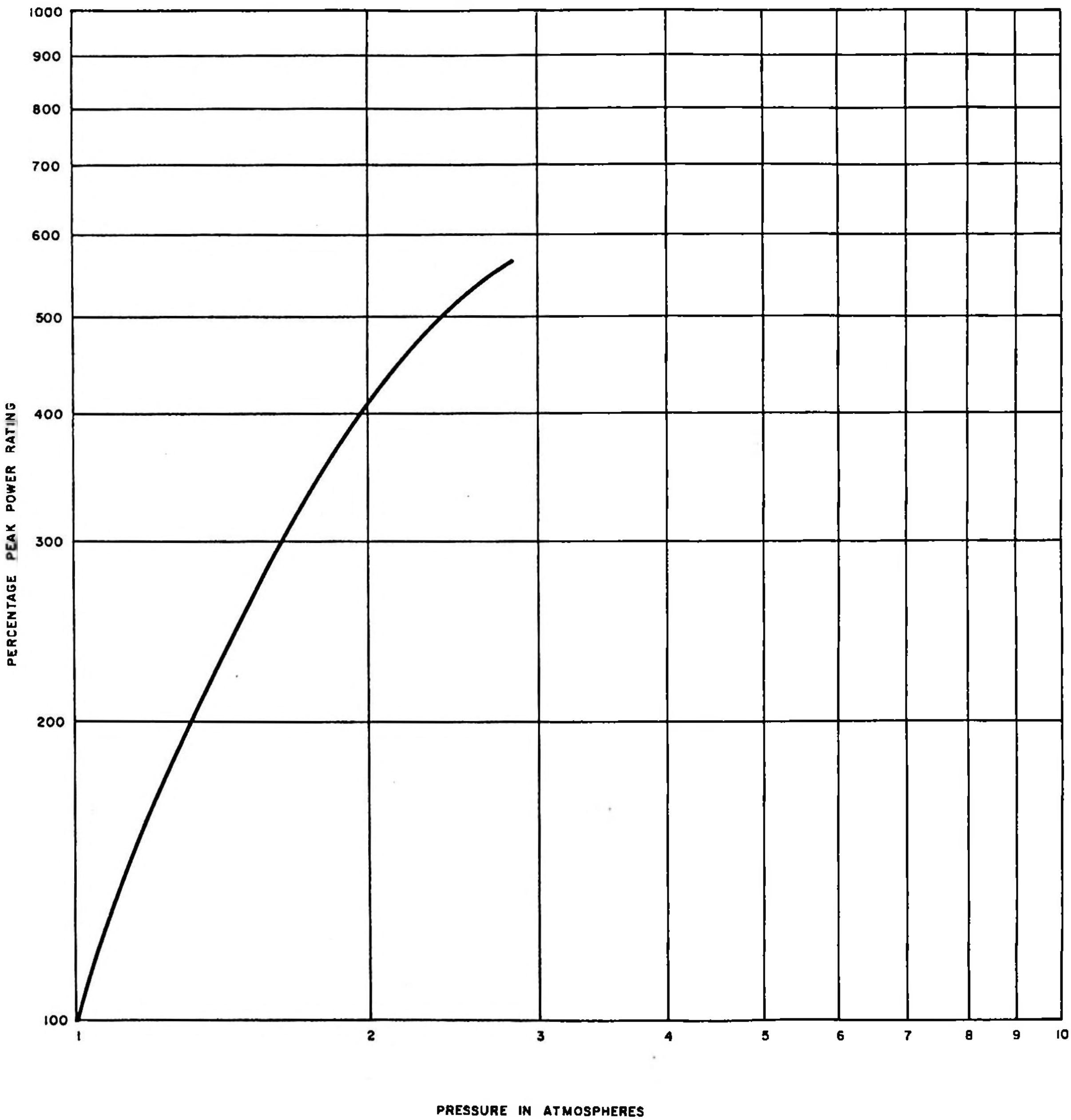


Figure 13-287. Dry-Air Power Rating vs Pressure

13-1069. GAS AND EQUIPMENT.

13-1070. Purging is usually accomplished using dry air or dry nitrogen gas. Nitrogen gas is generally preferable to dry air for small-diameter and medium-length cables. Standard nitrogen gas equipment is used. Dry air is obtained by using a manual dry air pump. The pump pushes the air through a desiccant, usually silica gel crystals, which extract any moisture. Silica gel crystals are normally colorless, and if used alone, their state of dryness (that is, their ability to extract moisture from the air) is usually very difficult to determine. To indicate the condition of the silica gel, a small sack of crystals is added which contain a chemical capable of changing color with moisture content. The darker their blue color, the dryer they are; and the darker the pink color, the more moisture they contain. Hence, if the crystals are pink, the silica gel must be dried by baking. A mechanical dehydrator, preferably one equipped with two desiccant chambers, is recommended for the continuous purging and pressurization of a large-volume cable. Dehydration is accomplished by forcing the air, using a compressor, through the desiccant chambers. The dry-air output from the dehydrator is connected by copper tubing to the cable through a pressure-regulating valve. Mechanical dehydrators are capable of maintaining pressures between 6 and 10 psi in 3/4 inch cable up to 50,000 feet long, and in 3-1/8 inch cable up to 10,000 feet long.

13-1071. PROCEDURE.

13-1072. GENERAL. The procedure for purging and pressurizing coaxial cables is as follows:

a. Connect the dehydration equipment to the cable. Suggested equipment arrangements are shown in figure 13-288, 13-289, and 13-290.

b. Pressurize the equipment to 10 psi and let it remain in this state for two hours.

c. Bleed the transmission line very slowly through the gas escape valve at the end of the line while maintaining pressure from the dry-gas source.

d. Continue the procedure in steps b and c until an insulation test indicates a satisfactory resistance.

e. Close the gas escape valve and pressurize the transmission line to 10 psi. A positive pressure should be maintained in the line.

13-1073. DETECTING LEAKS. When the pressure in the cable drops more than the normal amount caused by a change in temperature, efforts should be initiated to locate the leak. The first places you should examine are the cable connections and terminations, as these are the most likely areas for leaks to occur. A simple method is to apply undiluted liquid shampoo to the outside of the cable in the vicinity of the suspected area. If a leak exists, the escaping pressure will cause bubbles to appear. If nothing is detected, it may be necessary for you to segregate the cable into sections and carefully examine each section individually. First use all the divisions which already exist to determine the section containing the leak. When the leak is located and corrected, you should try to determine the cause so that it can be eliminated.

13-1074. WAVEGUIDE PRESSURIZATION AND SEALING.

13-1075. GENERAL.

13-1076. Microwave transmission lines are pressurized for protection against dirt, moisture, and, in tropical climates, mold growth and insects. These factors can

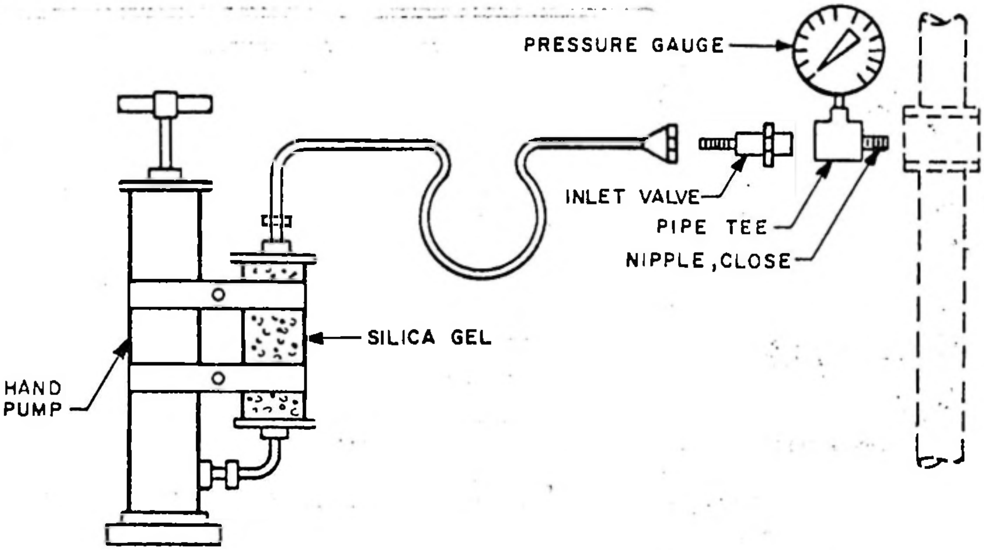


Figure 13-288. Dehydration Apparatus Using Manual Dry Air Pump

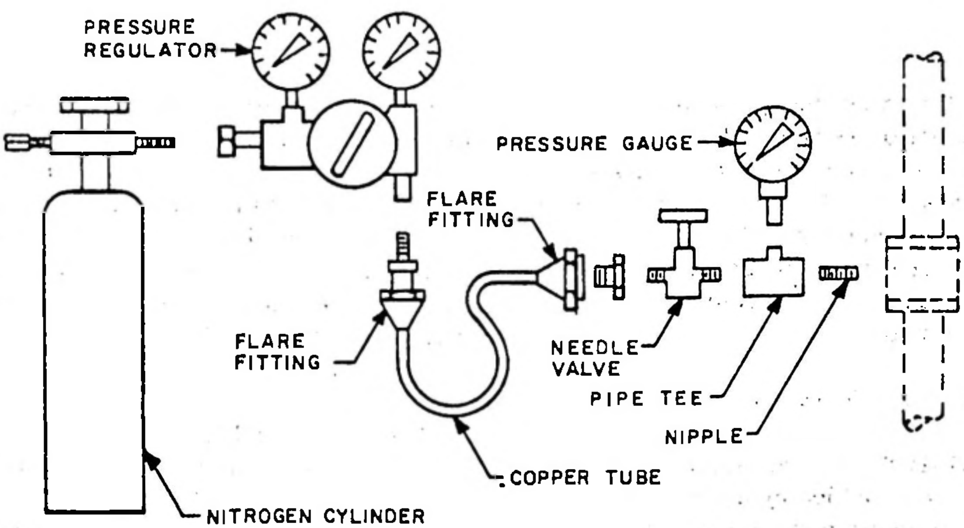


Figure 13-289. Dehydration Apparatus Using Nitrogen Gas

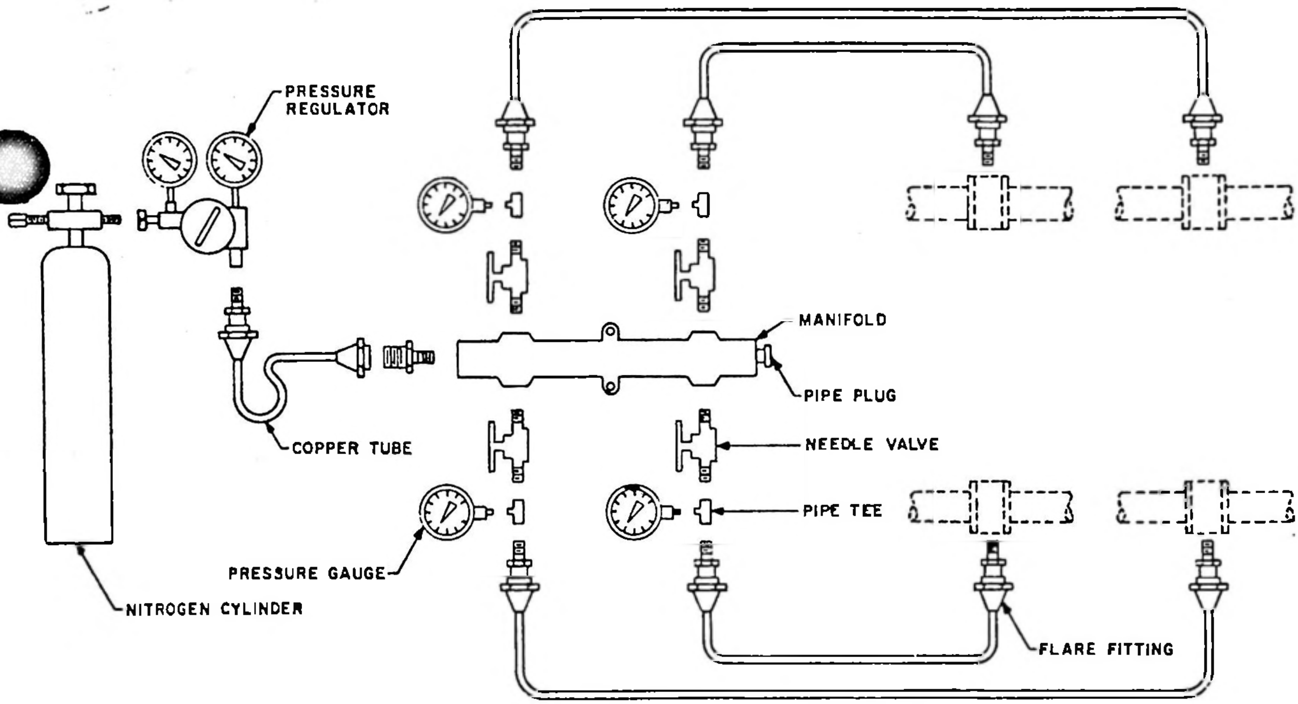


Figure 13-290. Dehydration Apparatus for Multiple Transmission Lines

seriously deteriorate the operation of waveguide equipment by reducing its power-handling capacity. Pressurization of a transmission line is also used to avoid voltage breakdown or corona discharge. In all cases, it is essential that the line be adequately sealed so that the correct working internal pressure can be maintained.

13-1077. TYPES OF SEALS.

13-1078. WINDOWS. The simplest seal is a viewing plane provided by a nonresonant thin sheet of dielectric material. Mica, teflon film, and polythene are some materials commonly used for this purpose. To completely seal a microwave assembly involves capping the antenna feed in some manner with a dielectric cover or window. This cover then becomes an integral part of the antenna feed and is involved in any measurement of its electrical characteristics. In addition, all branch lines must be sealed, preferably with some form of window. Most magnetrons and the newer klystron local oscillators can also be pressure-sealed to a waveguide assembly. Of the windows available for pressurization, most are rated for a differential pressure of 30 to 50 psi or more.

13-1079. GASKETS. When a pressurized waveguide assembly is inside a sealed compartment, the connecting flanges may be of the unpressurized type. In most cases, however, the line will be exposed and the connecting flanges are provided with grooves for sealing gaskets. Such gaskets must be made of a material which does not acquire a permanent set when compressed for long periods of time. The material should also resist deterioration from weather, oils and greases, and any elements peculiar to the area in which the gasket is to be used, such as mold growth in tropical climates. These gaskets are of two distinct types.

13-1080. One type of gasket is composed of a material containing a cross section which can be distorted but not compressed. Synthetic materials such as neoprene are generally used, but for operation down to -40 degrees C natural rubbers are preferable. The rectangular groove in the connector is designed so that when the gasket is pressed into place it extends beyond the groove by about 15 percent of its diameter. When the connector is bolted to the wave guide flange, the gasket expands to form a seal.

13-1081. The second type of gasket has a flat, rectangular cross section and is made of compressible material. The gasket has a thickness of about 50 percent greater than the groove depth of the flange and on clamping is compressed to form a seal. A cork-neoprene composition is the material generally used in this type of gasket. If no contact shim is used in the connector, the gaskets in the two mating flanges press against each other to form a satisfactory seal. Gaskets of this type require considerably more clamping force than the round type, and the flanges must be carefully mated to avoid distortion.

13-1082. USE OF GASES.

13-1083. When a microwave assembly is operated at high power levels, the air inside the waveguide ionizes and causes arcing and corona. The value of power at which this arcing and corona appear is referred to as the power-breakdown level. Pressurizing the air within the guide will often raise the power-breakdown level to a point which enables the guide to handle the power required by the equipment in use. However, in some instances, pressurization of the air alone may not be sufficient to overcome a power-level breakdown problem. More effective results can be obtained by using a gaseous dielectric in the waveguide rather than air.

13-1084. Sulfur hexafluoride (SF_6) is gas commonly used in conjunction with pressurization in waveguides. The properties of this gas are such that it can be used to significantly raise the power-handling capability of a transmission line without physically changing the line itself. Where the prevention of moisture and corrosion are the primary considerations for pressurizing a waveguide assembly, dry nitrogen gas is often used in place of air. The excellent dehydrating ability of nitrogen, along with ease of procurement, makes it extremely useful for this purpose.

13-1085. You may find that there are numerous other gases associated with waveguide pressurization, but sulfur hexafluoride and nitrogen are the most commonly used. Factors such as economy, safety, availability, etc., limit the use of other gases to special applications, and it is unlikely that you will encounter any of them.

13-1086. RECTANGULAR WAVEGUIDE DISTORTION.

13-1087. The internal pressure of a waveguide may be raised as high as 50 psi or more in order to increase its power-handling capacity. Rectangular waveguide, however, has a poor shape to withstand such a differential pressure, as the walls of the wide dimension have a tendency to bulge or bow outward. An exaggerated example of this is illustrated in figure 13-291. This outward deflection produces a small change in the wave impedance of the guide, but causes little trouble since it is reduced considerably in the neighborhood of the connecting flanges, where its effects would be mostly felt. This distortion does, however, produce an inward deflection of the narrow walls. This causes a change in the guide wavelength since the

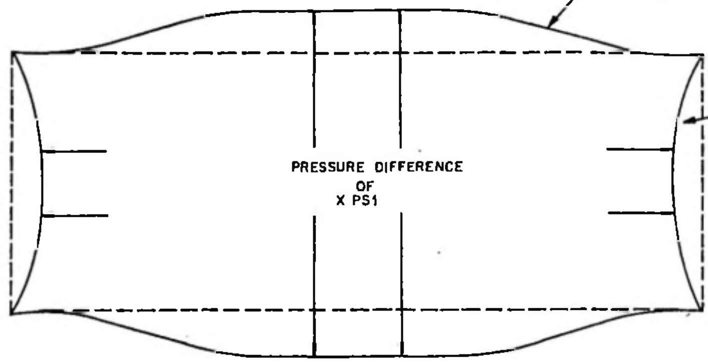


Figure 13-291. Distortion of Waveguide Having Excess Internal Pressure (Exaggerated)

wavelength is determined by the distance between the narrow walls. This will lead to trouble when the phase is critical over a given line length.

13-1088. CIRCULAR WAVEGUIDE DISTORTION.

13-1089. Due entirely to its physical configuration, a circular waveguide is much less susceptible to distortion than the rectangular type. The distortion in a circular guide is so small that it is considered negligible. However, the same physical factor which makes circular waveguide readily adaptable to pressurization also makes it very difficult to maintain proper mode orientation. Therefore, this type of guide is generally used only for specialized applications, such as in a rotating antenna, where the use of circular waveguide is advantageous.

13-1090. DISTORTION MEASUREMENT.

13-1091. An apparatus used to accurately measure the distortion of a waveguide consists of a strain gage-activated transducer, an amplifier, a deflectometer, and a recording system. The transducer itself, shown in figure 13-292, is essentially a cantilever beam which is deflected as the waveguide section passes beneath the probe point. These deflections are picked up as strain gage outputs and fed to an amplifying arrangement which controls a recorder chart. A deflectometer is used to synchronize the chart drive and the trace of the transducer probe along the waveguide wall. The result is a magnified profile of the waveguide wall on the chart paper.

13-1092. METHODS OF ELIMINATING DISTORTION.

13-1093. Several methods can be used to reduce or eliminate distortion in pressurized waveguide. The easiest and most effective method is to keep the internal pressure of the guide low enough to prevent the occurrence of distortion. Typical limiting values of pressure for a given size waveguide are:

5 psi for 5.50 in. x
3.25 in. guide

15 psi for 2.85 in. x
1.34 in. guide

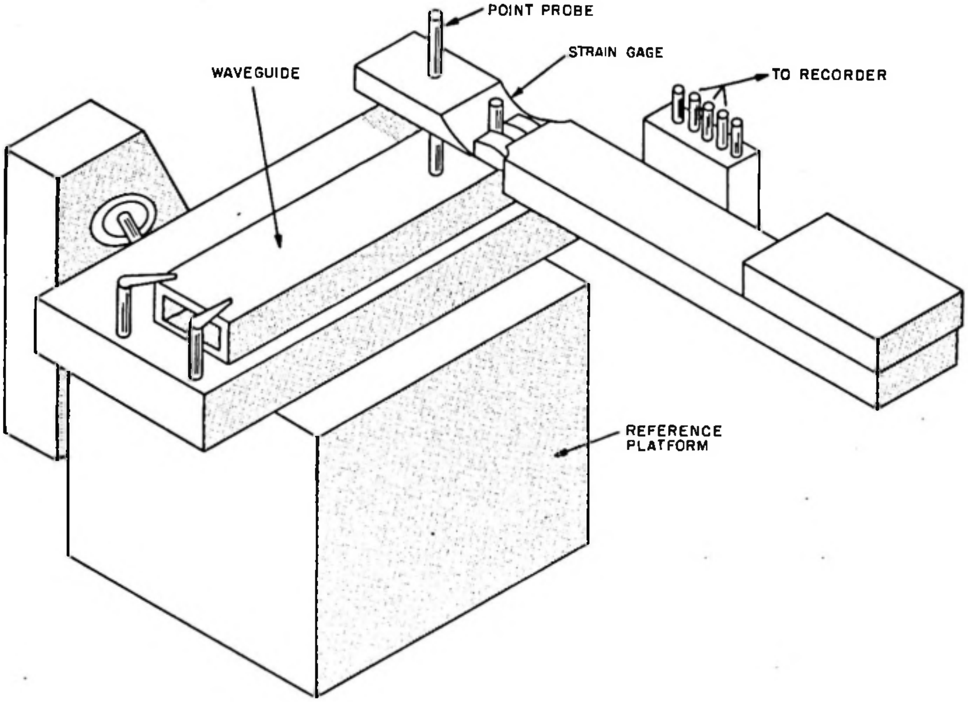


Figure 13-292. Strain Gage-Activated Transducer

30 psi for 0.90 in. x
0.40 in. guide

Although this is the most effective method, it is not always the most practical since pressure exceeding the limiting value is often necessary for satisfactory operation of the waveguide.

13-1094. When excess pressure is required for satisfactory operation, waveguide distortion can be minimized by artifically increasing the strength of the guide with external clamps placed at intervals along its length. Such clamps are generally made of steel and should be closely spaced if distortion is to be kept within acceptable limits. Also, if these clamps are to be used, you must exercise great care upon installation. Carelessness can result in a nicked, dented, or crushed waveguide.

13-1095. Another method of minimizing distortion is to enclose the rectangular waveguide in a circular sheath or cylinder and to pressurize this sheath and the waveguide to the same pressure. The pressure difference is then transferred to the sheath, which has a shape less susceptible to distortion. If any distortion of the sheath should occur, it will have no electrical effect on the waveguide assembly. Other advantages of this method are that it eliminates the need for clamps and provides excellent physical protection for the guide against accidental damage, such as dents and punctures. However, this method is at a distinct disadvantage in areas where space limitations are severe, since the diameter of the sheath must necessarily be about twice the waveguide's narrow dimension.

13-1096. PRESSURIZATION EQUIPMENT.

13-1097. The equipment used to pressurize waveguide sections is basically the same as that used for coaxial cable. However, a manual dry air pump is seldom used to pressurize waveguide assemblies of any appreciable length because of the large volume of air required. When sulfur hexafluoride is used as the dielectric, a charcoal filtering device is generally included as part of the equipment for purification of the gas.