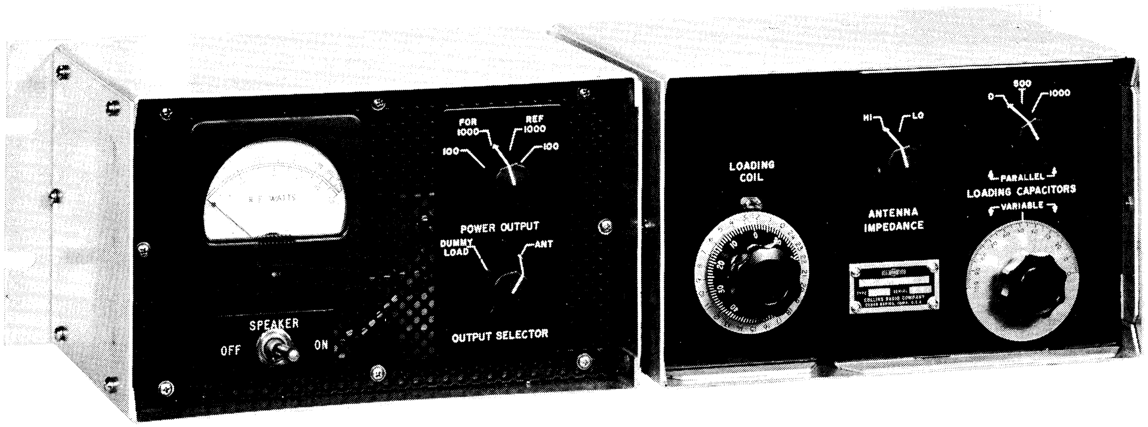
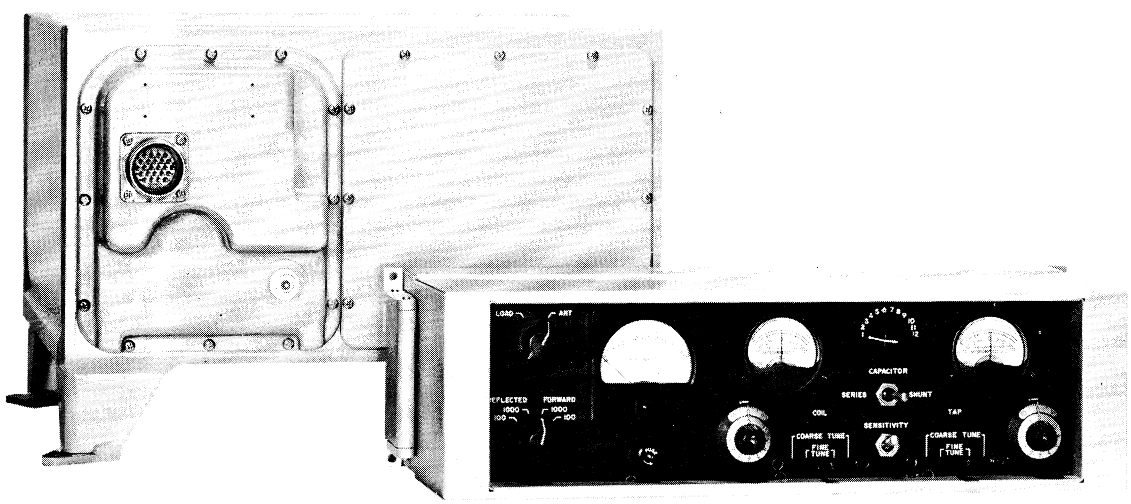


Collins 180U-2 Fixed Station Antenna Network Matches a 50-Ohm Output to a 50-Omh Transmission Line having a VSWR Up to 2 to 1. Power Handling Capacity is 1000 Watts over the Frequency Range of 2 - 30 Mc



Collins Designed the AN/SRA-22 Antenna Coupler to Satisfy the Tuning Requirements of Whip and Other Antennas Normally Encountered in Shipboard Applications. The Weathertite Enclosure Mounts at the Antenna Base with a Remote Control Unit for Convenient Operation

CHAPTER 10
ANTENNA FEED SYSTEMS

1. INTRODUCTION

The material presented in this chapter will be that concerning the network between the source and its load. As a matter of convenience, the transmitter will be considered the source, and the antenna the load; in most cases, however, the same principles apply to the receiving system. The primary subject is that of transmission lines, their physical and electrical properties, and their applications in high-frequency antenna feed systems. As with the antenna, the type of emission employed does not generally impose any particular requirements on the selection of transmission line.

Although the major part of the text is devoted to the practical aspects of feed systems, some basic transmission line theory is included as a review of the subject. Included at the end of the chapter is also a brief discussion pertaining to antenna and feedline measurements.

2. TYPES OF R-F LINES

Electrically there are two basic types of uniform transmission lines for r-f operation, balanced and unbalanced. When a line operates with potentials of equal magnitude and opposite polarity from each side of the line to ground, it is balanced; when a line operates with one side at ground potential and the other side at r-f potential, it is unbalanced. Mechanically there are also two basic forms, open wire and enclosed. Open wire lines are constructed of two or more conductors, uniformly spaced and supported in air above ground; enclosed lines utilize one or more transmission circuit conductors enclosed by a metallic sheath. Lines of the enclosed type may be either balanced or unbalanced.

3. TRANSMISSION LINE PROPERTIES

The electrical characteristics of a transmission line, which consist of inductance, capacitance, series resistance, and shunt conductance per unit length of line, are determined by its cross-sectional geometry and the nature of the dielectric medium. For most practical purposes, the series resistance and shunt conductance may be neglected, and the typical r-f transmission line may be represented by the equivalent circuit of figure 10-1.

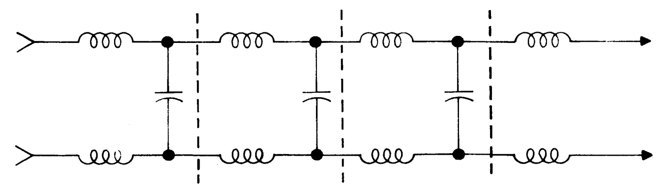


Figure 10-1. Approximate Representation of a Two-Wire Transmission Line By Means of Lumped Constants

The inductance, L, and capacitance, C, per unit length of line determine its characteristic or surge impedance, Z_0, which is equal to $\sqrt{\frac{L}{C}}$. The physical significance of the characteristic impedance is simply that impedance which the line offers to the flow of current (provided the line is terminated in its characteristic impedance). The characteristic impedance of typical r-f lines ranges from 50 ohms in the coaxial type to over 600 ohms in the open wire type.

When a transmission line is not terminated with a load equal to its characteristic impedance, standing waves of voltage and current will exist along the line as a result of reflection from the load end of the line. Their magnitude depends on the amount of mismatch between the characteristic impedance of the transmission line and the impedance of the termination. The standing-wave ratio, swr, is a measure of the degree of mismatch and is defined as the ratio of maximum rms voltage (or current) to the minimum rms voltage (or current) of the resultant standing waves. A line terminated in its characteristic impedance will exhibit no standing waves, and the swr will be 1:1. When a lossless (hypothetical) line is terminated in a complete short circuit or a complete open circuit, the resultant swr is infinite. Under actual conditions where lines are somewhat lossy and are affected by proximity, the swr is much less than infinite. When the load is purely resistive,

$$swr = \rho = \frac{Z_r}{Z_o} \text{ or } \frac{Z_o}{Z_r}$$

where Z_r = impedance of load

Z_o = characteristic impedance of the line

ρ = standing-wave ratio (by definition, swr $\geq$ 1.0)

Swr refers to voltage or current unless power is specifically mentioned. The voltage standing-wave ratio (vswr) must be squared to convert to power relationships. The power delivered to the load is equal to the difference between the energy contained in the incident wave and the energy contained in the reflected wave, which is determined by the degree of mismatch at the load end of the line. When both incident and reflected power are known, swr can be determined from

$$\rho = \frac{\sqrt{\frac{P_i}{P_r}} + 1}{\sqrt{\frac{P_i}{P_r}} - 1}$$

where P_i = incident power

P_r = reflected power

The power contained in the reflected wave, however, does not represent actual loss except as it is attenuated in traveling back to the input end of the line. Neglecting effects on the source, it is merely power which is not being coupled, resulting in inefficient operation. For most h-f antenna work, a vswr of 1.5:1 or under represents a desirable impedance match. Where an antenna is required to operate over a wide-frequency range, a vswr of 2:1 or under is usually satisfactory.

Frequently the term reflection coefficient, k, is used to designate terminal mismatch on a line and is related to the swr by

$$k = \frac{swr - 1}{swr + 1} = \frac{\rho - 1}{\rho + 1}$$

The power lost in a line is least when the line is terminated in a resistance equal to its characteristic impedance, and increases with an increase in standing-wave ratio, because the effective values of both current and voltage become greater. The increase in effective current raises the ohmic losses in the conductor, and the increase in effective voltage increases the losses in the dielectric. The effect of swr on line loss is shown in figure 10-2.

In addition, swr affects the power handling capability of a given transmission line, as limited by the

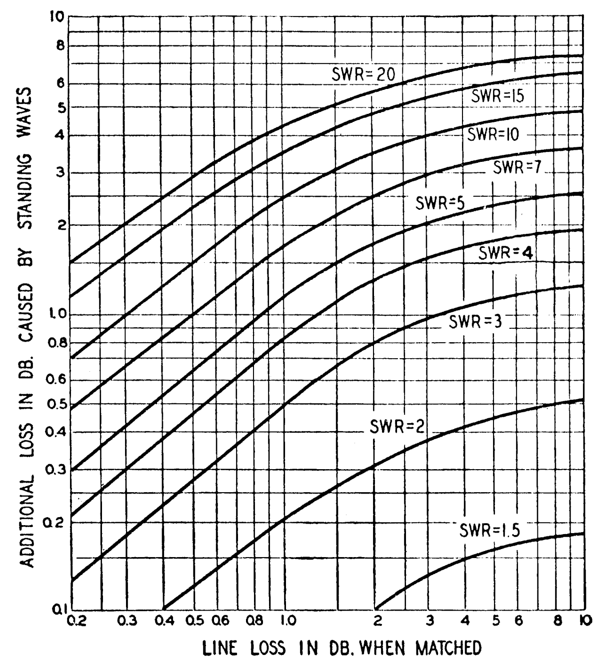


Figure 10-2. Effect of SWR on Line Loss

insulation breakdown voltage. This is of particular importance in transmission lines of the coaxial type. Most manufacturers derate their lines in direct proportion to the swr, so, in order to operate the line efficiently, it is necessary to maintain a low swr.

The velocity factor is another property of typical transmission lines. The velocity factor is the ratio of the actual propagation velocity along the line, to the propagation velocity in free space and accounts for the electrical length of a given line being somewhat longer than the physical length. Values of the velocity factor will differ for different line types and are available from manufacturers' specifications. The physical length corresponding to an electrical wave length is given by

$$L(\text{feet}) = \frac{984}{f} v$$

where f = frequency in mc

v = velocity factor

4. TYPICAL R-F TRANSMISSION LINES

a. TWO-WIRE BALANCED LINES--AIR DIELECTRIC

The two-wire balanced, air dielectric line is probably the most common type of feeder for balanced operation in the h-f range. When properly constructed and operated, the two-wire balanced line is capable of handling high power with comparatively low loss. If the lines are not properly balanced with respect to ground or are too widely spaced, they are subject to radiation. Although the two-wire balanced line is most frequently employed where a characteristic impedance in the range of 400 to 600 ohms is required, it can also be designed for a lower characteristic impedance. Transmission lines with surge impedances in the order of 200 ohms and up can be designed by employing larger diameter conductors (tubing). Minimum spacing of lines is limited by voltage breakdown considerations, mechanical difficulties, and by the increases in losses due to proximity effects. When the effect of insulating spacers is neglected, and the lines are high enough above ground ($h \gg b$), the characteristic or surge impedance of the two wire balanced line may be expressed as shown in figure 10-3.

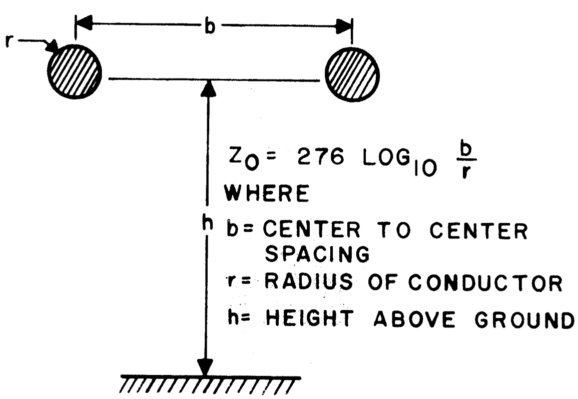


Figure 10-3. Cross Section of Two-Wire Balanced Line

Because of the effect of supporting structures, dielectric spreaders, line losses, and other factors, the velocity factor of open wire line is slightly less than unity, ranging from 0.95 to 0.98 for typical lines. Conductor size and spacing for a given characteristic impedance may be determined from the curves of figure 10-4.

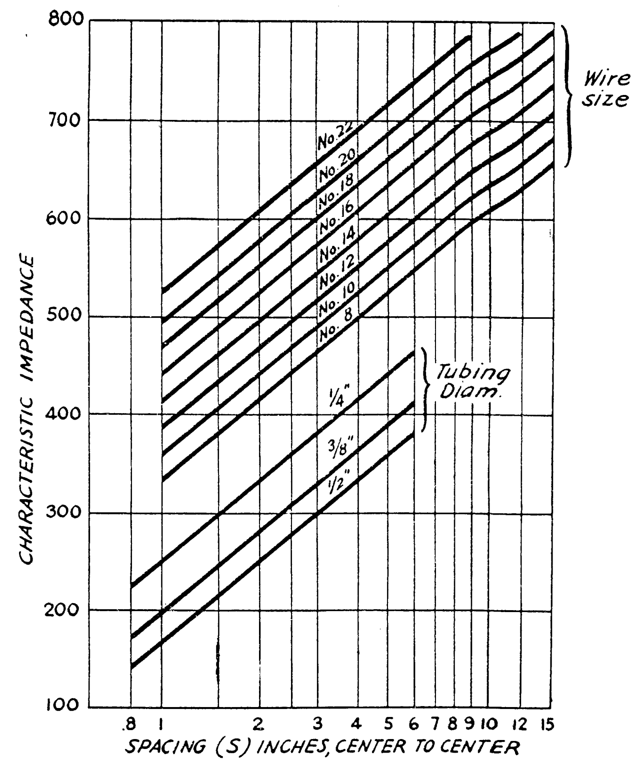


Figure 10-4. Characteristic Impedances for Typical Values of Conductor Size and Spacing (Two-Wire Balanced Lines)

b. FOUR-WIRE BALANCED LINES--AIR DIELECTRIC

Four-wire balanced transmission lines are commonly used where a lower characteristic impedance than that of the two-wire balanced line is required. The four-wire line can be side-connected as in figure 10-5a or cross-connected as in figure 10-5b. Both types of line exhibit a lower characteristic impedance than that of the two-wire line, but the cross-connected line shown in figure 10-5b is less subject to radiation or pickup. A popular application of the four-wire balanced line is for feeders to receiving systems in which several rhombic antennas are employed for diversity reception. The characteristic impedance required here is about 200 ohms. The characteristic impedance of the feeder shown in figure 10-5b may be expressed by

$$Z_o = 138 \log_{10} \frac{ab}{r\sqrt{a^2 + b^2}}$$

From figure 10-6, the characteristic impedance may be determined for various wire sizes and spacings for the special case where a = b and D is the diagonal dimension represented above as $\sqrt{a^2 + b^2}$.

c. TWO-WIRE BALANCED LINES--SOLID DIELECTRIC

Two-wire balanced solid dielectric lines are produced commercially in forms commonly known as molded pair, ribbon, or twin-lead. These lines are more lossy, but are flexible and fairly rugged. Because of the solid dielectric, lower characteristic impedances can be achieved than are practical with open wire, air dielectric lines. Typical size lines are 75, 100, 150, 200, and 300 ohms. The velocity factor of this type of line ranges from about 0.68 to 0.82, depending upon its construction.

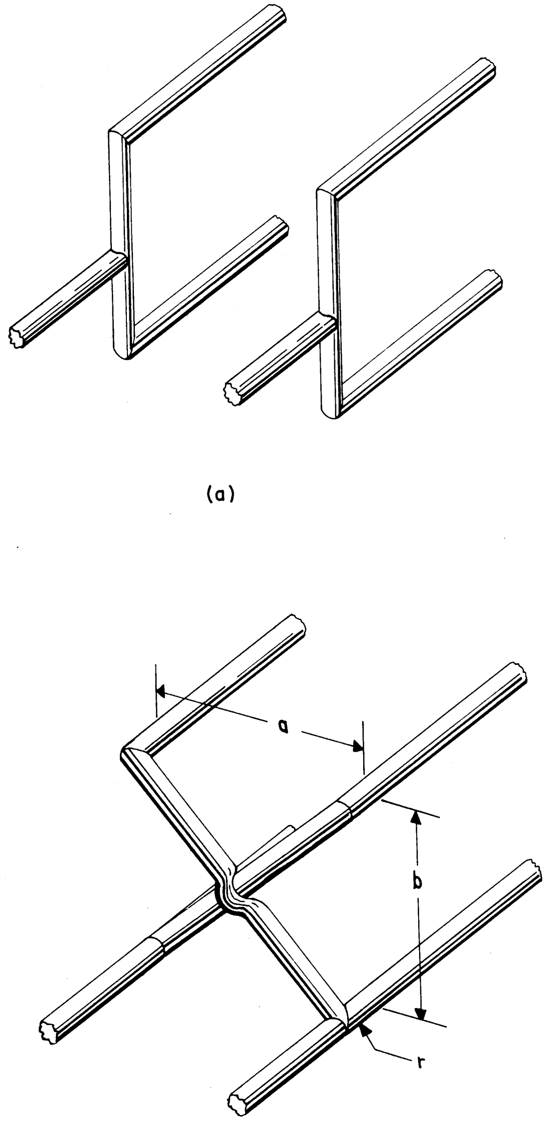


Figure 10-5. Cross Section of Four-Wire Balanced Lines

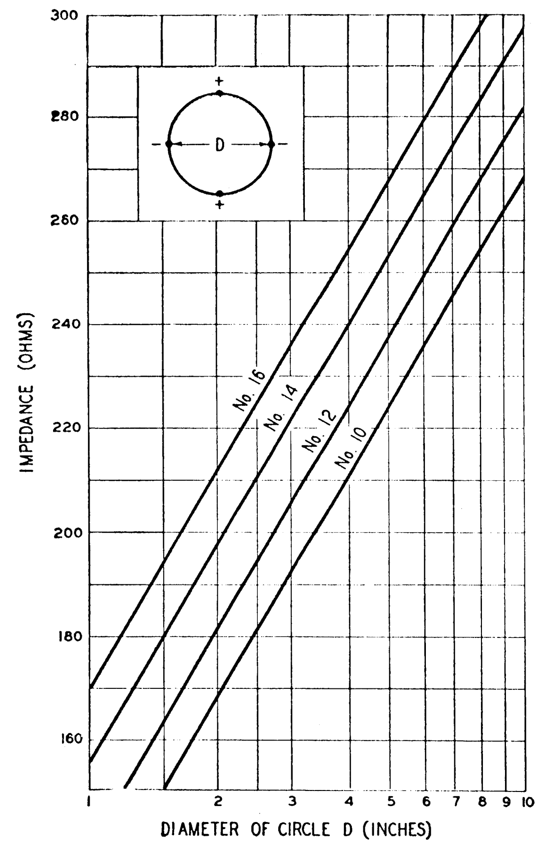


Figure 10-6. Characteristic Impedances of Four-Wire Balanced Line with Typical Values of Conductor Size and Spacing

d. AIR DIELECTRIC COAXIAL LINES

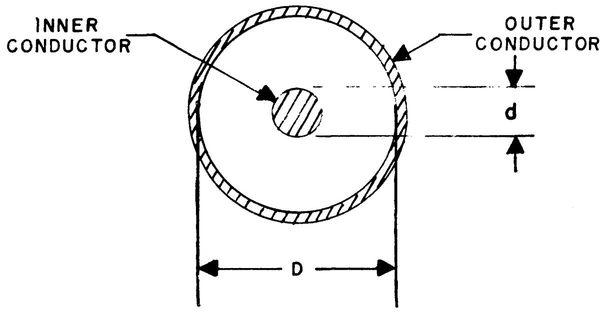
With coaxial construction, it is possible to achieve a much lower characteristic impedance than is practical with the parallel-conductor type line. For air as the dielectric, the 50-and 70-ohm lines are the most popular. The 50-ohm coaxial line, in fact, is fast becoming a standard transmission line in modern communications systems. Since the fields are contained within the outer conductor, loss by radiation is eliminated and, by the same virtue, the line is free from pickup.

Air dielectric coaxial line is available in the rigid or the semiflexible type and in various diameters depending on the power handling requirements; some common sizes are 7/8 inch, 1-5/8 inch, 3-1/8 inches, and 6-1/8 inches (outer diameter). To prevent moisture from forming inside the line, pressurization with dry-air or nitrogen is usually required. The velocity factor ranges from 0.83 to 0.99.

The characteristic impedance of air dielectric coaxial line may be determined with sufficient accuracy from the expression shown in figure 10-7.

e. SOLID DIELECTRIC COAXIAL LINES

Solid dielectric coaxial lines differ from air dielectric types in that the inner and outer conductors are separated by a dielectric material, such as Polyethylene or Teflon. Such cables have the advantage of easy handling and installing, but for a given size, the attenuation is higher and the power-handling



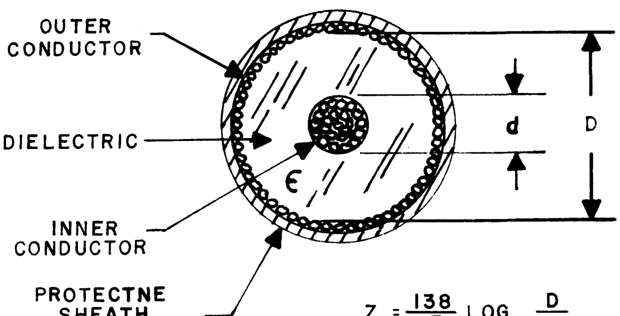
NOTE INNER CONDUCTOR IS SUPPORTED BY DIELECTRIC BEADS OR SPIRAL SPACERS

$$Z_0 = 138 \; \text{LOG}_{10} \frac{D}{d}$$

WHERE D= INSIDE DIAMETER OF OUTER CONDUCTOR
d= OUTSIDE DIAMETER OF INNER CONDUCTOR

Figure 10-7. Cross Section of Air Dielectric Coaxial Line

capabilities are lower than that of air dielectric types. The velocity factor for lines using Polyethylene is 0.659 and for lines using Teflon is 0.695. The characteristic impedance of solid dielectric coaxial lines may be determined from the expression of figure 10-8.



$$Z_0 = \frac{138}{\sqrt{\epsilon}} \; \text{LOG}_{10} \frac{D}{d}$$

WHERE ε= DIELECTRIC CONSTANT
D= INSIDE DIAMETER OF OUTER CONDUCTOR
d= OUTSIDE DIAMETER OF INNER CONDUCTOR

Figure 10-8. Cross Section of Solid Dielectric Coaxial Line

f. TWO CONDUCTOR SHIELDED CABLES

Two conductor shielded cables are commercially available for specialized applications requiring a shielded balanced line. However, this type of line is not commonly used in high-frequency antenna feed systems.

g. SINGLE WIRE LINE

A single wire line, well above ground and surrounding objects, exhibits a characteristic impedance of about 500 ohms when operated with ground as the return circuit. Although this type of feed is subject to radiation and is usually avoided, it provides a quick and simple method of improvising a temporary feedline.

5. TRANSMISSION LINE SECTIONS AS CIRCUIT ELEMENTS

The characteristics of low-loss resonant sections of transmission lines make them desirable for use as high-Q circuit elements. Their usefulness as circuit elements in the high-frequency range can be reduced to three basic applications: (a) parallel resonant circuits, (b) series-resonant circuits, and (c) low-loss reactances. The reactance of circuit elements comprised of transmission line sections varies differently with frequency than does the reactance of the ordinary inductance and capacitance type circuit. This is due to the linear configuration of the transmission line section as compared to the lumped form of the ordinary inductance and capacitance. However, the distributed inductance and capacitance of the line section do not determine directly the reactive effects of the linear circuit elements. The reactance is a result of the reflection effects which depend primarily on the relationship between L, C, line length, and line termination.

a. LINE SECTIONS AS PARALLEL RESONANT CIRCUITS

When a section of transmission line is terminated in an open or short circuit, high magnitude standing waves of voltage and current exist on the line. At current standing-wave nodes, the impedance appears as a pure resistance of very high value (infinite in the case of the hypothetical lossless line) and resembles the impedance of a conventional high-Q parallel resonant circuit consisting of lumped inductance and capacitance. At points substantially removed from the current node (exact resonance), the impedance characteristics of the line section no longer resemble those of the lumped constant "tank" circuit. This is due to the effect of the linear circuit.

The use of transmission line sections as parallel resonant circuits is more popular in the vhf and uhf regions than in the h-f range, because the lines become excessively long at the lower frequencies. The quarter-wave section with its end terminated in a short circuit, as shown in figure 10-9, is the most commonly used line section for this type of circuit element, but, as can be seen from figure 10-10, other varieties of line sections will produce the same circuit. The line sections can be of the balanced or unbalanced, open wire, or coaxial type. The length of line designated for the particular circuit element is the effective electrical length, and where applicable, the velocity factor of the line must be considered.

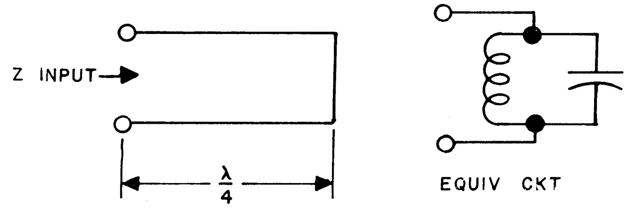


Figure 10-9. Lumped Circuit Analogy for Quarter-Wave Section of Transmission Line Terminated in a Short Circuit

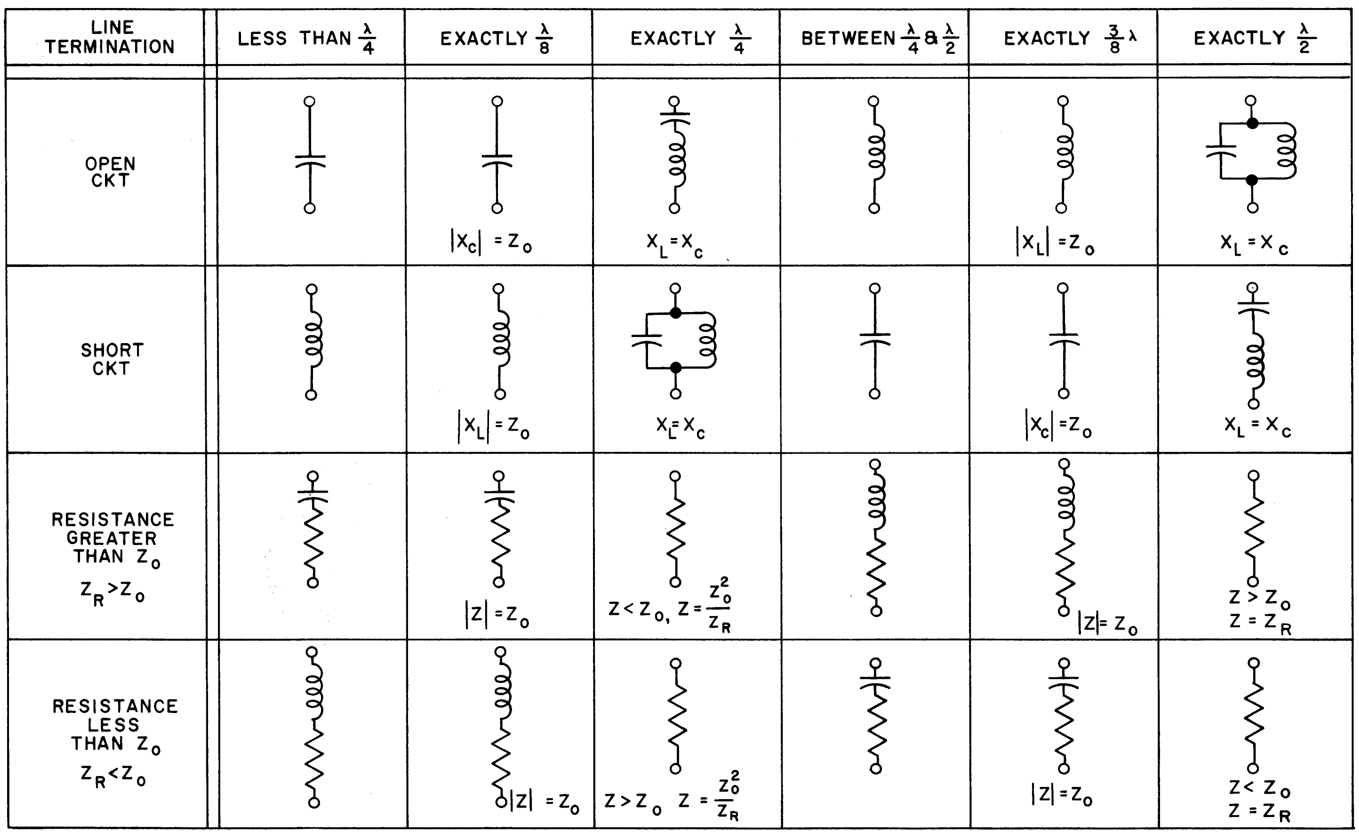


Figure 10-10. Lumped Circuit Analogy of Transmission Line Sections of Various Lengths

For high-frequency applications where the line section length required for resonance becomes excessive, a short-circuited quarter-wave parallel resonant line can be made physically much shorter by connecting a capacitor across the open end of the line section. The degree to which a resonant quarter-wave line section is electrically lengthened or physically shortened by addition of a given lumped capacitance depends upon the characteristic impedance of the line. The higher the characteristic impedance of the line section, the greater the effect of a given lumped capacitance. For resonance, the reactance of the capacitor must equal the inductive reactance presented by the shortened line section. As shown in figure 10-10, the effective impedance at the open end of a shorted line section less than a quarter-wave in length is inductive, and its value depends upon both its characteristic impedance and length. A practical high-frequency application for the shortened parallel resonant line section is the balanced-to-unbalanced transformer.

b. LINE SECTIONS AS SERIES RESONANT CIRCUITS

At voltage standing-wave nodes on a transmission line terminated in an open or short circuit, the impedance appears as a pure resistance having a very low value (zero in the case of the hypothetical lossless line) and resembles the impedance of a conventional high-Q series-resonant circuit consisting of lumped inductance and capacitance. At points on the line substantially removed from the voltage node (exact resonance), the impedance characteristics of the line section no longer resemble those of the lumped series-resonant circuit. Figure 10-10 illustrates that for series resonance, an open-circuited line may be any odd number of quarter-wave lengths long, while a short-circuited line may be any even number of quarter-wave lengths long. Although series-resonant linear circuits are not as frequently employed in antenna systems as are the parallel resonant line sections, they are quite useful for such purposes as the suppression of harmonics and as compensating reactances. When applicable, the quarter-wave line section shown in figure 10-11 is the most common.

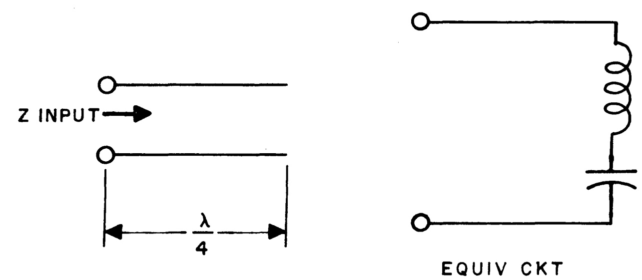


Figure 10-11. Quarter-Wave Open Transmission Line Section and its Equivalent Circuit

c. LINE SECTIONS AS LOW-LOSS REACTANCES

The reactance produced by linear circuit elements can have almost any magnitude and can be either capacitive or inductive, depending upon the electrical length of the line section and upon the type of termination (open or short-circuited). This is illustrated in figure 10-10 for the lossless line; losses present in actual line sections result in a small resistive component. In a well constructed line section, however, the losses are so low that this resistive component usually can be neglected. The characteristics shown in figure 10-10 are repeated for any multiple of an electrical half wave length that is added.

(1) LINE MATCHING STUBS

The most common use of line sections as low-loss reactive elements is for impedance matching on antenna feedlines. In this application, they are ordinarily referred to as stubs and are employed to cancel or eliminate undesirable reactive components of impedance (figure 10-12). Although only current standing waves are shown in the figure, standing waves of both current and voltage exist, due to the mismatch at the line termination. For every wave length on the line there are four recurring points where the resistive component is equal to the characteristic impedance of the line. If a stub is placed across the line at one of these points and is adjusted so that its reactance is equal in magnitude but opposite in sign to that of the reactive component existing at that point, the line will become "flat" or matched from that point back to the generator. As shown in figure 10-12, the standing waves have been removed only between the stub and the generator. Therefore, to operate a mismatched line with the highest efficiency, the stub should be placed as close to the load as practicable. In addition, this placement will allow operation of the over-all antenna system over a broader frequency range.

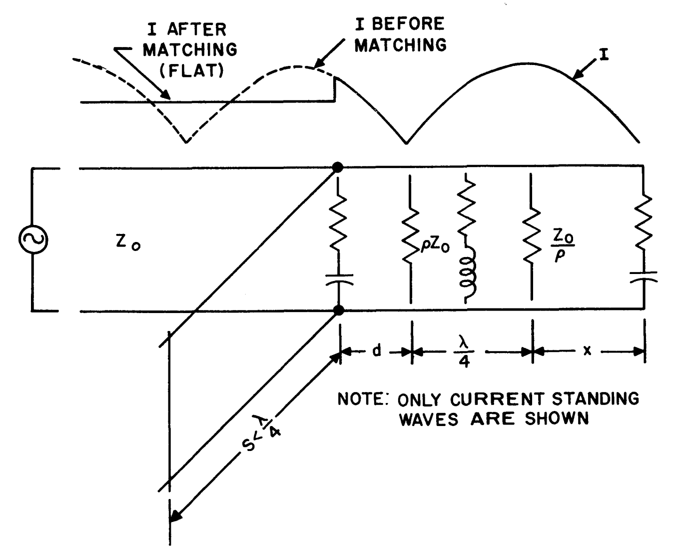


Figure 10-12. Effect of Shorted Stub on SWR Between Stub and Generator

The reactance of short-circuited or open-circuited line sections less than one quarter-wave in length may be determined from the universal reactance curves of figure 10-13, or from the equations illustrated in figure 10-14.

(2) DOUBLE STUB MATCHING

It often is impractical to locate a single stub at the desired point along the line; this is particularly true with coaxial lines when experimental final adjustments make it necessary to vary the point of attach-

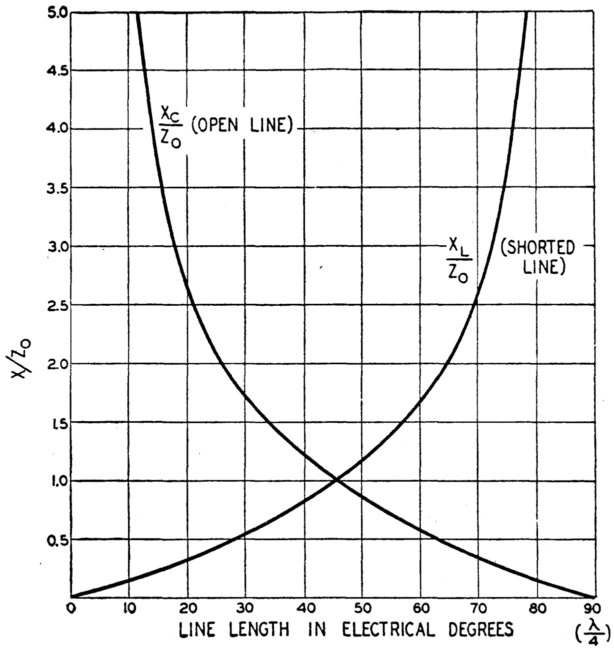
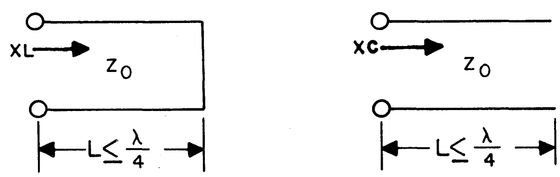


Figure 10-13. Universal Reactance Curves for Open or Shorted Lines Less than a Quarter-Wave in Length



X_L (OHMS) $= Z_O \tan \frac{2\pi}{\lambda} L$

X_C (OHMS) $= Z_O \cot \frac{2\pi}{\lambda} L$

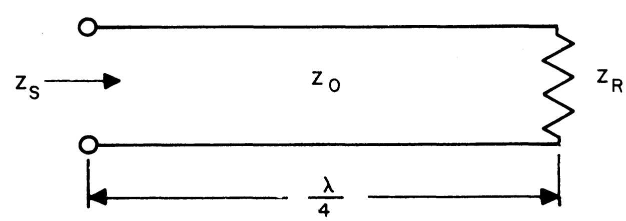
Figure 10-14. Line Sections as Reactive Elements

ment. This difficulty frequently can be overcome by employing two stubs of adjustable length with a predetermined spacing. Common separation of the two stubs is 1/8 wave length or odd multiples of 1/8 wave length. The range of complex impedances which the double stub arrangement is capable of matching is limited, but is usually adequate for the practical antenna installation. The two stubs, together with the line, act as a combined impedance transformer and reactance-canceling circuit. The resulting effect is essentially the same as changing the location of the single stub.

d. LINE SECTIONS AS IMPEDANCE TRANSFORMERS

The properties of a transmission line that enable it to perform as an impedance transformer are illustrated in figures 10-10 and 10-12, where the resultant or apparent impedance is shown to be a function of characteristic impedance and length of the line section.

An important application of this characteristic is the popular quarter-wave transformer. Referring to figure 10-12, it can be seen that at given distance, x, from the load end of the line, the apparent impedance is a resistance equal to $\frac{Z_o}{\rho}$ where ρ is the standing-wave ratio. Then, a quarter-wave farther down the line, the apparent impedance is another resistance, but now equal to ρZ_o, or multiplied by an amount equal to the square of the swr. Thus, a section of line a quarter-wave in length (or any odd multiple of a quarter-wave length) transforms the load at one end to an impedance at the other end by a factor of ρ^2 or $\frac{1}{\rho^2}$, depending on the reference end. Since ρ can be defined in terms of the characteristic impedance of the line section and the impedance of the load, the relationship shown in figure 10-15 can be derived. The line section may be of either the parallel-conductor or coaxial type with the velocity factor applied where needed to make the line an electrical quarter-wave length. However, the transformer functions properly only when the load is nonreactive.



$$Z_O = \sqrt{Z_S Z_R}$$

WHERE:

Z_O = CHARACTERISTIC IMPEDANCE OF $\frac{\lambda}{4}$ LINE SECTION

Z_S = SENDING END IMPEDANCE

Z_R = LOAD IMPEDANCE

Figure 10-15. Quarter-Wave Line Section as Transformer

Theoretically, any value of load impedance Z_r can be transformed to any desired value of impedance Z_s by the quarter-wave line. In actual practice, the range of impedance transformation is limited by the value of Z_o which in turn depends on the physical size of the line. Practical lines yield a range of characteristic impedances from about 50 to 600 ohms. This type of matching section is widely used in antenna feed systems. A few particular applications will be shown later in the chapter.

e. LINE SECTIONS AS BALANCING DEVICES

Frequently it is required to couple energy from an unbalanced system to a balanced system, or vice versa. A common example is center-feeding an elevated dipole with a coaxial transmission line. There is then a need for a device to convert the unbalanced voltage of the coaxial cable to the balanced voltage required by the antenna. If the example antenna were fed directly with the coaxial line, it would not be a strictly balanced load and the line would not operate normally. Currents would be present on the outer surface of the outer conductor of the coaxial line, causing it to radiate, and, in addition, the load presented to the source would not be strictly unbalanced. In the interest of system performance, these conditions should be avoided. A device commonly employed for this transformation is a balun, a contraction for "balanced to unbalanced."

One balun which performs the desired conversion without affecting the impedance characteristics of the system is shown in figure 10-16. Because of its physical appearance, it is sometimes referred to as a "bazooka." When the length, $L = \frac{\lambda}{4}$, the outer sleeve acts with the outer conductor of the enclosed section of line to form a shorted quarter-wave coaxial section, causing a high impedance to exist between points 1 and 2. The result is little or no shunt path to ground from 2 to 1, no division of current at junction 2 and, consequently, equal currents in each conductor of the dual line. The impedance to ground is then only that due to the distributed capacitance and conductance of the balanced load.

Since the currents in each of the dual conductors are equal and 180° out of phase, the voltages to ground are equal and 180° out of phase, which then results in balanced operation.

The impedance between points 2 and 1 remains high only when the length, L, is very near a quarter-wave in length. Consequently, this type of balun provides good performance over a relatively narrow band of frequencies (about 10% of the frequency of operation). Where broader band operation is required, the "double bazooka" type balun, shown in figure 10-17, may be employed. It consists of two sections of the balun in figure 10-16 joined at their open ends as illustrated. From the equivalent circuit, it can be seen that whatever the impedance between points 2 and 1 due to A and B, it will have at all frequencies an exact counterpart between 3 and 1 due to A' and B'.

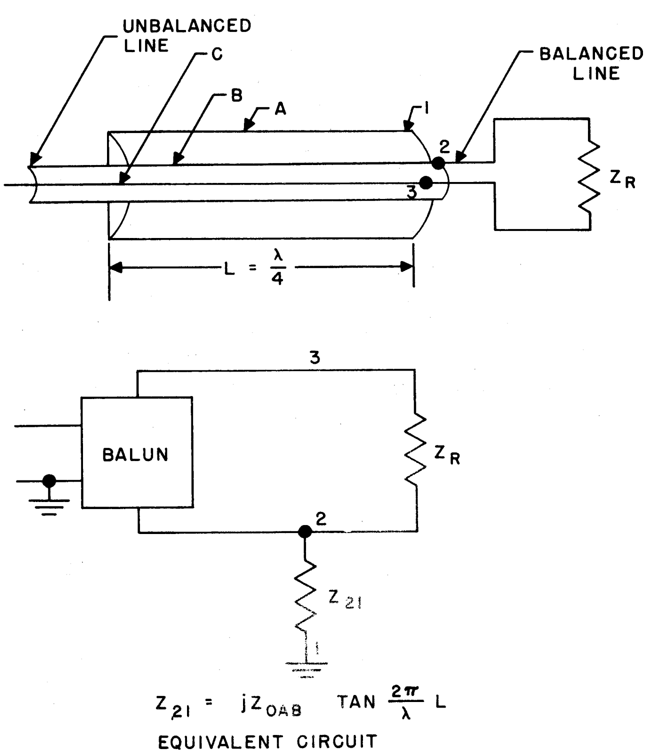


Figure 10-16. Bazooka Type Balun with 1:1 Z-Transfer

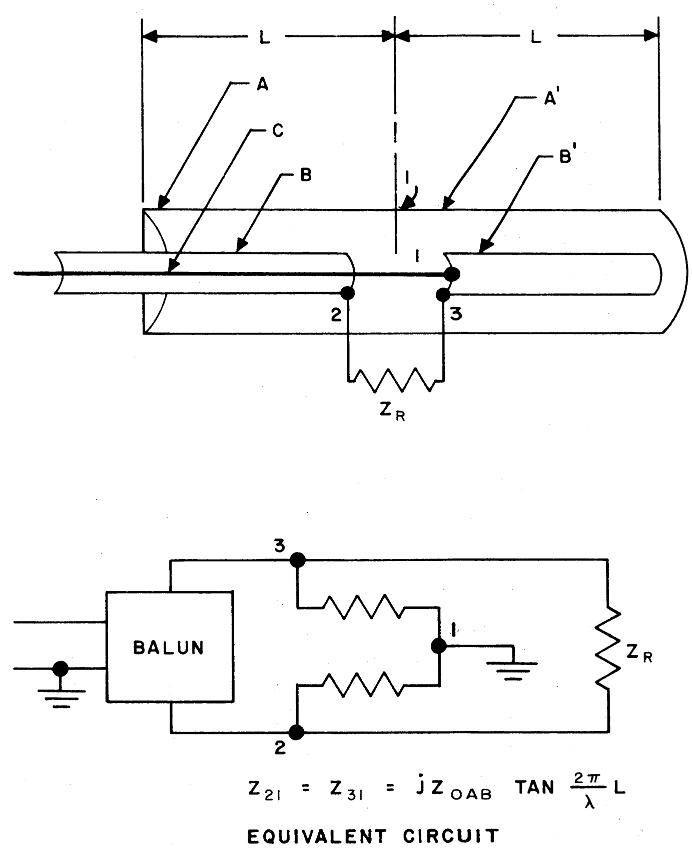


Figure 10-17. Double Bazooka Type Balun with 1:1 Z-Transfer

At the frequency at which each half of the balun is a quarter wave length, the impedance across the transmission line is only that of the dual lines. At any other frequency, the balun will shunt the dual line with an impedance of $2Z_{21}$ or $2Z_{31}$ as shown with the equivalent circuit. Z_{oab} is the surge impedance of each section of the balun.

For best performance, the surge impedance, Z_{oab}, of the coaxial sleeve should be relatively high in the order of 100 ohms. With average size coaxial line for the enclosed portion, this results in a fairly large diameter for the sleeve. For high-frequency application, the physical size and cost may not warrant its use.

A balun, the performance of which is comparable to that of the bazooka for h-f applications, and the construction of which is simpler and more economical, is shown in figure 10-18. This type balun functions

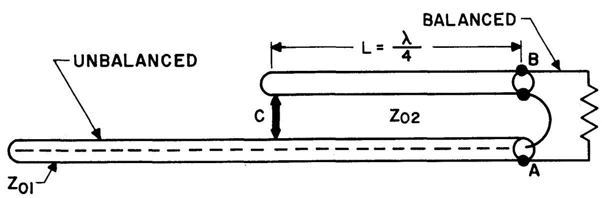


Figure 10-18. Parallel-Conductor Type Balun with 1:1 Impedance Transfer

similarly to the types just described by preventing undesirable currents from flowing along the outside of the coaxial line. Here, however, this is accomplished by a "canceling" effect rather than "choking" off the current as with the quarter-wave sleeve section. Any unbalance currents that flow from A to C due to its direct connection to the antenna (or load) will be canceled at point C by the exact counterpart of equal current flowing from B to C. Thus, there is no current flow in the outside of the coaxial feedline beyond point C.

The length, L, need not be of any particular length for this canceling effect to take place. However, when the length $L = \frac{\lambda}{4}$, the normal impedance characteristics of the load are not upset. It is frequently desirable to operate the balun at other than a quarter wave length in order to take advantage of the shunt reactance it presents to the load for impedance matching purposes. From the discussion on the transmission line section as a parallel-resonant circuit, it becomes apparent that the impedance characteristics of the shorted line section can be utilized here. Since the impedance characteristics of a center-fed half-wave dipole resemble those of a series-resonant circuit, the combined effect of the balun and the dipole result in a wider frequency range of zero reactance and, thus broader band operation. Adding a lumped capacitance across the open end of the balun will permit its physical length to be shortened considerably, as was previously discussed. This is commonly done at the lower frequencies where the parallel resonant circuit properties of the quarter-wave balun are desirable but where its physical length becomes impractical.

In constructing a balun of this type (for high-frequency application) dimensions are usually not very critical. The most important consideration is to establish and maintain the two conductors parallel, which can be done suitably with dielectric spreaders. Spacing between the two conductors depends on the characteristic impedance Z_{O2} desired. In most cases this impedance is not critical, varying from about 150 to 400 ohms. The impedance Z_{O2} can be readily determined from the set of curves shown in figure 10-4. Generally, a higher Z_{O2} is desirable for broad-band purposes. If the spacing between the conductors becomes too wide, however, a series inductance will be added due to the length of the loop connecting points A and B. The effective inductive reactance depends upon the frequency of operation (this may be a desirable or an undesirable effect).

The properties of the balun may be utilized for impedance correction in a variety of ways. Another application of this simple balun is shown in figure 10-19, where it serves also as an impedance transformer. The conversion from unbalance to balance is similar to that of the balun shown in figure 10-18, but due to the indirect loop connection at B and to the length, L, being less than a quarter-wave, the impedance transfer is no longer 1:1. The inductance, L, and capacitance, C, shown in the equivalent circuit are respectively determined by L_1 and Z_{O2} of the shorted section of open line, and by L_2 and Z_{O3} of the open section of coaxial line. These reactances may be calculated from the information given in the discussion of "Line Sections as Reactive Elements" or more conveniently from figure 10-13. The magnitude of X_L and X_C will vary with frequency, so for a given impedance transformation, this device is comparatively narrowband. It is most commonly employed

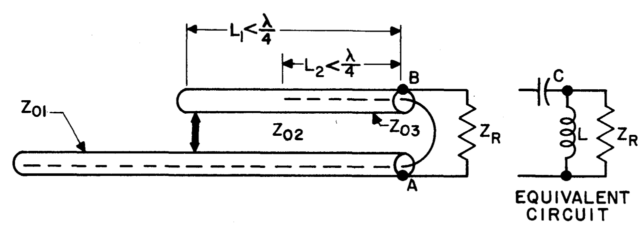


Figure 10-19. Simple Balun Impedance Transformer

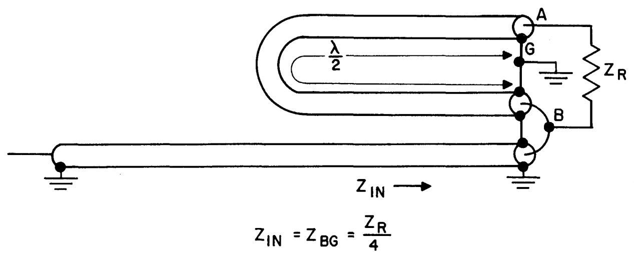


Figure 10-20. Half-Wave Balun with 4:1 Impedance Transfer

where the load impedance Z_R is several times as great as Z_{O1}.

A balun which provides a fixed impedance transformation of 4:1 is shown in figure 10-20. Advantage is taken of the fact that a half-wave section of line repeats its load but with a 180° phase reversal in voltage. Thus the voltages to ground from each side of the load impedance, Z_R, will be equal and 180° out of phase, resulting in balanced operation. The 4:1 impedance transformation may be explained as follows: $Z_{AG} = \frac{Z_R}{2}$ due to Z_R being balanced to ground; this top half of the load impedance appears across BG due to the half-wave line repeating its load. Also appearing across BG is another value of $\frac{Z_R}{2}$ due to the bottom half of the balanced load, Z_R. The two apparent impedances of $\frac{Z_R}{2}$ are in parallel, thus, $Z_{BG} = \frac{Z_R}{4}$. Since the performance of the balun depends on the added section of line being one half-wave in length, it is a narrow-band device. This type balun finds many applications in antenna feedlines, one example of which is coupling the simple folded dipole whose input impedance is about 300 ohms, to a standard 75 ohm coaxial line.

A form of broad-band balun which is useful for application in high-frequency antenna feed systems, particularly in receiving and low-power installations, is illustrated in figure 10-21. It may be visualized as the balun shown in figure 10-18 wound into a coil. Because of the large amount of inductance and distributed capacitance formed by the coil, the length of line required for resonance is considerably reduced. The length of line actually required for a balun of this type can most easily (and practically) be determined by experiment. The coil length, diameter, number of turns, spacing between turns, location, and mounting all have an appreciable effect on the resultant distributed capacitance and inductance.

Although the coil balun of figure 10-21 appears to be a "coiled" version of the parallel-conductor type of figure 10-18, its operation is somewhat more complex. For most practical purposes its operation can be compared to that of the balun shown in figure 10-17, where the linear circuit constants of the coaxial sleeve sections are replaced by the inductance and capacitance formed by the coil. A coil balun is capable of operating over a much wider range of frequencies than is the parallel-conductor type without upsetting the impedance of the system to which it is connected. This can be explained partly by its performing similarly to the balun of figure 10-17 and partly because of its lower Q.

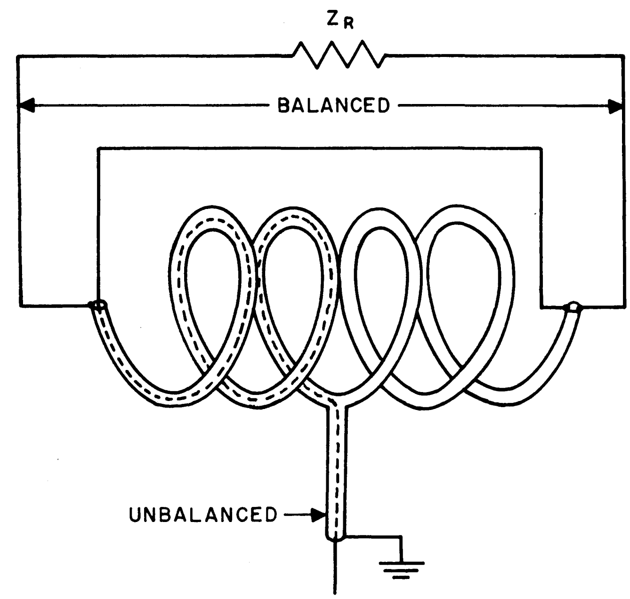


Figure 10-21. Broad Band Coil Balun with 1:1 Impedance Transfer

In constructing a coil balun, assuming the designer has no previous experience with coils of this type, it is reasonable to begin with a length of a pair of lines as required for the full quarter-wave balun of figure 10-18. The coil should then be wound as illustrated to some given diameter and spacing between turns. By experiment, the two ends of the coil should be trimmed symmetrically until resonance is achieved at the desired frequency. A popular method for checking resonance is by the use of the grid-dip meter. This must be done with the balun disconnected from the load and the transmission line, and with the loop connection open. The grid-dip meter method, in fact, may be employed quite conveniently for determining the resonant frequency of baluns previously described. Generally, baluns should be designed to resonate at about the mean frequency of operation.

This chapter does not purport to describe all possible balancing devices which may be employed in high-frequency antenna feed systems. The devices discussed here, however, are the more basic types and are representative of those commonly employed.

6. METHODS OF COUPLING TO THE ANTENNA

a. DIRECT FEEDING

At the lower frequencies, particularly below the high-frequency region, the radiator is commonly brought directly in to the transmitter rather than employing a transmission line. This is done quite frequently in shipboard installations, for example, where a single wire antenna must be used over a wide range of frequencies. Since the antenna is usually not resonant at more than one frequency, an impedance matching network is necessary. In its simplest form, it consists of sufficient series reactance to compensate for the reactance presented by the antenna when it is not resonant. While this matching scheme is simple electrically, it possesses several disadvantages, chiefly the limited frequency range over which it is capable of matching. There are a number of reactive networks that are capable of providing better coupling to the transmitter, one of which is the adjustable "Pi" section shown in figure 10-22.

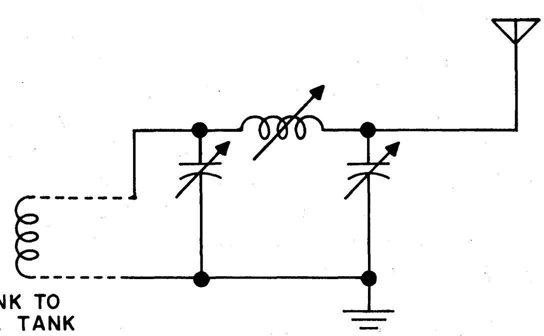


Figure 10-22. Coupling the Directly Fed Antenna Through a Pi Network

b. RESONANT LINE FEEDING

The resonant or tuned-line method of coupling to the antenna is employed where no attempt is made to match the input impedance of the antenna to that of the feedline. The most common application for this type is in harmonically operated antennas, where, in effect, the combination of antenna and feed circuit constitute a resonate system. Consequently, the system is highly frequency sensitive, and a rather complex matching network must be employed at the transmitter unless the feedlines are cut to some critical length which will eliminate the reactive component.

Most generally, the feedline is connected either at a voltage or current loop on the antenna and, therefore, properly designated voltage feed or current feed, respectively. Several typical cases of voltage and current feed on a harmonically operated radiator are shown in figure 10-23. The transmission line is most often the open wire type because the standing-wave ratio on the feedline is usually high.

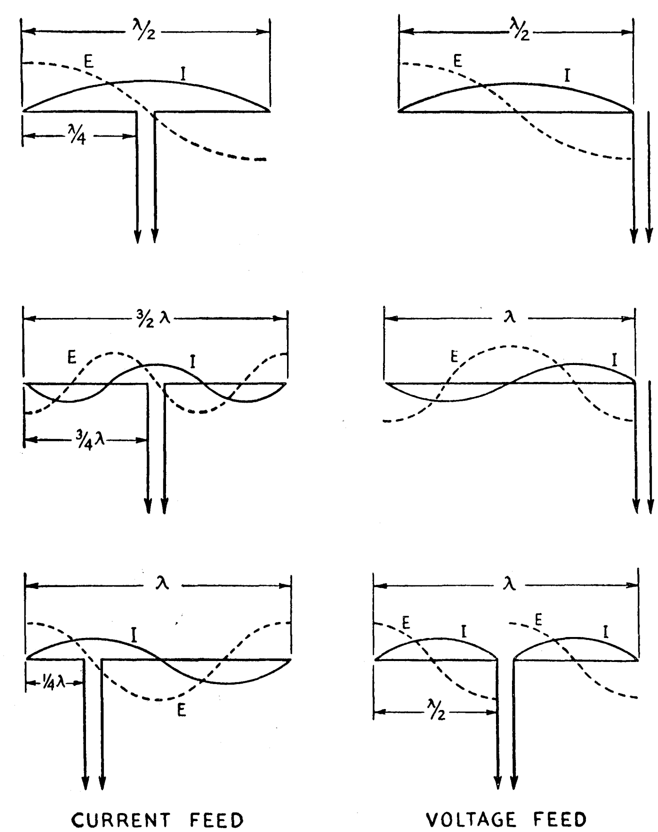
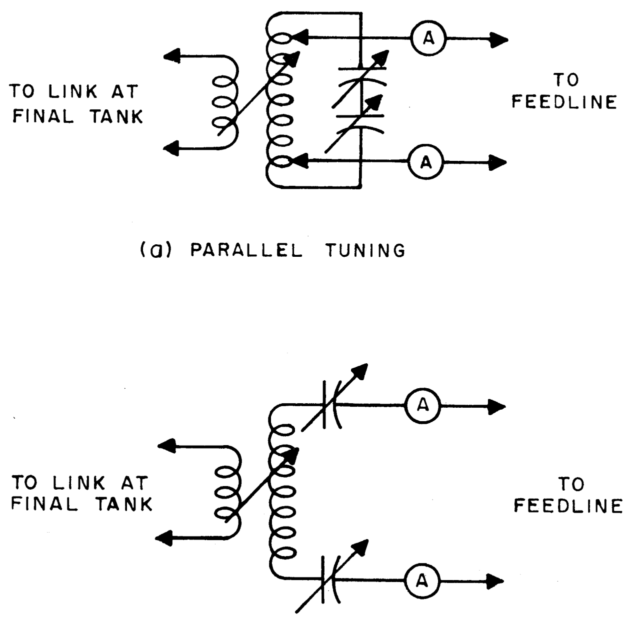


Figure 10-23. Current and Voltage Feed in Antennas Operated at One, Two, and Three Times Fundamental Frequency

Impedance conditions at the transmitter (input) end of the feedline will vary considerably with frequency, since its termination is not matched. The impedance presented to the transmitter depends on the terminating impedance and the feedline length. It is common practice to cut the line to a length where it will be a multiple of a quarter-wave at the operating frequency, so either a voltage or a current loop will occur near the input end. This is assuming that the antenna is resonant at the harmonic frequency and that it is being fed at either a voltage or current loop. The impedance at the transmitter end will then be essentially resistive, and its value will be above or below that of the line characteristic impedance depending upon whether the loop at the input end is voltage or current, respectively.

It is usually more practical to compensate for the unmatched condition by employing an adjustable coupling network at the transmitter end of the line than by varying the line length for a given coupling coil. The coupling networks shown in figure 10-24 are frequently employed. When a voltage loop appears at the input end ($Z_{in} > Z_o$), the parallel-tuned coupler of figure 10-24a is used; when a current loop appears ($Z_{in} < Z_o$), the series-tuned coupler of figure 10-24b is used.



(a) PARALLEL TUNING

(b) SERIES TUNING

Figure 10-24. Coupling Networks for Resonant Feedlines

Since the swr is usually high when the tuned-line method of feeding is used, the line is subject to more losses and lower power limitations. Therefore, when the antenna is operated at only one frequency, there is generally no point in using this method.

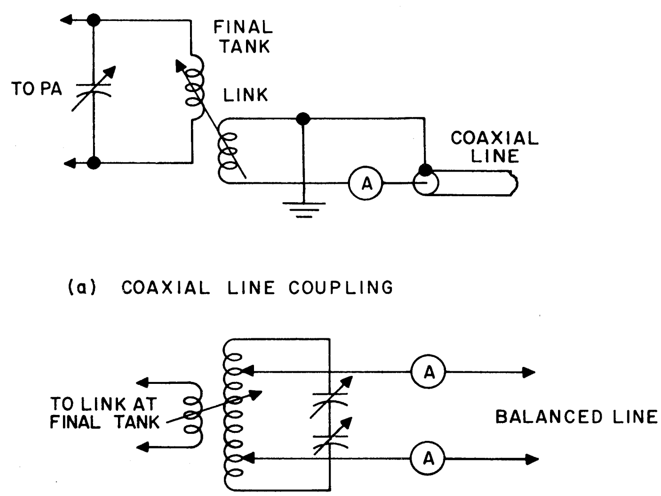
c. NONRESONANT LINE FEEDING

Unlike the resonant line method of coupling power to the antenna, the use of the nonresonant or untuned line imposes the restriction that the line be operated with a low standing-wave ratio. Except in a few special cases, the system characteristics are such that some impedance matching and/or balancing techniques are required. This method of antenna coupling is highly efficient and is by far the one most commonly used in modern communication systems.

It is important to note that before an attempt is made to match the antenna to the feedline, the antenna itself should be made resonant so that a resistive load will appear at the end of the line. An over-all matched system will then be less difficult to achieve, particularly in systems employing antennas with high Q. Furthermore, for a less frequency sensitive and more efficient system, the impedance matching should be performed as near to the load as practical.

In the case where the input impedance of a given antenna matches the characterisitc impedance of a practical transmission line, no problem of coupling exists provided the transmitter and the radiator are both designed for the same condition of balance or unbalance. If either operates unbalanced while the other operates balanced, one of the balanced-to-unbalanced techniques (with a 1:1 impedance transfer) previously described should be incorporated in the feed system. Since this condition is not one of actual impedance correction, the position of the balun in the feedline is not important.

With untuned line feeding, where the swr on the line is low, there is no particular problem in coupling the line to the transmitter. The feedline can be matched to the final tank with various simple coupling schemes, examples of which are illustrated in figure 10-25.



(b) BALANCED LINE COUPLING

Figure 10-25. Coupling to Nonresonant Balanced and Unbalanced Lines

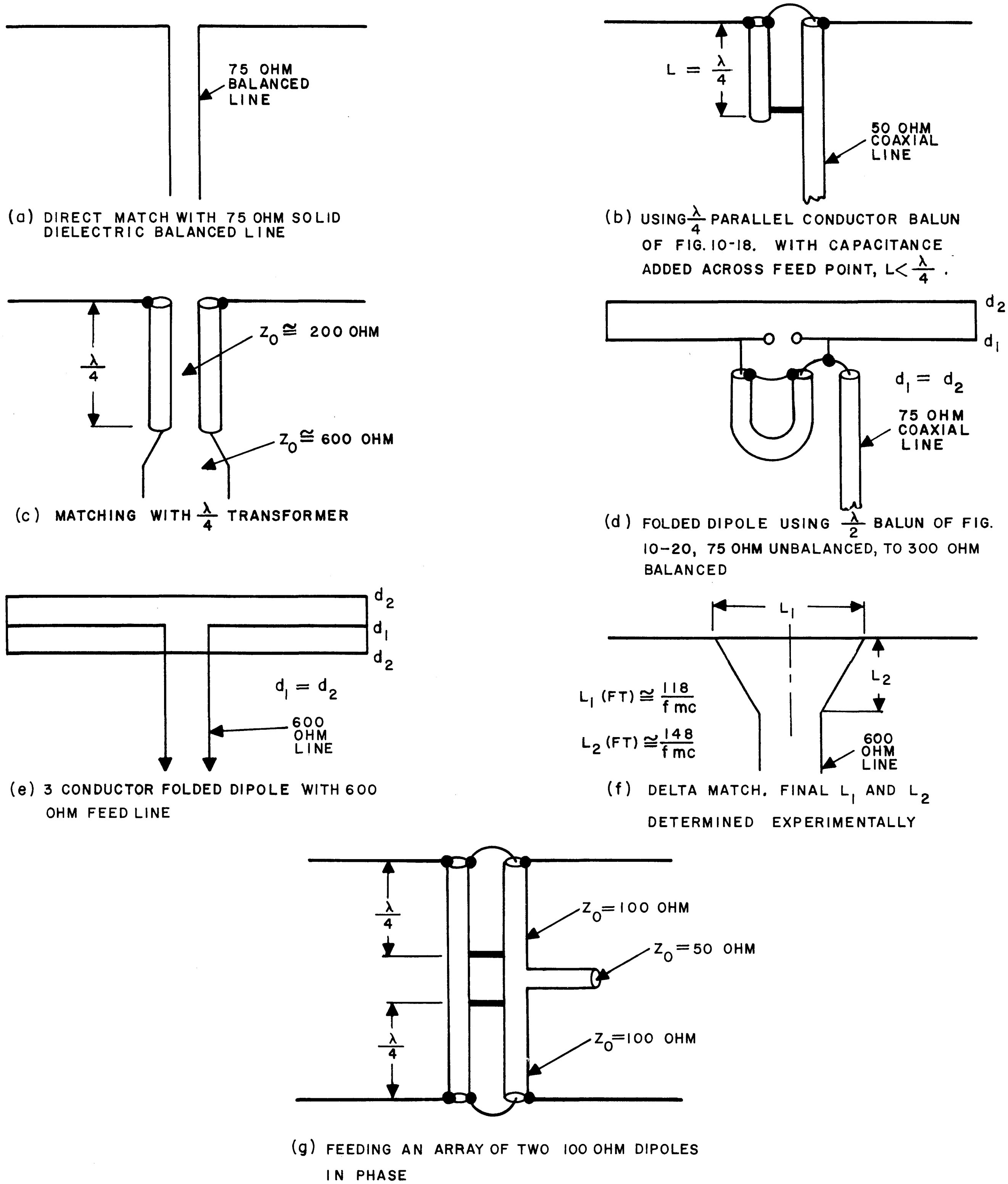
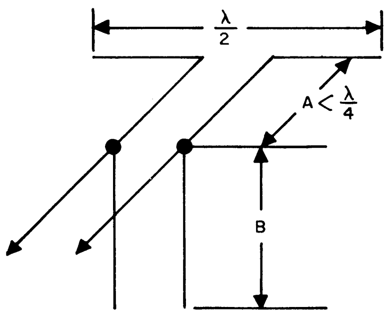
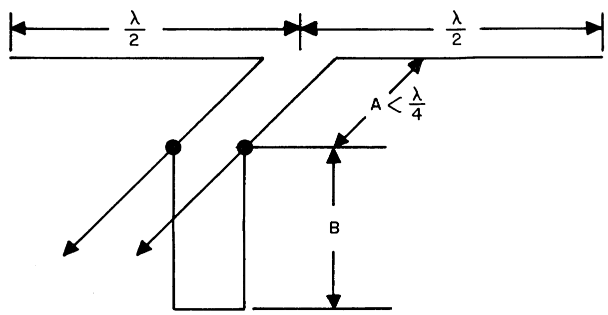


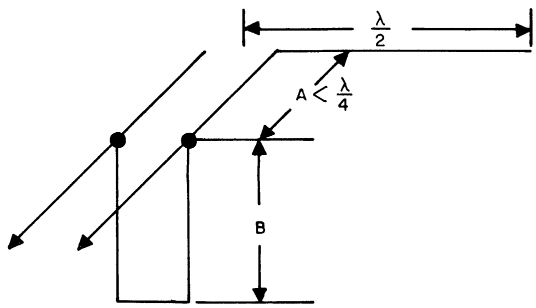
Figure 10-26. Typical Coupling to Center-Fed Half-Wave Dipoles



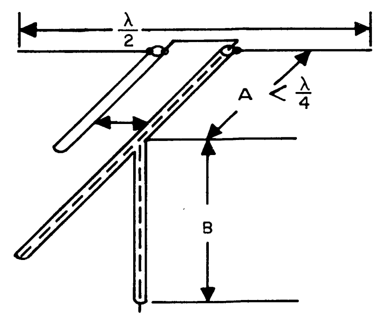
(a) CENTER-FED HALF-WAVE



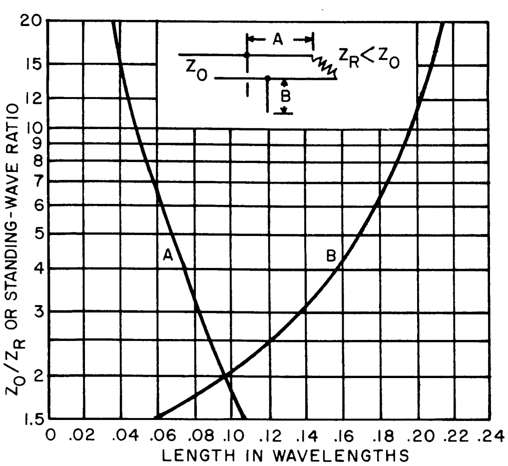
(b) CENTER-FED FULL-WAVE



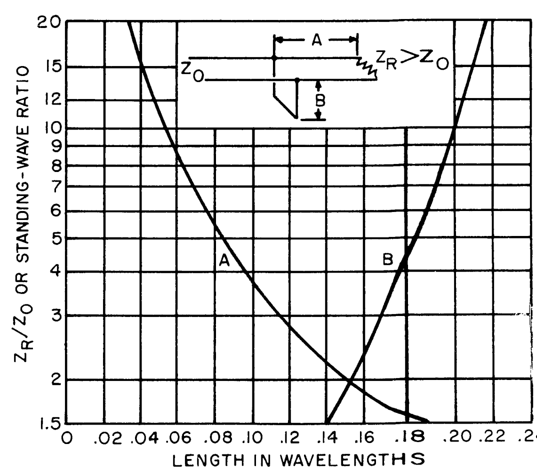
(C) END-FED HALF-WAVE



(d) CENTER-FED HALF-WAVE WITH COAXIAL LINE AND STUB



SWR FOR VARIOUS OPEN STUB LENGTHS AND POSITIONS AS A FUNCTION OF Z_O/Z_R OR STANDING WAVE RATIO.



SWR FOR VARIOUS CLOSED STUB LENGTHS AND POSITIONS AS A FUNCTION OF Z_R/Z_O OR STANDING WAVE RATIO.

Figure 10-27. Typical Applications of Matching Stubs. The curves shown apply only if the characteristic impedance of the line and the stub are the same and the load is resistive. When applicable, modify lengths of A and B by the appropriate velocity factor. Where coaxial lines are used, the same principles apply.

The following illustrations (figures 10-26, 10-27, and 10-28) are typical applications of some of the antenna-to-line coupling methods and matching techniques previously described.

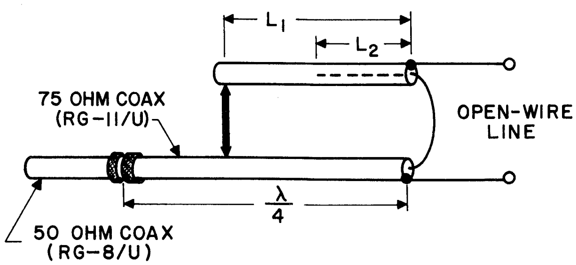
7. MEASUREMENTS

Although there are a number of measurements that can be made during the adjustment or evaluation of a complex antenna system, the following discussion will be limited to measurements of impedance and directivity. These two measurements are generally the most important and usually will adequately describe the operating characteristics of the antenna system.

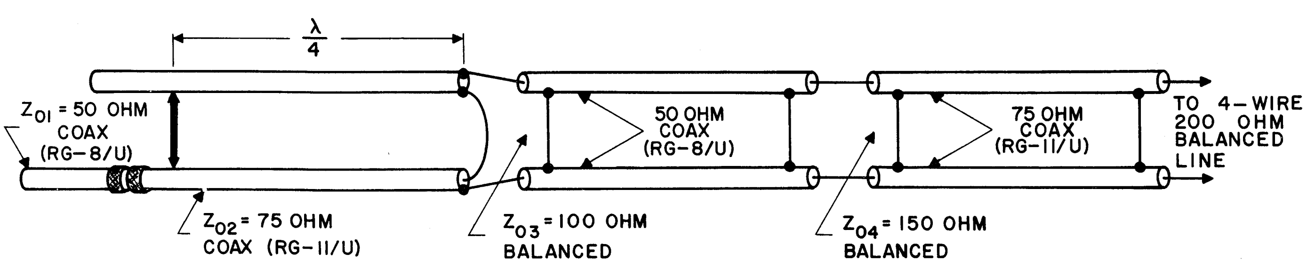
Probably the most common method of measuring the actual impedance of the antenna or the impedance at some point on the transmission line is by the use of an r-f impedance or admittance bridge. The information obtained from bridge measurements often requires a great deal of data reduction before the desired information is obtained. However, when the resistive and reactive components of impedance are required, particularly in coaxial feed systems, the bridge is an essential item.

In feed systems employing open wire lines, it becomes practical to locate the nodes and loops of the standing waves on the line and to measure their relative magnitude. When both null position and swr is known, the resistive and reactive components of a complex impedance may be determined. The swr can be obtained more directly by this method, but to determine the actual impedance, the amount of data reduction required is comparable to that when using a bridge. Since this method locates the loops and nodes of the standing waves, it is very useful when applying impedance-matching stubs. The standing waves are generally investigated with a voltage or current indicator which has adequate sensitivity and does not disturb the normal operation of the line.

When only the swr is required, a directional coupler device which responds to power in only one direction is frequently used. The incident power and the reflected power can then be measured and related to the swr by



(a) BALUN AND Z TRANSFORMER FOR MATCHING 50 OHM UNBALANCED LINE TO OPEN-WIRE LINE, USING $\frac{\lambda}{4}$ TRANSFORMER OF Z_0 = 75 OHMS AND Z STEP-UP BALUN OF FIG 10-19. WHERE L_1 AND $L_2 << \frac{\lambda}{4}$, CURVES OF FIG 10-13 MAY BE USED. A GIVEN OUTPUT IMPEDANCE IN THE RANGE OF ABOUT 250-550 OHMS MAY BE MATCHED WITH PROPER COMBINATION OF L_1 AND L_2.



(b) BROADBAND Z TAPERED TRANSFORMER FOR MATCHING 50 OHM UNBALANCED TO 200 OHM BALANCED.

Figure 10-28. Typical Applications of Practical Transmission Lines for Impedance Matching

$$\text{swr} = \rho = \frac{\sqrt{\frac{P_i}{P_r}} + 1}{\sqrt{\frac{P_i}{P_r}} - 1}$$

where P_i = Incident or forward power

P_r = Reflected power

For convenience, this relationship is plotted in figure 10-29 for typical values of power and swr.

An important factor to consider when making swr or impedance measurements is the effect of line attenuation on the actual swr. On a lossy line, the measured swr is less than the actual swr. The degree of error introduced is dependent on the amount of line attenuation and the amount of mismatch at the point of interest. Refer to the swr correction curves of figure 10-30, and apply the correction to the apparent swr when necessary. The attenuation of some typical coaxial transmission lines is shown in figure 10-31.

The measurement of the directional properties of a high-frequency antenna usually presents a more complex problem than that of measuring its impedance. This is especially true when the antenna under test is a fixed installation.

In order that the directional characteristics of a given antenna be completely known, the relative field intensity must be measured in at least the two principal planes (azimuth and elevation). This requires that in both planes field strength measurements be taken at sufficient increments around paths of constant range from the antenna. In most cases where the test antenna is fixed, these measurements are taken in flight with a properly equipped aircraft and experienced personnel. It is apparent that measurements requiring this technique can be involved and costly.

If the antenna under test can be rotated, pattern measurements, at least in the azimuth plane, are less difficult. With a pick-up antenna and a field strength meter located at a fixed remote point, the test antenna may be rotated through 360° in the azimuth plane and its relative field strength recorded. When its vertical directivity is required, a problem similar to that of the fixed installation exists. Because of the effect of ground on vertical directivity, vertical plane patterns derived from azimuth plane rotation would be under free-space conditions and would not be representative of the actual antenna. In a case like this, however, if the free-space vertical pattern is known, a theoretical ground factor may be applied which will modify the measured pattern. If actual ground conditions are known, the corrected pattern will be representative of the antenna under actual conditions.

In the development of high-frequency antennas large in dimension, the engineer will most often use scale model techniques. The antenna is scaled down to a size that is practical to rotate for determining pattern characteristics. The frequency is scaled up by the same factor. Scale model techniques are also applicable to investigation of impedance.

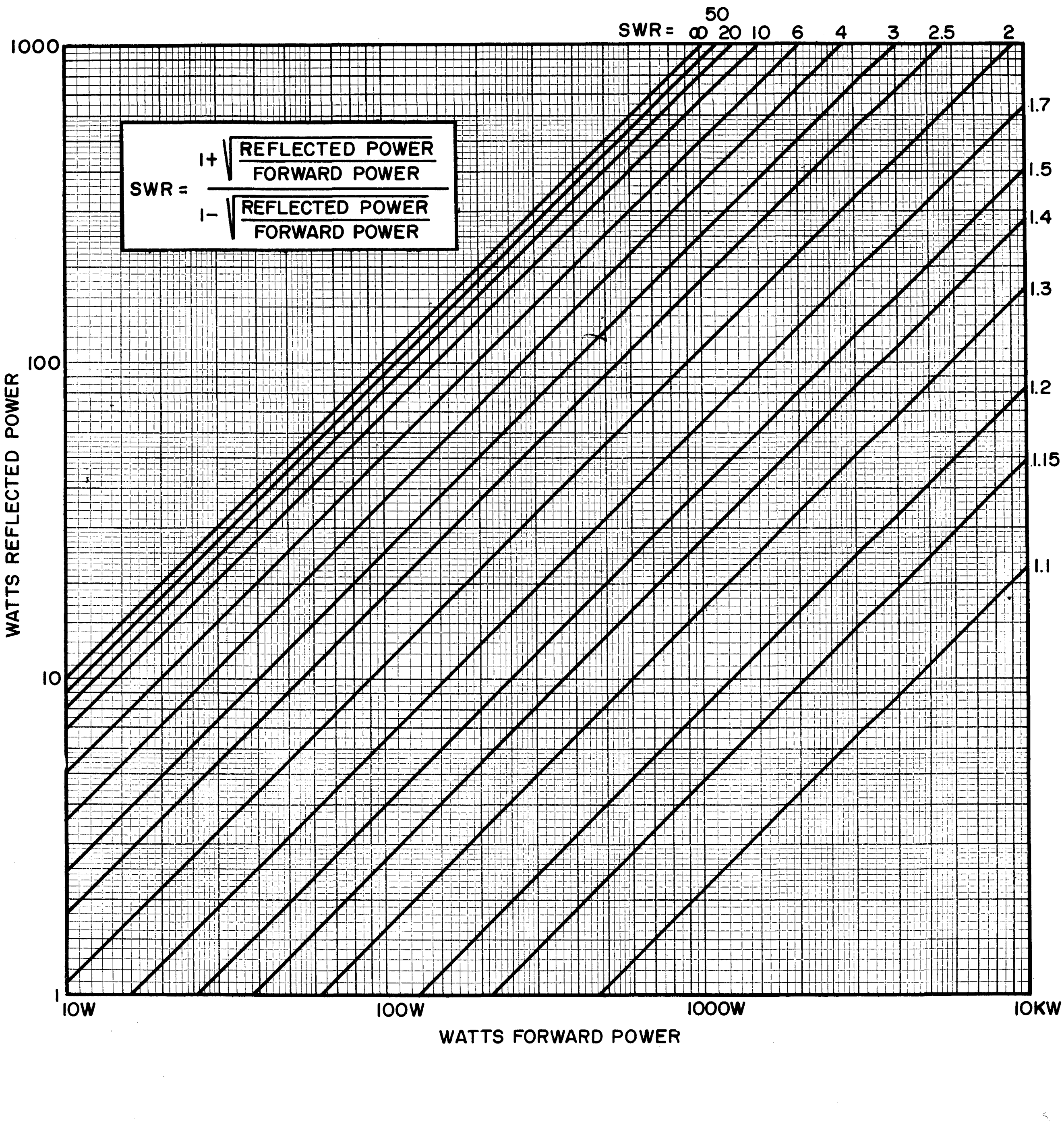


Figure 10-29. Relationship of SWR to Incident and Reflected Power

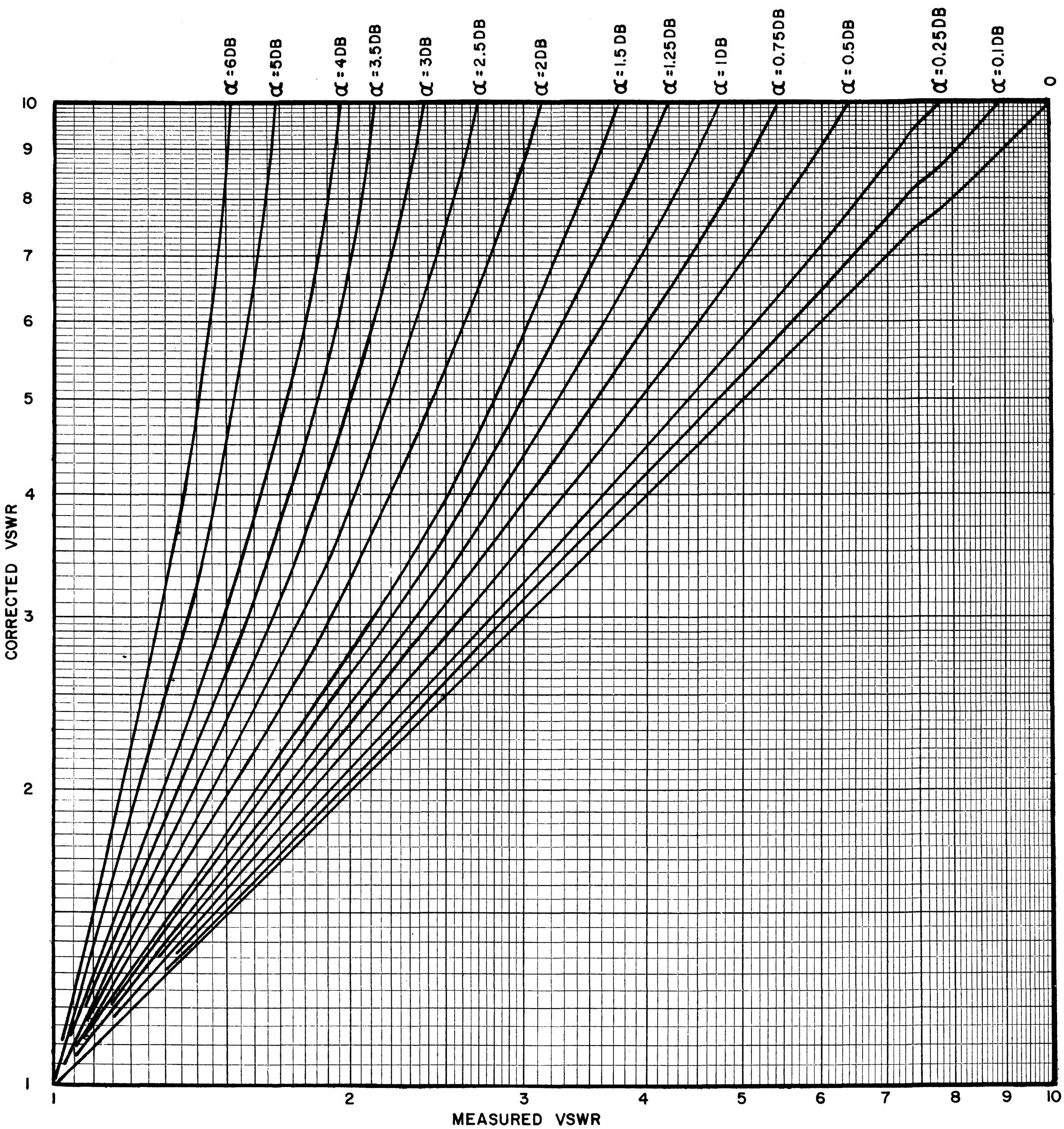


Figure 10-30. VSWR Correction Curves for Attenuation on Transmission Line

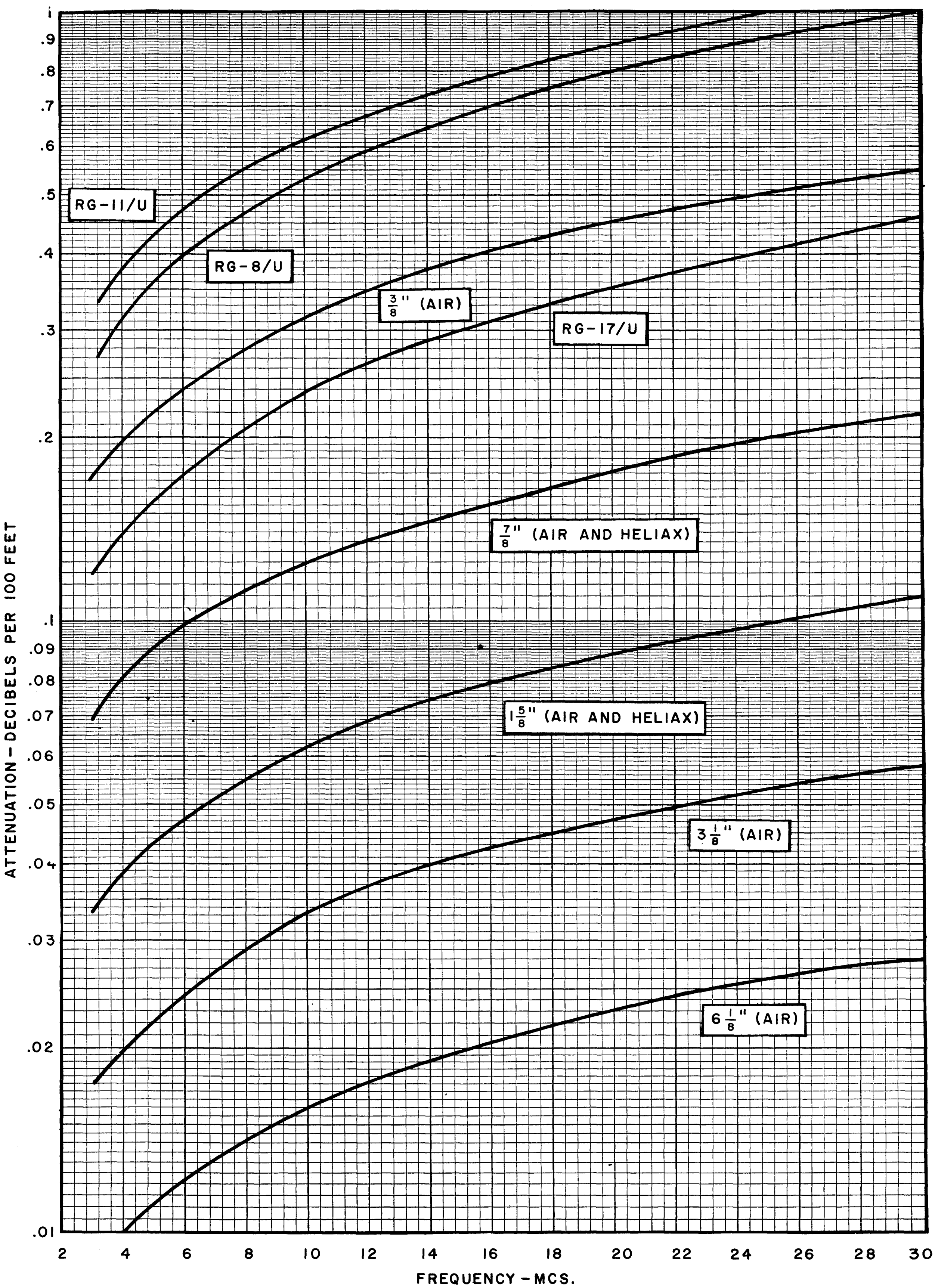


Figure 10-31. Attenuation Data for Common Types of Coaxial Transmission Lines