

# **Practical Antenna Handbook**

*Fourth Edition*

*Joseph J. Carr*

**McGraw-Hill**

# **Practical Antenna Handbook**

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# Practical Antenna Handbook

Fourth Edition

Joseph J. Carr

**McGraw-Hill**

New York Chicago San Francisco Lisbon London  
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## **Dedication**

Dedicated in memoriam to Johnnie Harper Thorne, K4NFU:  
a friend and colleague for nearly 30 years who is sorely missed.  
Johnnie was a genius who knew some real smoke about antennas.  
Killed by a drunk driver. . .

## **From the Publisher**

The fourth edition of this book is published in memory of Joe Carr who passed away shortly after completing his manuscript. Joe was a valuable member of the McGraw-Hill family of authors and the electronics community as a whole. He will be missed by us all.

*Scott L. Grillo*  
*Editor-in-Chief*

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# Contents

Introduction to the Fourth Edition	<i>ix</i>
1 Introduction to Radio Broadcasting and Communications	<i>1</i>
2 Radio-wave Propagation	<i>5</i>
3 Transmission Lines	<i>59</i>
4 The Smith Chart	<i>95</i>
5 Fundamentals of Radio Antennas	<i>123</i>
6 High-Frequency Dipole and Other Doublet Antennas	<i>141</i>
7 Vertically Polarized HF Antennas	<i>173</i>
8 Multiband and Tunable-Wire Antennas	<i>203</i>
9 Longwire Directional Antennas	<i>213</i>
10 Hidden and Limited-Space Antennas	<i>231</i>
11 Directional Phased Vertical Antennas	<i>245</i>
12 Directional Beam Antennas	<i>255</i>

<b>13</b>	<b>Antennas for Shortwave Reception</b>	<b>271</b>
<b>14</b>	<b>Large Wire Loop Antennas</b>	<b>287</b>
<b>15</b>	<b>Small Loop Receiving Antennas</b>	<b>299</b>
<b>16</b>	<b>Small Transmitting Loop Antennas</b>	<b>319</b>
<b>17</b>	<b>Antenna Modeling Software</b>	<b>327</b>
<b>18</b>	<b>VHF/UHF Transmitting and Receiving Antennas</b>	<b>339</b>
<b>19</b>	<b>Microwave Waveguides and Antennas</b>	<b>369</b>
<b>20</b>	<b>Antenna Noise Temperature</b>	<b>417</b>
<b>21</b>	<b>Antennas for Radio Astronomy</b>	<b>421</b>
<b>22</b>	<b>Adjusting, Installing, and Troubleshooting Antennas and Transmission Lines</b>	<b>433</b>
<b>23</b>	<b>Antennas for Radio Direction Finding (RDF)</b>	<b>439</b>
<b>24</b>	<b>Impedence Matching in Antenna Systems</b>	<b>457</b>
<b>25</b>	<b>Mobile, Emergency, Portable, and Marine Antennas</b>	<b>479</b>
<b>26</b>	<b>Antennas for Low-Frequency Operation</b>	<b>501</b>
<b>27</b>	<b>Measurement and Adjustment Techniques</b>	<b>515</b>
<b>28</b>	<b>General Antenna Mechanical Construction Techniques</b>	<b>543</b>
<b>29</b>	<b>Grounding the Antenna: What Is a Good Ground?</b>	<b>573</b>
	<b>Index</b>	<b>583</b>

# Introduction to the fourth edition

IT WAS WITH GREAT PLEASURE THAT I WROTE THIS FOURTH EDITION OF *PRACTICAL Antenna Handbook*. The original intent when the first edition was planned was to provide the reader with a practical, yet theoretical, book that could be used with only a minimal effort to actually design and install radio antennas. It was assumed that the readership would possess a wide range of levels of antenna sophistication, from the novice “newbie” to the professional engineer. That assumption proved to be correct as I have received letters from a wide variety of people.

The success of the three previous editions of *Practical Antenna Handbook* has been most gratifying. Clerks in radio stores told me that this book was outselling other titles 2:1. They confirmed that the types of people who buy the book meet the profile above. It was with surprise and delight that I noted that clerks in one radio store could cite the McGraw-Hill catalog number from memory, when they could not do that for other antenna books.

While the sales are deeply appreciated (they are an honor), there was one comment that stood out. The salesman at one radio store introduced me to an instructor from a U.S. Government communications school. He typically bought twenty copies of *Practical Antenna Handbook* at a time for use by his students in a training class. He told me that the reason why he selected my book over others was “...it’s the *only book on the market* that people can give to a secretary, or clerk-typist, and expect them to be able to put up a working half-wavelength dipole two hours later.” And, he stated, in his business that could literally happen any time.

Antennas have changed a lot over the years. Figure I-1 shows how antennas were between 1913 and 1940. This facility was the first U.S. Navy radio station (NAA) in Arlington, VA.<sup>1</sup> Those two 600- and 400-ft towers were taken down in 1940 to make room for National Airport (Washington, D.C.). The towers were reassembled in Annapolis, MD, where they lasted until recently.

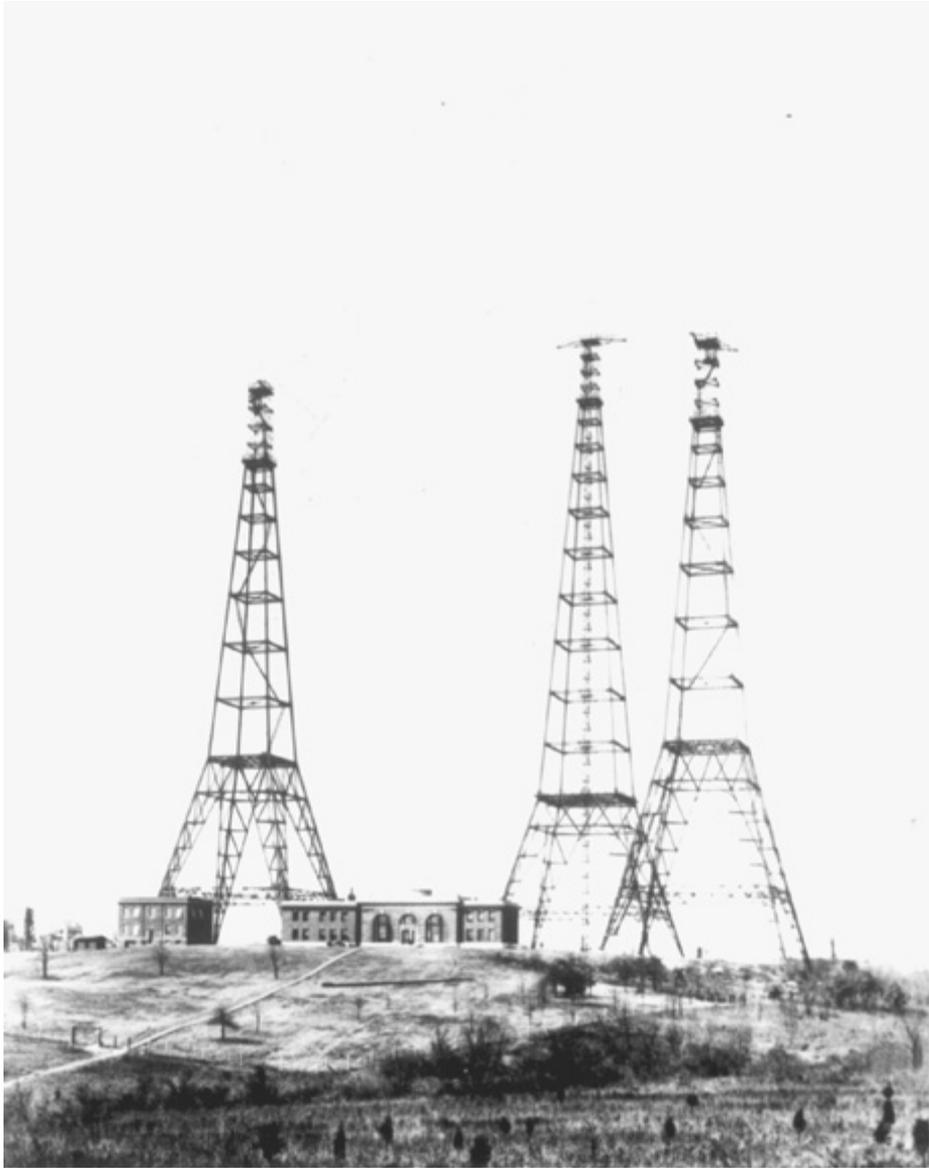
Antenna technology has changed since 1940. In preparing this fourth edition of *Practical Antenna Handbook* all previous material was reviewed for accuracy and relevance. Additional material was added for the following topics:

- Wire antenna construction methods
- Antenna modeling software (miniNEC, WinNEC, and EZNEC for Windows)
- Antenna noise temperature
- Antennas for radio astronomy
- Antennas for Radio Direction Finding

I hope that you find this fourth edition as useful as the previous three editions. And thank you very much for honoring me by spending your hard-earned money on my book.

Joseph J. Carr, MSEE

<sup>1</sup> The NAA callsign has since been reassigned to the VLF station at Cutler, ME.



**I-1** Massive antenna towers at Navy “Radio Arlington” (1913-1940). (Courtesy of the NAA.)

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# 1 CHAPTER

# Introduction to radio broadcasting and communications

RADIO BROADCASTING AND COMMUNICATIONS SEEMS TO HOLD A STRANGE KIND OF magical allure that attracts a wide variety of people and holds them for years. There is something fascinating about the ability to project yourself over vast intercontinental distances.

Radio communications have been with us now for the entire twentieth century and into the twenty-first. Experiments are on record as early as 1867, and by the turn of the century “wireless telegraphy” (as radio was called then) sparked the imaginations of countless people across the world. Radio communications began in earnest, however, when Guglielmo Marconi successfully demonstrated wireless telegraphy as a commercially viable entity. The “wireless” aspect to radio so radically changed communications that the word is still used to denote radio communications in many countries of the world. Marconi made a big leap to international fame on a cold December day in 1903, when he and a team of colleagues successfully demonstrated transatlantic wireless telegraphy. Until that time, wireless was a neighborhood—or cross-town at best—endeavor that was of limited usefulness. Of course, ships close to shore, or each other, could summon aid in times of emergency, but the ability to communicate over truly long distances was absent. All that changed on that fateful day in Newfoundland when Marconi heard the Morse letter “S” tickle his ears.

Wireless telegraphy was pressed into service by shipping companies because it immediately provided an element of safety that was missing in the prewireless days. Indeed, a ship that sank, leaving its crew and passengers afloat on a forbidding sea, was alone. Any survivors often succumbed to the elements before a chance encounter with a rescue vessel. Some early shipping companies advertised that their ships were safer because of wireless aboard. It was not until 1909, however, that wireless telegraphy proved its usefulness on the high seas. Two ships collided in the foggy Atlantic Ocean and were sinking. All passengers and crew members of both ships were in imminent danger of death in the icy waters. But radio operator Jack

## 2 *Introduction to radio broadcasting and communications*

Binns became the first man in history to send out a maritime distress call. There is some debate over which distress call Binns transmitted, but one thing is certain: It was not “SOS” (today’s distress call), because SOS was not adopted until later. Binns probably transmitted either “CQD” or “CQE,” both of which were recognized in those days before standardization. Regardless of which call was sent, however, it was received and relayed from ship to ship, allowing another vessel to come to the aid of the stricken pair of ships.

All radio prior to about 1916 was carried on via telegraphy (i.e., the on-off keying of a radio signal in the Morse code). But in 1916 some more magic happened. On a little hill in Arlington, Virginia, on a site that now overlooks the Pentagon and the U.S. Marine Corps base called *Henderson Hall*, there were (and still are) a pair of two-story brick buildings housing the naval radio station NAA (callsign since reassigned to the VLF station at Cutler, ME). On a night in 1916, radio operators and monitors up and down the Atlantic seaboard—from the midwest to the coast and out to sea for hundreds of miles—heard something that must have startled them out of their wits, for crackling out of the “ether,” amidst the whining of Alexanderson alternators and “ZZZCHHT” of spark-gap transmitters, came a new sound—a human voice. Engineers and scientists at the Naval Research Laboratory had transmitted the first practical amplitude-modulated (AM) radio signal. Earlier attempts, prior to 1910, had been successful as scientific experiments, but they did not use commercially viable equipment.

Although radio activity in the early years was unregulated and chaotic, today it is quite heavily regulated. Order was brought to the bands (don’t laugh, ye who tune the shortwaves) that was lacking before. Internationally radio is regulated by the International Telecommunications Union (ITU) in Geneva, Switzerland through the treaties arising from World Administrative Radio Conferences (WARC) held every 10 to 15 years. In the United States, radio communications are regulated by the Federal Communications Commission (FCC), headquartered in Washington, D.C.

Amateur radio has grown from a few thousand “hams” prior to World War I to more than 900,000 today, about one-third of them in the United States. Amateur operators were ordered off the air during World War I, and almost did not make a comeback after the war. There were, by that time, many powerful commercial interests that greedily coveted the frequencies used by amateurs, and they almost succeeded in keeping postwar amateurs off the air. But the amateurs won the dispute, at least partially. In those days, it was the frequencies with wavelengths longer than 200 m (i.e., 20 to 1500 kHz) that were valuable for communications.

The cynical attitude attributed to the commercial interests regarding amateurs was, “put ’em on 200 meters and below . . . they’ll never get out of their backyards there!” But there was a surprise in store for those commercial operators, because the wavelengths shorter than 200 m are in the high-frequency region that we now call “shortwaves.” Today, the shortwaves are well-known for their ability to communicate over transcontinental distances, but in 1919 that ability was not suspected.

I once heard an anecdote from an amateur operator “who was there.” In the summer of 1921 this man owned a large, beautiful wire “flattop” antenna array for frequencies close to 200 m on his family’s farm in southwestern Virginia. Using those frequencies he was used to communicating several hundred miles into eastern Ohio and down to the Carolinas. But, in September 1921 he went to college to study

electrical engineering at the University of Virginia in Charlottesville. When he returned home for Thanksgiving he noticed that his younger brother had replaced the long flattop array with a short dipole antenna. He was furious, but managed through great effort to contain the anger until after dinner. Confronting his brother over the incredible sacrilege, he was told that they no longer used 150 to 200 m, but rather were using 40 m instead. Everyone “knew” that 40 m was useless for communications over more than a few blocks, so (undoubtedly fuming) the guy took a turn at the key. He sent out a “CQ” (general call) and was answered by a station with a call sign like “8XX.” Thinking that the other station was in the 8th U.S. call district (WV, OH, MI) he asked him to relay a message to a college buddy in Cincinnati, OH. The other station replied in the affirmative, but suggested that “. . . you are in a better position to reach Cincinnati than me, I am *FRENCH* 8XX.” (Call signs in 1921 did not have national prefixes that are used today.) The age of international amateur communications had arrived! And with it came a new problem—the national identifiers in call signs became necessary (which is why American call signs begin with K, W, or N).

During the 1930s, radio communications and broadcasting spread like wildfire as technology and techniques improved. World War II became the first war to be fought with extensive electronics. Immediately prior to the war, the British developed a new weapon called RADAR (radio detection and ranging; now spelled radar). This tool allowed them to see and be forewarned of German aircraft streaming across the English Channel to strike targets in the United Kingdom. The German planes were guided by (then sophisticated) wireless highways in the sky, while British fighters defended the home island by radio vectoring from ground controllers. With night fighters equipped with the first “centimetric” (i.e., microwave) radar, the Royal Air Force was able to strike the invaders accurately—even at night. The first kill occurred one dark, foggy, moonless night when a *Beaufighter* closed on a spot in the sky where the radar in the belly of the plane said an enemy plane was flying. Briefly thinking he saw a form in the fog, the pilot cut loose a burst from his quad mount of 20-mm guns slung in the former bomb bay. Nothing. Thinking that the new toy had failed, the pilot returned to base—only to be told that ground observers had reported that a German *Heinkle* bomber fell from the overcast sky at the exact spot where the pilot had his ghostly encounter.

Radio, television, radar, and a wide variety of services, are available today under the general rubric “radio communications and broadcasting.” Homeowners, and other nonprofessionals in radio, can own a receiver system in their backyard that picks up television and radio signals from satellites in geosynchronous orbit 23,000 mi out in space. Amateur operators are able to communicate worldwide on low power, and have even launched their own “OSCAR” satellites.

Some people had written off the HF radio spectrum in recent years, citing satellite technology as the reason. But the no-code license for amateur radio operators, which does not carry HF privileges, has proven to be a stepping stone to higher-class licenses, which do. Also, the shortwave broadcasting market received a tremendous boost during the Gulf War. When the troops of *Operation Desert Shield* and *Desert Storm* were assembling to take back Kuwait from the Iraqis, the sales of shortwave receivers jumped dramatically. And, following January 16, 1991, when the forces started pouring across the border into the actual fight, the sales

#### 4 *Introduction to radio broadcasting and communications*

skyrocketed out of sight. One dealer told me that he couldn't keep receivers in stock, and that he had sold out most models. That interest seems to have matured into long-term interest on the part of a significant number of listeners and new ham operators.

The antenna is arguably one of the most important parts of the receiving and/or transmitting station (Fig. 1-1). That is what this book is all about.



**1-1** This AM/FM broadcast antenna tower bristles with two-way antennas.

## 2 CHAPTER

# Radio-wave propagation

THE PROPAGATION OF RADIO SIGNALS IS NOT THE SIMPLE MATTER THAT IT SEEMS AT first glance. Intuitively, radio signal propagation seems similar to light propagation; after all, light and radio signals are both electromagnetic waves. But simple inverse square law predictions, based on the optics of visible light, fall down radically at radio frequencies because other factors come into play. In the microwave region of the spectrum, the differences are more profound because atmospheric pressure and water vapor content become more important than for light. For similar reasons, the properties of microwave propagation differ from lower VHF and HF propagation. In the HF region, solar ionization of the upper reaches of the atmosphere causes the kind of effects that lead to long-distance “skip” communications and intercontinental broadcasting. This chapter examines radio propagation phenomena so that you have a better understanding of what an antenna is used for and what parameters are important to ensure the propagation results that you desire.

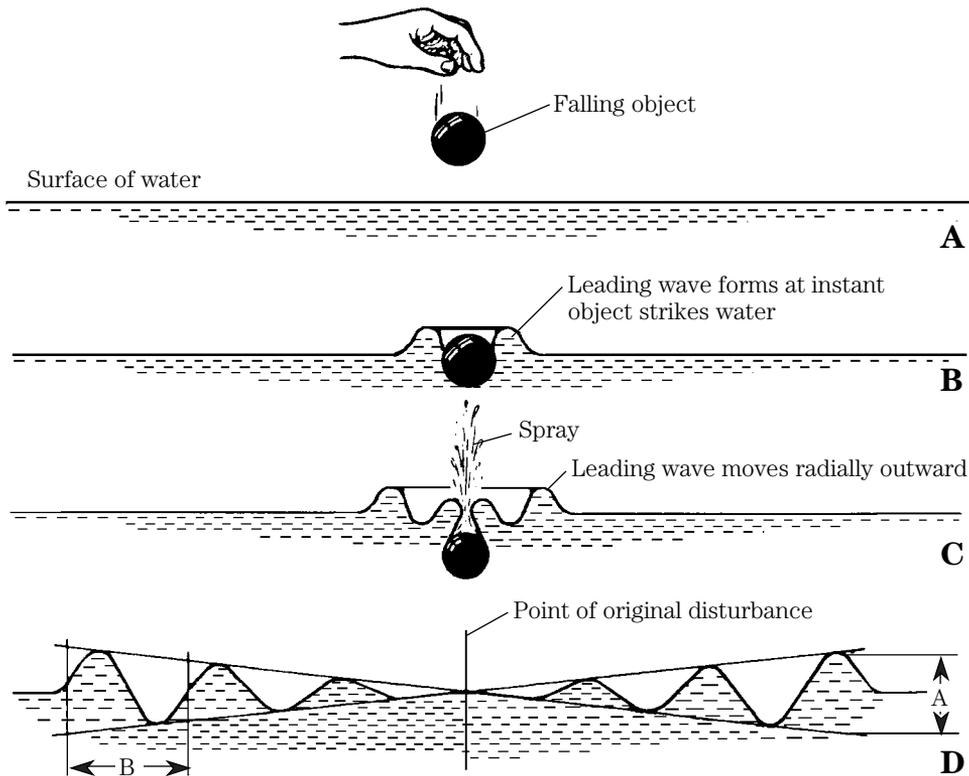
## Radio waves

Although today it is well recognized that radio signals travel in a wave-like manner, that fact was not always so clear. It was well known in the first half of the nineteenth century that wires carrying electrical currents produced an *induction field* surrounding the wire, which is capable of causing action over short distances. It was also known that this induction field is a magnetic field, and that knowledge formed the basis for electrical motors. In 1887, physicist Heinrich Hertz demonstrated that radio signals were *electromagnetic waves*, like light. Like the induction field, the electromagnetic wave is created by an electrical current moving in a conductor (e.g., a wire). Unlike the induction field, however, the *radiated field* leaves the conductor and propagates through space as an electromagnetic wave.

## 6 Radio-wave propagation

The *propagation* of waves is easily seen in the “water analogy.” Although not a perfect match to radio waves, it serves to illustrate the point. Figure 2-1 shows a body of water into which a ball is dropped (Fig. 2-1A). When the ball hits the water (Fig. 2-1B), it displaces water at its point of impact, and pushes a leading wall of water away from itself. The ball continues to sink and the wave propagates away from it until the energy is dissipated. Although Fig. 2-1 shows the action in only one dimension (a side view), the actual waves propagate outward in all directions, forming concentric circles when viewed from above.

The wave produced by a dropped ball is not continuous, but rather is *damped* (i.e., it will reduce in amplitude on successive crests until the energy is dissipated and the wave ceases to exist). But to make the analogy to radio waves more realistic, the wave must exist in a continuous fashion. Figure 2-2 shows how this is done: a ball is dipped up and down in a rhythmic, or cyclic manner, successively rein-



Notes: A Amplitude of leading wave  
B Corresponds to 1 cycle of oscillation

**2-1** A ball dropped into water generates a wavefront that spreads out from the point of original disturbances.

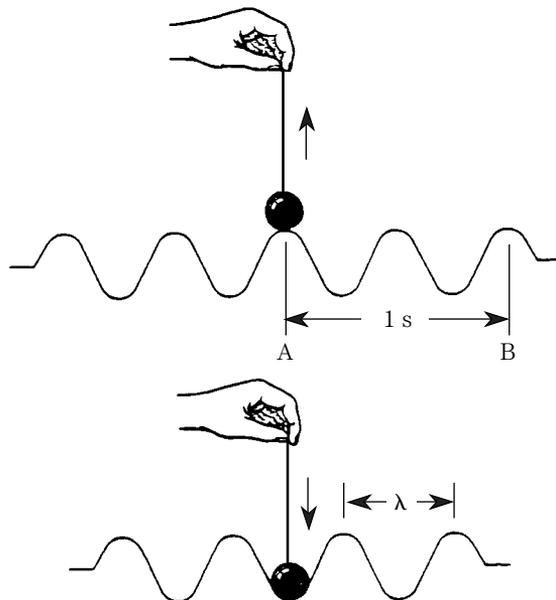
forcing new wave crests on each dip. The waves continue to radiate outward as long as the ball continues to oscillate up and down. The result is a *continuous wave train*.

There are two related properties of all waves that are important to radio waves as well: *frequency* ( $f$ ) and *wavelength* ( $\lambda$ ). The frequency is the number of oscillations (or cycles) per unit of time. In radio waves, the unit of time is the second, so frequency is an expression of the number of *cycles per second* (*cps*). If the period of time required for the leading wave to travel from point "A" to "B" is one second (1 s), and there are two complete wave cycles in that space, then the frequency of the wave created by the oscillating ball is 2 cps.

At one time, radio frequencies (along with the frequencies of other electrical and acoustical waves) were expressed in cps, but in honor of Heinrich Hertz, the unit was renamed the hertz (Hz) many years ago. Because the units are equal (1 Hz = 1 cps), the wave in Fig. 2-2 has a frequency of 2 Hz.

Because radio frequencies are so high, the frequency is usually expressed in kilohertz (kHz—1000s of Hz) and megahertz (MHz—1,000,000s of Hz). Thus, the frequency of a station operating in the middle of the AM broadcasting band can be properly expressed as 1,000,000 Hz, or 1000 kHz, or 1 MHz, all of which are equivalent to each other. Radio dials in North America are usually calibrated in kHz or MHz. In Europe and the rest of the world, on the other hand, it is not uncommon to find radio dials calibrated in meters, the unit of wavelength, as well as in frequency. In most equations used in radio antenna design, the proper units are hertz, kilohertz, and megahertz.

2-2 Bobbing ball on a string demonstrates continuous wave generation action.



The *wavelength* of any wave is the distance between like features on the waveform. In the case of Fig. 2-2, the wavelength ( $\lambda$ ) is the distance between successive positive peaks. We could also measure the same distance between successive negative peaks, or between any two similar features on successive waves. In radio work, the wavelength of the signal is expressed in *meters or its subunits*.

The wavelength is proportional to the reciprocal of the frequency. The wavelength of any wave is related to the frequency so that  $f\lambda = v$ , where  $f$  is the frequency in Hz,  $\lambda$  is the wavelength in meters, and  $v$  is the *velocity of propagation* in meters per second (m/s). Because radio waves propagate at the speed of light (which is also an electromagnetic wave), approximately 300,000,000 m/s in both free space and the earth's atmosphere, the lowercase letter  $c$  is used to represent velocity (rather than  $v$ ), so you can rewrite this expression in the form

$$f_{\text{Hz}} = \frac{c}{\lambda_{\text{meters}}} = \frac{300,000,000}{\lambda_{\text{meters}}} \quad [2.1]$$

These equations are sometimes abbreviated for use with the units kHz and MHz:

$$F_{\text{kHz}} = \frac{300,000}{\lambda_{\text{meters}}} \quad [2.2]$$

$$f_{\text{MHz}} = \frac{300}{\lambda_{\text{meters}}} \quad [2.3]$$

You can get an idea of the order of length of these waves by solving Eq. 2.3 for several different frequencies: 100 kHz, 1 MHz (in the AM broadcast band), 10 MHz (in the shortwave bands), and 1000 MHz (microwave bands). If you work the equations, then you will find that these wavelengths are 3000 m (100 kHz), 300 m (1 MHz), 30 m (10 MHz), and 0.3 m, or 30 cm (1000 MHz). You can see from these numbers why 1 MHz is in what is called the *medium-wave band*, 10 MHz is in the *short wave band*, and 1,000 MHz is in the *microwave* (“very small” wave) band. At 100 kHz, which is 0.1 MHz, the wavelength is 3000 m, so this frequency is in the *long-wave band*.

The place where the water analogy falls down most profoundly is in the nature of the medium of propagation. Water waves move by moving water molecules; water is said to be the *medium* in which the wave propagates. At one time, scientists could not conceive of the “action at a distance” provided by radio waves, so they invented a hypothetical medium called *ether* (or *aether*) for propagating electromagnetic waves (such as radio waves and light). It was not until the late nineteenth century that American physicists Michaelson and Morley proved that the ether does not exist. Nonetheless, radio enthusiasts still refer to the “stuff” out of which radio waves arrive as the “ether.” This terminology is merely an archaic, linguistic echo of the past.

## The electromagnetic field: a brief review

A great deal of heavily mathematical material can be presented about electromagnetic waves. Indeed, developing Maxwell's equations is a complete field of study for

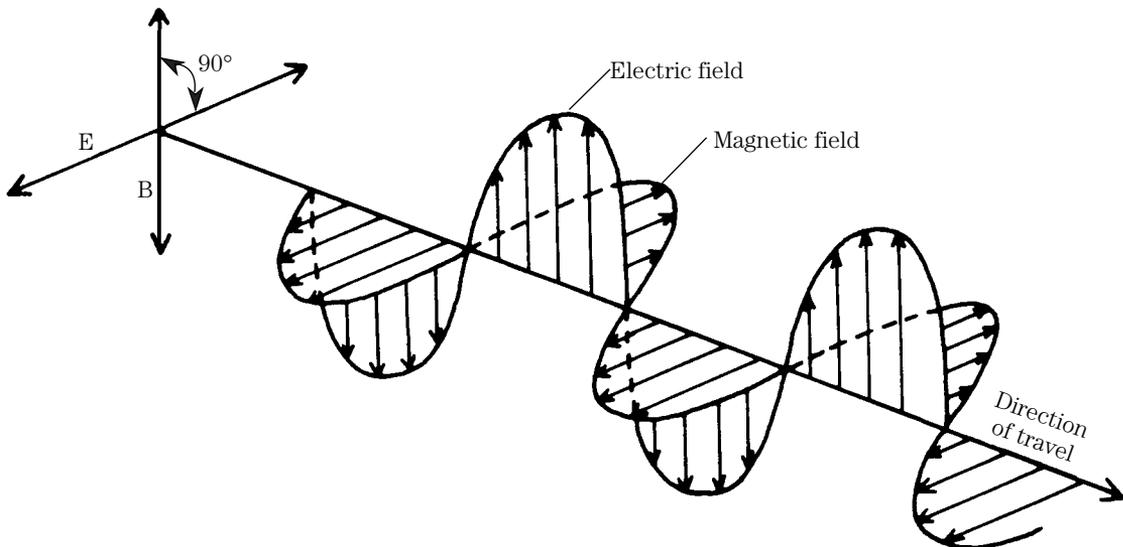
specialists. In this section, you will not use this rigorous treatment because you can refer to engineering textbooks for that depth of information. The purpose here is to present a descriptive approach that is designed to present you with a basic understanding of the phenomena. The approach here is similar to the learning of a “conversational” foreign language, rather than undertaking a deep study of its grammar, syntax, and context. For those whose professional work routinely involves electromagnetic waves, this treatment is hopelessly simplistic. For that I make no apology, because it serves a greater audience. The goal here is to make you more comfortable when thinking about the propagation of electromagnetic fields in the radio portion of the electromagnetic spectrum.

Radio signals are *transverse electromagnetic* (TEM or EM) waves exactly like light, infrared (IR), and ultraviolet (UV), except for frequency. Radio waves have much lower frequencies than light, IR, or UV, hence they have much longer wavelengths. The TEM (EM) wave consists of two mutually perpendicular oscillating fields (see Fig. 2-3) traveling together in phase. One of the fields is an *electric field* and the other is a *magnetic field*.

### Radio-wave intensity

The radio wave is *attenuated* (i.e. reduced in apparent power) as it propagates from the transmitter to the receiver. Although at some very high microwave frequencies, there is additional path loss as a result of the oxygen and water vapor content of the air. At other frequencies other losses exist. Radio waves at all frequencies suffer losses due to the *inverse square law*. Let’s take a look at that phenomenon.

The electric field vector falls off in direct proportion to the distance traveled. The E field is measured in terms of volts per meter (V/m), or in the subunits millivolts per meter (mV/m) or microvolts per meter ( $\mu\text{V/m}$ ). That is, if an E field of



2-3 Electromagnetic wave consists of right angle electric and magnetic fields.

## 10 Radio-wave propagation

10 V/m crosses your body from head to toe, and you are about 2 m tall, then an electrical voltage is generated of (2 m)×(10 V/m), or 20 V. The reduction of the E field is linearly related to distance (i.e., if the distance doubles the E-field voltage vector halves). Thus, a 100 mV/m E field that is measured 1 mi from the transmitter will be 50 mV/m at 2 mi.

The power in any electrical system is related to the voltage by the relationship

$$P = \frac{E^2}{R} \quad [2.4]$$

where

$P$  is the power in watts (W)

$R$  is the resistance in ohms ( $\Omega$ )

$E$  is the electrical potential in volts (V)

In the case of a radio wave, the  $R$  term is replaced with the impedance ( $Z$ ) of free space, which is on the order of  $377 \Omega$ . If the E field intensity is, for example, 10 V/m, then the *power density* of the signal is

$$P = \frac{(10 \text{ V/m})^2}{377 \Omega} = 0.265 \text{ W/m}^2 = 26.5 \text{ mW/m}^2 \quad [2.5]$$

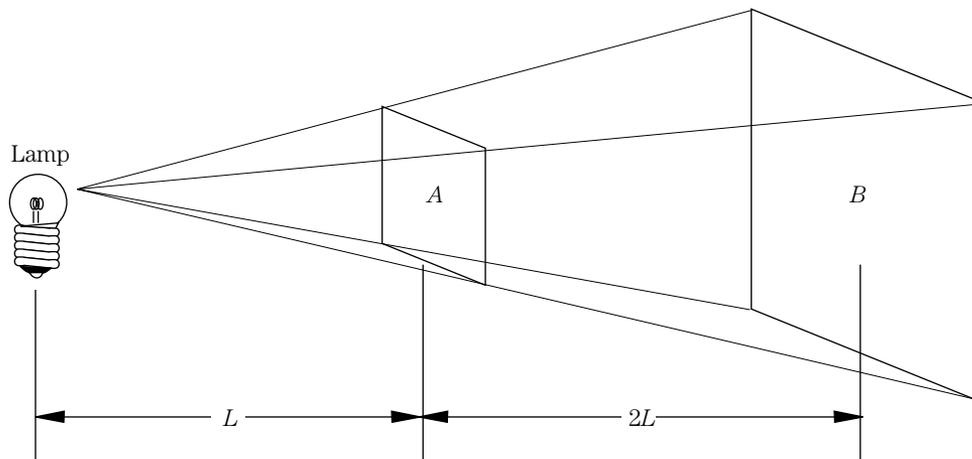
The power density, measured in watts per square meter ( $\text{W/m}^2$ ), or the subunits (e.g.,  $\text{mW/cm}^2$ ), falls off according to the square of the distance. This phenomenon is shown graphically in Fig. 2-4. Here, you can see a lamp emitting a light beam that falls on a surface ( $A$ ), at distance  $L$ , with a given intensity. At another surface ( $B$ ), that is  $2L$  from the source, the same amount of energy is distributed over an area ( $B$ ) that is twice as large as area  $A$ . Thus, the power density falls off according to  $1/d^2$ , where  $d$  is the difference in distance. This is called the *inverse square law*.

### Isotropic sources

In dealing with both antenna theory and radio wave propagation, a theoretical construct called an *isotropic source* is sometimes used for the sake of comparison, and for simpler arithmetic. You will see the isotropic model several places in this book. An isotropic source assumes that the radiator (i.e., an “antenna”) is a very tiny spherical source that radiates energy equally well in all directions. The radiation pattern is thus a sphere with the isotropic antenna at the center. Because a spherical source is uniform in all directions, and its geometry is easily determined mathematically, the signal intensities at all points can be calculated from basic geometric principles.

For the isotropic case, you can calculate the average power in the extended sphere from

$$P_{\text{av}} = \frac{P_t}{4\pi d^2} \quad [2.6]$$



**2-4** As a wave propagates, it spreads out according to the inverse square law, i.e., the area of *B* is 4 times that of *A* because it is twice as far from the source.

where

$P_{av}$  is the average power per unit area

$P_t$  is the total power

$d$  is the radius of the sphere in meters (i.e., the distance from the radiator to the point in question)

The *effective aperture* ( $A_e$ ) of the receiving antenna relates to its ability to collect power from the EM wave and deliver it to the load. Although typically smaller than the surface area in real antennas, for the theoretical isotropic case  $A_e = \lambda^2/4\pi$ . The power delivered to the load is

$$P_L = P_{av} A_e \quad [2.7]$$

By combining the two previous equations, the power delivered to a load at distance  $d$  is given by

$$P_L = \frac{P_t \lambda^2}{(4\pi)^2 d^2} \quad [2.8]$$

where

$P_L$  is the power to the load

$\lambda$  is the wavelength ( $c/F$ ) of the signal

From these expressions, there can then be derived an expression for ordinary path losses between an isotropic transmitter antenna and a receiver antenna:

$$L_{dB} = 10 \log \left( \frac{P_t}{P_L} \right) \quad [2.9]$$

## 12 Radio-wave propagation

or, by rearranging to account for individual terms:

$$L_{\text{dB}} = [20 \log d] + [20 \log F_{\text{MHz}}] + k \quad [2.10]$$

where

$L_{\text{dB}}$  is the path loss in decibels (dB)

$d$  is the path length

$F_{\text{MHz}}$  is frequency in megahertz (MHz)

$k$  is a constant that depends on the units of  $d$  as follows:

$k = 32.4$  if  $d$  in kilometers

$= 36.58$  if  $d$  in statute miles

$= 37.80$  if  $d$  in nautical miles

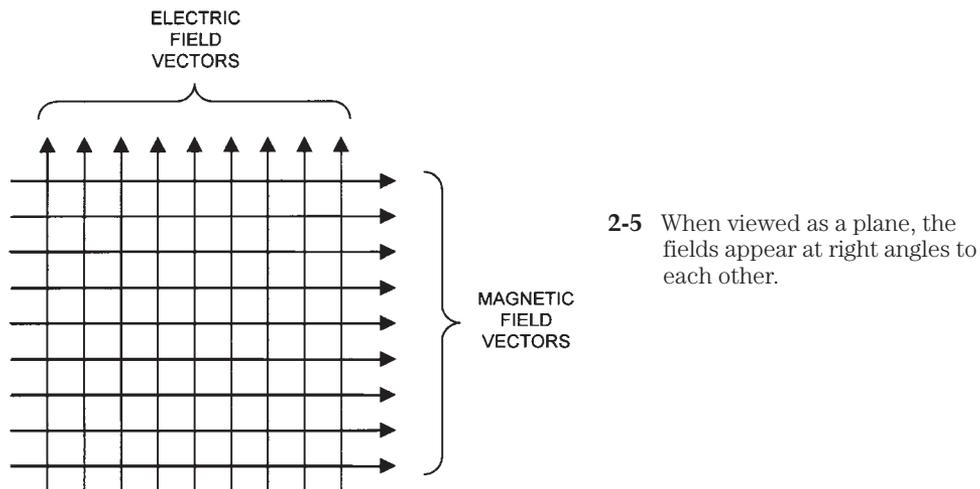
$= -37.87$  if  $d$  in feet

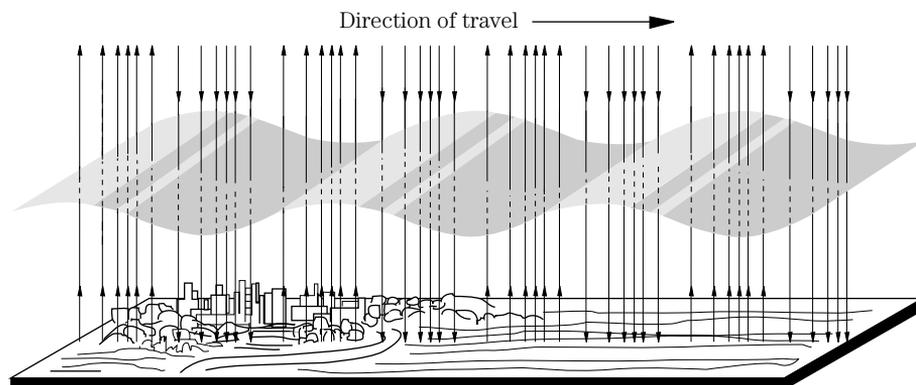
$= -27.55$  if  $d$  in meters

The radiated sphere of energy gets ever larger as the wave propagates away from the isotropic source. If, at a great distance from the center, you take a look at a small slice of the advancing wavefront you can assume that it is essentially a flat plane, as in Fig. 2-5. This situation is analogous to the apparent flatness of the prairie, even though the surface of the earth is nearly spherical. You would be able to “see” the electric and magnetic field vectors at right angles to each other in the flat plane wavefront.

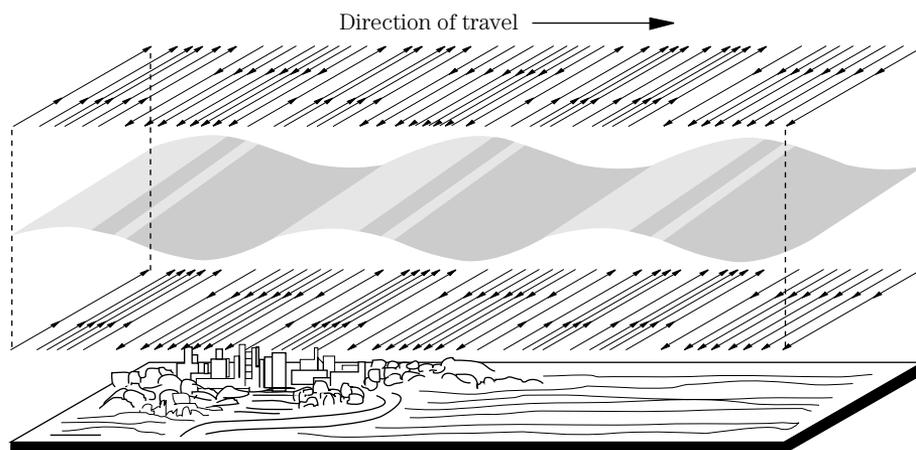
The *polarization* of an EM wave is, by definition, the *direction of the electric field*. Figure 2-6A shows vertical polarization, because the electric field is vertical with respect to the earth. If the fields were exchanged (as in Fig. 2-6B), then the EM wave would be horizontally polarized.

These designations are especially convenient because they also show the type of antenna used: vertical antennas produce vertically polarized signals, while hori-





**A**



**B**

**2-6** Wave polarization is determined by the direction of the electric field lines of force: (A) vertical polarized electromagnetic wave; (B) horizontally polarized wave.

horizontal antennas produce horizontally polarized signals. At least one text erroneously states that antennas will not pick up signals of the opposite polarity. Such is not the case, especially in the high-frequency (HF) and lower very high-frequency (VHF) regions. At VHF, ultrahigh frequency (UHF), and microwave frequencies a loss of approximately 20 to 30 dB can be observed due to cross-polarization.

An EM wave travels at the speed of light, designated by the letter  $c$ , which is about 300,000,000 m/s (or 186,000 mi/s). To put this velocity in perspective, a radio signal originating on the sun's surface would reach earth in about 8 minutes. A terrestrial radio signal can travel around the earth seven times in one second. The velocity of the

wave slows in dense media, but in air the speed is so close to the “free-space” value of  $c$ , that the same figures are used for both air and the near vacuum of outer space in practical problems. In pure water, which is much denser than air, the speed of radio signals is about one-ninth that of the free-space speed. This same phenomenon shows up in practical work in the form of the *velocity factor* ( $V$ ) of transmission lines. In foam dielectric coaxial cable, for example, the value of  $V$  is 0.80, which means that the signal propagates along the line at a speed of  $0.80c$ , or 80 percent of the speed of light.

This coverage of radio propagation considers the EM wave as a very narrow “ray” or “pencil beam” that does not diverge as it travels. That is, the ray remains the same width all along its path. This convention makes it easy to use ray tracing diagrams. Keep in mind, however, that the real situation, even when narrow-beamwidth microwave signals are used, is much more complicated. Real signals, after all, are sloppier than textbook examples: they are neither infinitesimally thin, nor nondivergent.

### **The earth’s atmosphere**

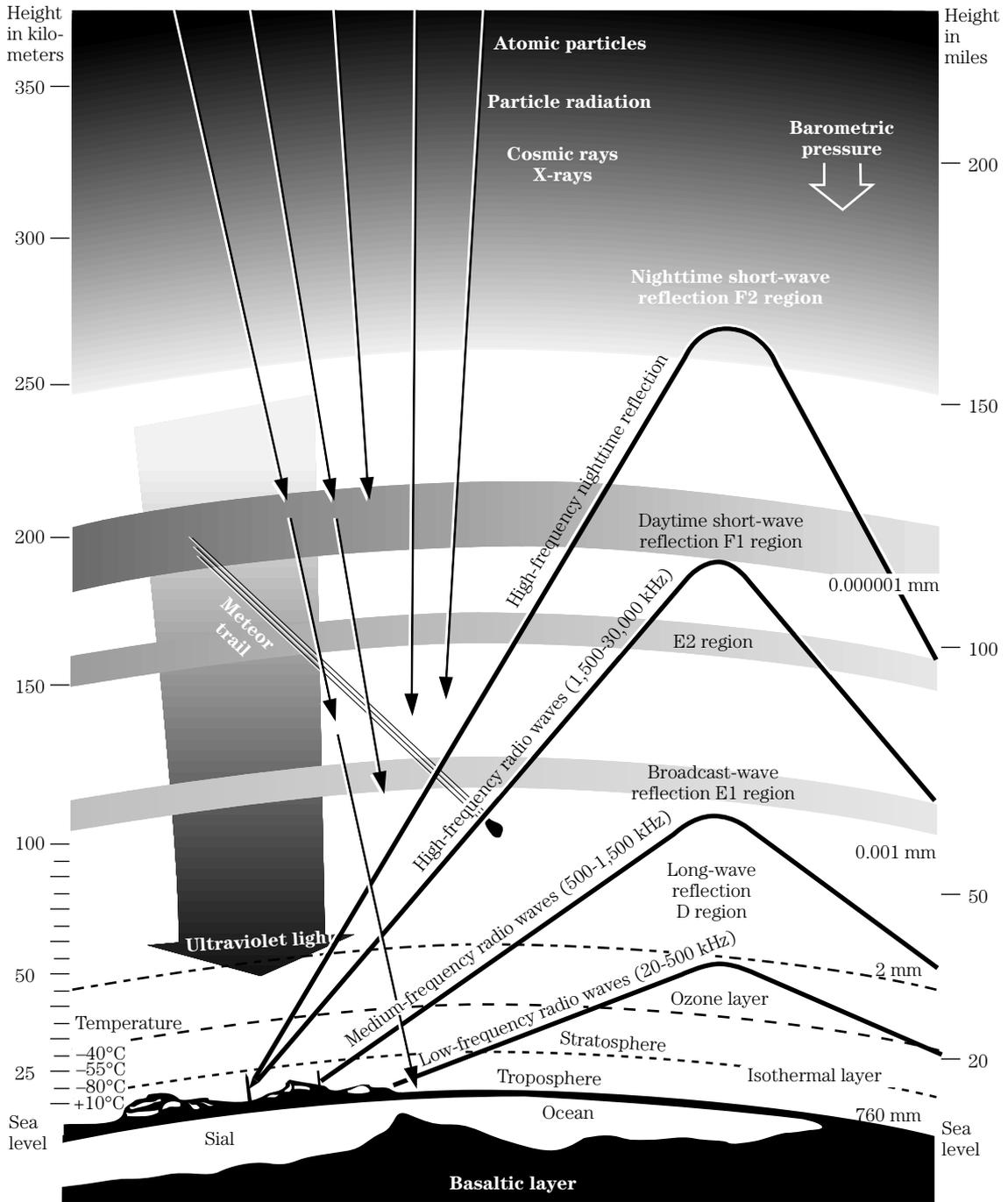
The electromagnetic waves do not need an atmosphere in order to propagate, as you will undoubtedly realize from the fact that space vehicles can transmit radio signals back to earth in a near vacuum. But when a radio wave does propagate in the earth’s atmosphere, it interacts with the atmosphere, and its path of propagation is altered. A number of factors affect the interaction, but it is possible to break the atmosphere into several different categories according to their respective effects on radio signals.

The atmosphere, which consists largely of oxygen ( $O_2$ ) and nitrogen (N) gases, is broken into three major zones: *troposphere*, *stratosphere*, and *ionosphere* (Fig. 2-7). The boundaries between these regions are not very well defined, and change both diurnally (over the course of a day) and seasonally.

The troposphere occupies the space between the earth’s surface and an altitude of 6 to 11 km (4 to 7 mi). The temperature of the air in the troposphere varies with altitude, becoming considerably lower at greater altitude compared with ground temperature. For example, a  $+10^\circ\text{C}$  surface temperature could reduce to  $-55^\circ\text{C}$  at the upper edges of the troposphere.

The stratosphere begins at the upper boundary of the troposphere (6 to 11 km), and extends up to the ionosphere ( $\approx 50$  km). The stratosphere is called an *isothermal region* because the temperature in this region is somewhat constant, despite altitude changes.

The ionosphere begins at an altitude of about 50 km (31 mi) and extends up to approximately 300 km (186 mi). The ionosphere is a region of very thin atmosphere. Cosmic rays, electromagnetic radiation of various types (including ultraviolet light from the sun), and atomic particle radiation from space (most of these from the sun also), have sufficient energy to strip electrons away from the gas molecules of the atmosphere. These freed electrons are called *negative ions*, while the  $O_2$  and N molecules that lost electrons are called *positive ions*. Because the density of the air is so low at those altitudes, the ions can travel long distances before neutralizing each other by recombining. Radio propagation on some bands varies markedly between daytime and nighttime because the sun keeps the level of ionization high during



2-7 Radio propagation in the ionosphere is affected by a number of different physical factors: cosmic rays, atomic particles, solar radiation.

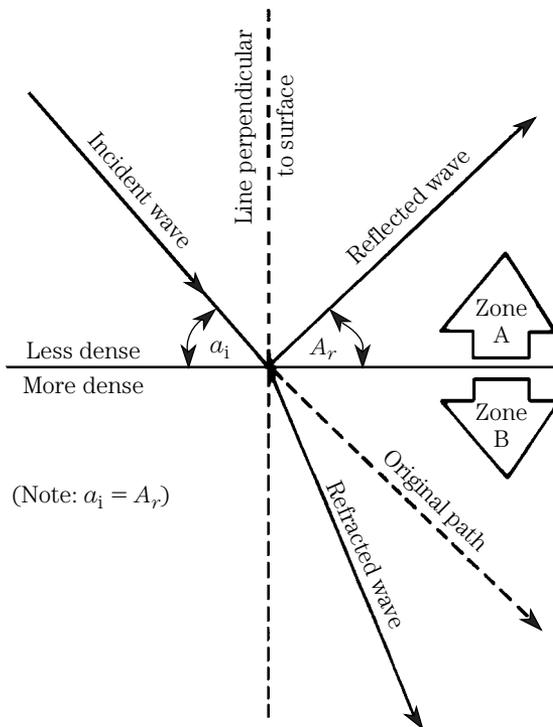
daylight hours, but the ionization begins to fall off rapidly after sunset, altering the radio propagation characteristics after dark. The ionization does not occur at lower altitudes because the air density is such that the positive and negative ions are numerous and close together, so recombination occurs rapidly.

## EM wave propagation phenomena

Because EM waves are waves, they behave in a wave-like manner. Figure 2-8 illustrates some of the wave behavior phenomena associated with light and radio waves: *reflection*, *refraction*, and *diffraction*. All three play roles in radio propagation.

Reflection and refraction are shown in Fig. 2-8A. Reflection occurs when a wave strikes a denser reflective medium, as when a light wave strikes a glass mirror. The incident wave (shown as a single ray) strikes the interface between less dense and more dense mediums at a certain *angle of incidence* ( $a_i$ ), and is reflected at exactly the same angle (now called the *angle of reflection* ( $a_r$ )). Because these angles are equal, a reflected radio signal can often be traced back to its origin.

Refraction occurs when the incident wave enters a region of different density, and thereby undergoes both a velocity change and a directional change. The amount and direction of the change are determined by the ratio of the densities between the two media. If Zone B is much different from Zone A, then bending is great. In radio



2-8A Reflection and refraction phenomena.

systems, the two media might be different layers of air with different densities. It is possible for both reflection and refraction to occur in the same system. Indeed, more than one form of refraction might be present. These topics will be covered in greater depth shortly.

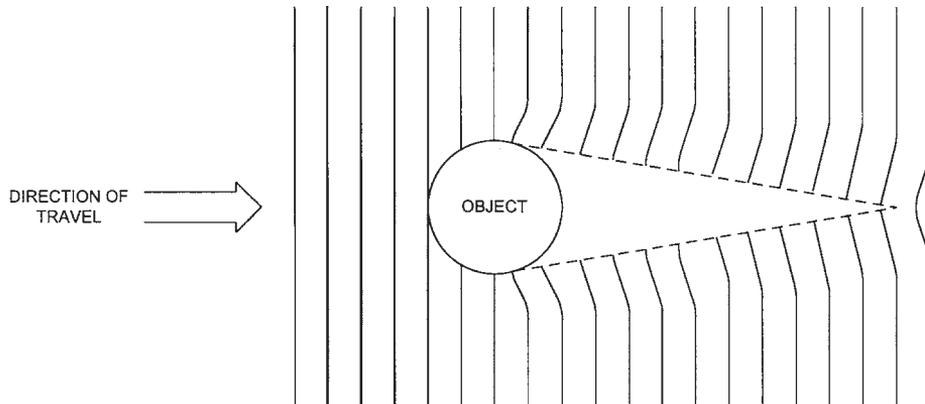
Diffraction is shown in Fig. 2-8B. In this case, an advancing wavefront encounters an opaque object (e.g., a steel building). The shadow zone behind the building is not simply perpendicular to the wave, but takes on a cone shape as waves bend around the object. The “umbra region” (or diffraction zone) between the shadow zone (“cone of silence”) and the direct propagation zone is a region of weak (but not zero) signal strength. In practical situations, signal strength in the cone of silence rarely reaches zero. A certain amount of reflected signals scattered from other sources will fill in the shadow a little bit. The degree of diffraction effect seen in any given case is a function of the wavelength of the signal, the size of the object, and its electromagnetic properties.

### Propagation paths

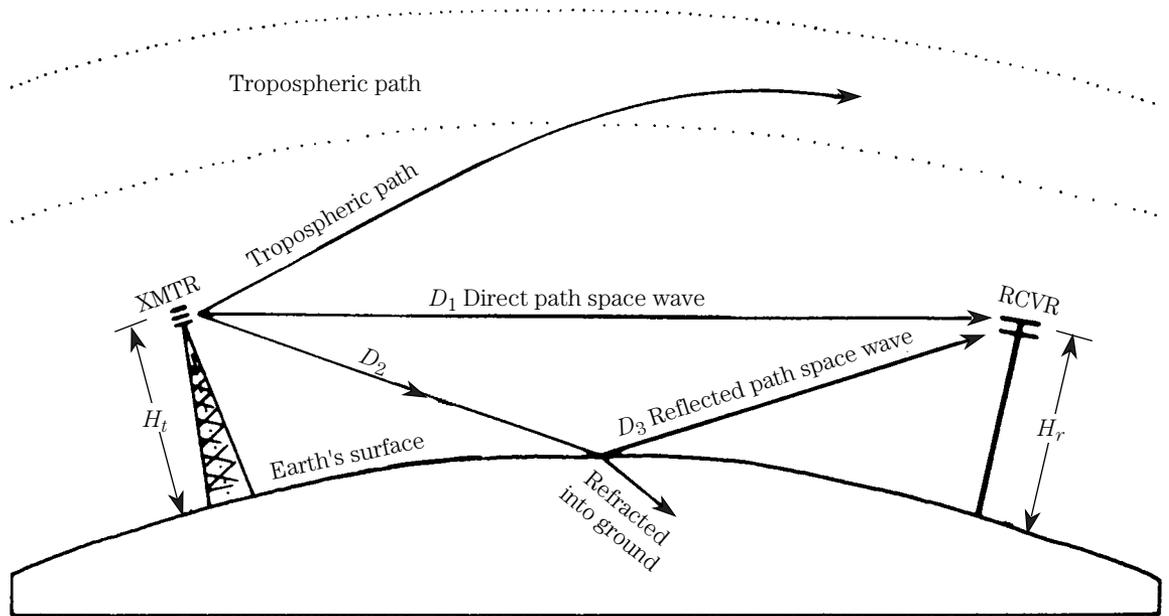
There are four major propagation paths: *surface wave*, *space wave*, *tropospheric*, and *ionospheric*. The ionospheric path is important to medium-wave (MW) and HF propagation, but is not important to VHF, UHF, or microwave propagation. The space wave and surface wave are both *ground waves*, but they behave differently enough to warrant separate consideration. The surface wave travels in direct contact with the earth’s surface and it suffers a severe frequency-dependent attenuation caused by absorption into the ground.

The space wave is also a ground-wave phenomenon, but it is radiated from an antenna many wavelengths above the surface. No part of the space wave normally travels in contact with the surface; VHF, UHF, and microwave signals are usually space waves. There are, however, two components of the space wave in many cases: *direct* and *reflected* (see Fig. 2-9).

The tropospheric wave is lumped with the direct space wave in some texts, but it has properties that actually make it different in practical situations. The troposphere



2-8B Diffraction phenomena.



2-9 Space wave propagation.

is the region of earth's atmosphere between the surface and the stratosphere, or about 4 to 7 mi above the surface. Thus, most forms of ground wave propagate in the troposphere. But because certain propagation phenomena (caused mostly by weather conditions) only occur at higher altitudes, tropospheric propagation should be differentiated from other forms of ground wave.

The ionosphere is the region of earth's atmosphere that is above the stratosphere. The peculiar feature of the ionosphere is that molecules of air gas ( $O_2$  and  $N$ ) can be ionized by stripping away electrons under the influence of solar radiation and certain other sources of energy. In the ionosphere, the air density is so low that ions can travel relatively long distances before recombining with oppositely charged ions to form electrically neutral atoms. As a result, the ionosphere remains ionized for long periods of the day—even after sunset. At lower altitudes, however, air density is greater and recombination thus occurs rapidly. At those altitudes, solar ionization diminishes to nearly zero immediately after sunset, or never achieves any significant levels even at local noon.

Ionization and recombination phenomena in the ionosphere add to the noise level experienced at VHF, UHF, and microwave frequencies. The properties of the ionosphere are important to microwave technology because of the noise contribution. In satellite communications, there are some additional transionospheric effects.

### Ground-wave propagation

The *ground wave*, naturally enough, travels along the ground, or at least in close proximity to it. There are three basic forms of ground wave: *space wave*, *surface*

*wave*, and *tropospheric wave*. The space wave does not actually touch the ground. As a result, space wave attenuation with distance in clear weather is about the same as in free space (except above about 10 GHz, where H<sub>2</sub>O and O<sub>2</sub> absorption increases dramatically). Of course, above the VHF region, weather conditions add attenuation not found in outer space.

The surface wave is subject to the same attenuation factors as the space wave, but in addition it also suffers ground losses. These losses are caused by ohmic resistive losses in the conductive earth. In other words, the signal heats up the ground. Surface wave attenuation is a function of frequency, and it increases rapidly as frequency increases. Both of these forms of ground-wave communications are affected by the following factors: wavelength, height of both the receiving and transmitting antennas, distance between the antennas, and the terrain and weather along the transmission path. Figure 2-10 is a nomograph that can be used to calculate the line of sight distances in miles from a knowledge of the receiver and transmitter antenna heights. Similarly, Figs. 2-11A and 2-11B show power attenuation with frequency and distance (Fig. 2-11A) and power attenuation in terms of field intensity (Fig. 2-11B).

Ground-wave communications also suffer another difficulty, especially at VHF, UHF, and microwave frequencies. The space wave is like a surface wave, but it is radiated many wavelengths above the surface. It is made up of two components (see Fig. 2-9 again): the *direct* and *reflected* waves. If both of these components arrive at the receiving antenna, they will add algebraically to either increase or decrease signal strength. There is always a phase shift between the two components because the two signal paths have different lengths (i.e.,  $D_1$  is less than  $D_2 + D_3$ ). In addition, there may possibly be a 180° ( $\pi$  radians) phase reversal at the point of reflection (especially if the incident signal is horizontally polarized), as in Fig. 2-12. The following general rules apply in these situations:

- A phase shift of an odd number of half wavelengths causes the components to add, increasing signal strength (*constructive interference*).
- A phase shift of an even number of half wavelengths causes the components to subtract (Fig. 2-12), thus reducing signal strength (*destructive interference*).
- Phase shifts other than half wavelength add or subtract according to relative polarity and amplitude.

You can characterize the loss of signal over path  $D_1$  with a parametric term  $n$  that is defined as follows:

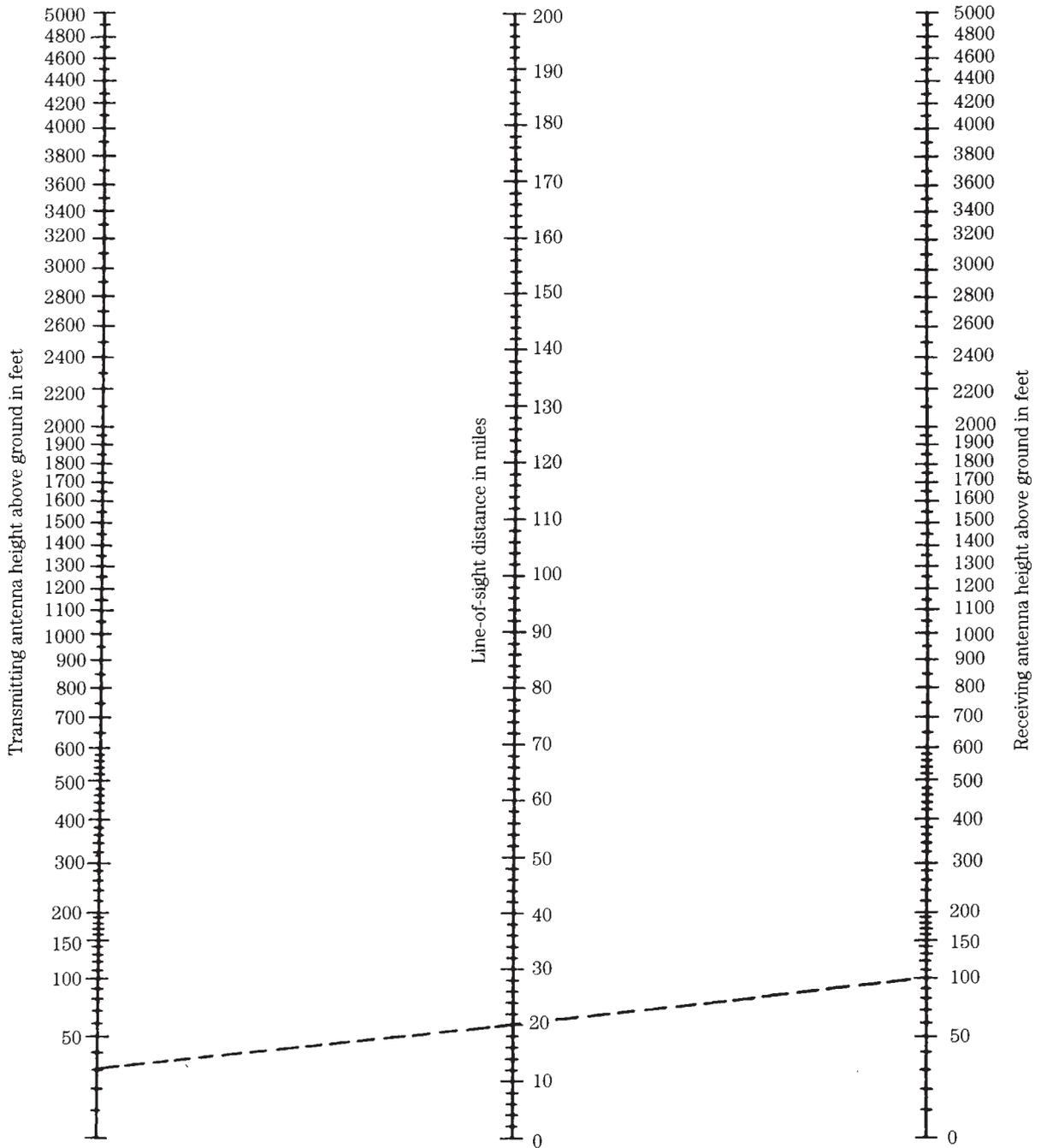
$$n = \frac{S_r}{S_f} \quad [2.11]$$

where

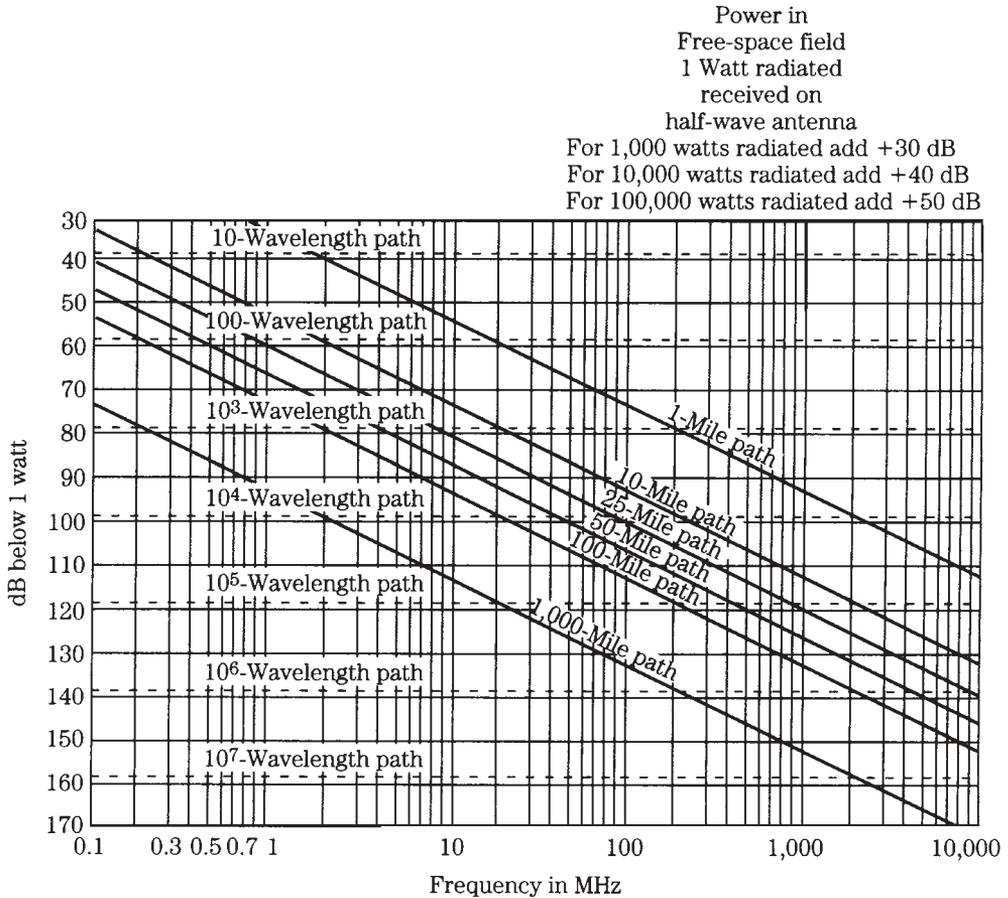
$n$  is the signal loss coefficient

$S_r$  is the signal level at the receiver in the presence of a ground reflection component

$S_f$  is the free-space signal strength over path  $D_1$  if no reflection took place.



**2-10** Nomograph showing the line-of-sight transmission distance as a function of receiving and transmitting antenna heights.



2-11A Power in a free-space field (normalized to 1 W).

You can calculate as follows:

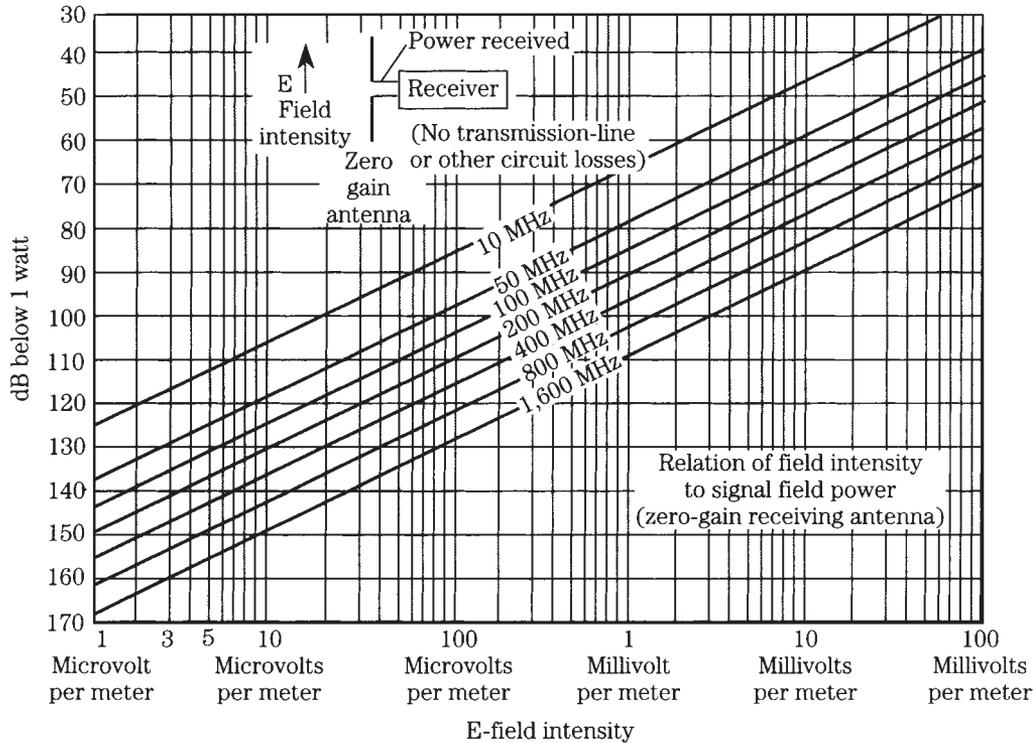
$$n^2 = 4 \sin^2 \left( \frac{2\pi h_t h_r}{\lambda D_1} \right) \tag{2.12}$$

or

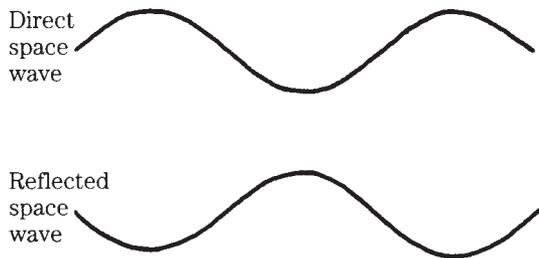
$$n = 2 \sin \left( \frac{2\pi h_t h_r}{\lambda D_1} \right) \tag{2.13}$$

The reflected signal contains both amplitude change and phase change. The phase change is typically  $\pi$  radians (180°). The amplitude change is a function of frequency and the nature of the reflecting surface. The *reflection coefficient* can be characterized as

$$\gamma = pe^{j\phi} \tag{2.14}$$



2-11B Relation of field strength to signal field power.



2-12 Out-of-phase waves cancel.

where

- $\gamma$  is the reflection coefficient
- $p$  is the amplitude change
- $\phi$  is the phase change
- $j$  is the imaginary operator ( $\sqrt{-1}$ )

For smooth, high-reflectivity surfaces and a horizontally polarized microwave signal that has a shallow angle of incidence, the value of the reflection coefficient is close to  $-1$ .

The phase change of the reflected signal at the receiving antenna is at least  $\pi$  radians because of the reflection. Added to this change is an additional phase shift that is a function of the difference in path lengths. This phase shift can be expressed in terms of the two antenna heights and path length.

$$s = \pi + \left( \frac{4\pi h_t h_r}{\lambda D_1} \right) \quad [2.15]$$

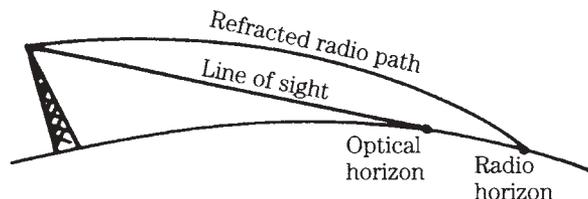
A category of reception problems called *multipath phenomena* exists because of interference between the direct and reflected components of the space wave. The form of multipath phenomena that is, perhaps, most familiar to many readers is ghosting in off-the-air television reception. Some multipath events are transitory in nature (as when an aircraft flies through the transmission path), while others are permanent (as when a large building, or hill, reflects the signal). In mobile communications, multipath phenomena are responsible for reception dead zones and “picket fencing.” A *dead zone* exists when destructive interference between direct and reflected (or multiply reflected) waves drastically reduces signal strengths. This problem is most often noticed at VHF and above when the vehicle is stopped; and the solution is to move the antenna one-half wavelength (which at VHF and UHF is a matter of a few inches). *Picket fencing* occurs as a mobile unit moves through successive dead zones and signal enhancement (or normal) zones, and it sounds like a series of short noise bursts.

At VHF, UHF, and microwave frequencies, the space wave is limited to so-called line-of-sight distances. The horizon is theoretically the limit of communications distance, but the radio horizon is actually about 15 percent farther than the optical horizon (Fig. 2-13). This phenomenon is caused by refractive bending in the atmosphere around the curvature of the earth, and it makes the geometry of the situation look as if the earth’s radius is  $\frac{4}{3}$  the actual radius.

The refraction phenomenon occurs at VHF through microwave frequencies, but not in the visible light spectrum, because water and atmospheric pressure (which relates to the effects of atmospheric gases on microwave signals) become important contributors to the phenomenon. The *K factor* expresses the degree of curvature along any given path, while the *index of refraction*  $n$  measures the differential properties between adjacent zones of air.

The *K factor*, also called the *effective earth’s radius factor*, is defined according to the relationship of two hypothetical spheres, both centered at the exact center of

2-13 Phenomenon by which “greater than line of sight” communications occurs.



the earth. The first sphere is the earth's surface, which has a radius  $r_o$  (3440 nmi or 6370 km). The second sphere is larger than the first by the curvature of the signal "ray path," and has a radius  $r$ . The value of  $K$  is approximately

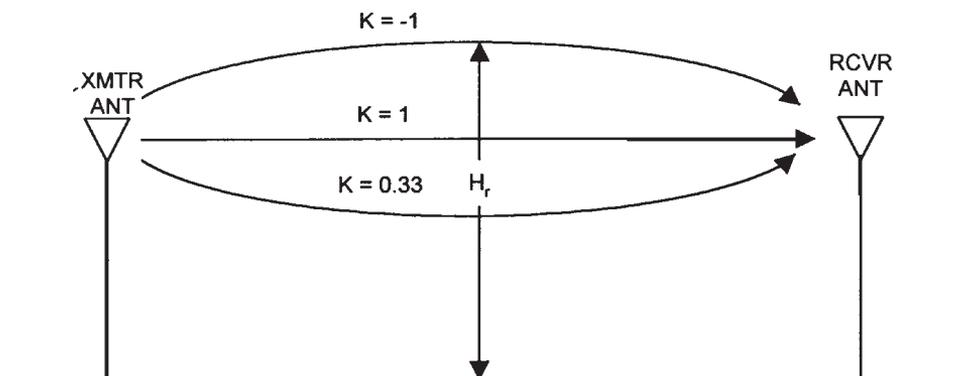
$$K = \frac{r}{r_o} \quad [2.16]$$

A value of  $K = 1$  indicates a straight path (Fig. 2-14); a value of  $K > 1$  indicates a positively curved path (refraction); and a value of  $K < 1$  indicates a negatively curved path (subrefraction). The actual value of  $K$  varies with local weather conditions, so one can expect variation not only between locations, but also seasonally. In the arctic regions,  $K$  varies approximately over the range 1.2 to 1.34. In the "lower 48" states of the United States,  $K$  varies from 1.25 to 1.9 during the summer months (especially in the south and southeast), and from 1.25 to 1.45 in the winter months. The index of refraction  $n$  can be defined in two ways, depending on the situation. When a signal passes across boundaries between adjacent regions of distinctly different properties (as occurs in temperature inversions, etc.), the index of refraction is the ratio of the signal velocities in the two regions. In a homogeneous region  $n$  can be expressed as the ratio of the free-space velocity  $c$  to the actual velocity in the atmosphere  $V$ :

$$n = \frac{c}{V} \quad [2.17]$$

At the surface, near sea level, under standard temperature and pressure conditions, the value of  $n$  is approximately 1.0003, and in homogeneous atmospheres it will decrease by  $4 \times 10^{-8}$  per mile of altitude. The units of  $n$  are a bit cumbersome in equations, so the UHF/microwave communities tend to use a derivative parameter,  $N$ , called the *refractivity of the atmosphere*:

$$N = (n - 1) \times 10^6 \quad [2.18]$$



2-14 Refraction changes path length.

The value of  $N$  tends to vary from about 280 to 320, and both  $n$  and  $N$  vary with altitude. In nonhomogeneous atmospheres (the usual case), these parameters will vary approximately linearly for several tenths of a kilometer. All but a few microwave relay systems can assume an approximately linear reduction of  $n$  and  $N$  with increasing altitude, although airborne radios and radars cannot. There are two methods for calculating  $N$ :

$$N = \left( \frac{77.6}{T} \right) \left( P + \frac{4810 e_s H_{\text{rel}}}{T} \right) \quad [2.19]$$

and

$$N = \frac{77.6}{T} \left( \frac{3.73 \times 10^5 e_s}{T^2} \right) \quad [2.20]$$

where

$P$  is the atmospheric pressure in millibars (1 torr = 1.3332 mbar)

$T$  is temperature in kelvins

$e_s$  is saturation vapor pressure of atmospheric water in millibars

$H_{\text{rel}}$  is the relative humidity expressed as a decimal fraction (rather than a percentage)

Ray path curvature ( $K$ ) can be expressed as a function of either  $n$  or  $N$ , provided that the assumption of a linear gradient  $d_n/d_h$  holds true:

$$K = \frac{1}{\left( 1 + \frac{r_o d_n}{d_h} \right)} \quad [2.21]$$

or

$$K = \frac{1}{\left( 1 + \frac{d_N/d_h}{157} \right)} \quad [2.22]$$

For the near-surface region, where  $d_n/d_h$  varies at about  $3.9 \times 10^{-8}$  m, the value of  $K$  is 1.33. For most terrestrial microwave paths, this value ( $K = \frac{4}{3} = 1.33$ ) is called a *standard refraction*, and is used in calculations *in the absence of additional data*. For regions above the linear zone close to the surface, you can use another expression of refractivity:

$$N_a = N_s e^{-C_e(h_r - h_t)} \quad [2.23]$$

where

$N_1$  is the refractivity at 1 km altitude

$h_r$  is the height of the receiver antenna

$h_t$  is the height of the transmit antenna

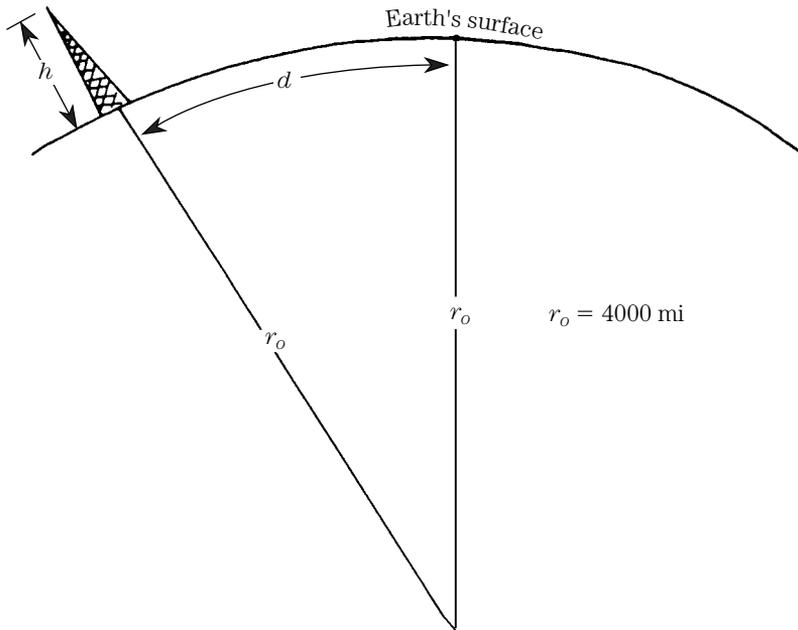
26 *Radio-wave propagation*

$$C_e = L_n(N_s/N_1)$$

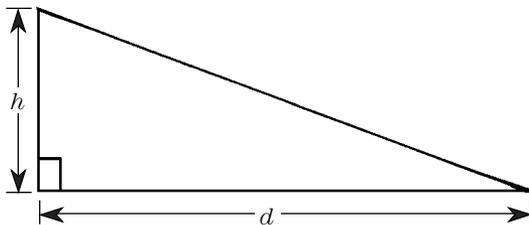
$N_a$  is the refractivity at altitude

$N_s$  is the refractivity at the earth's surface

For simple models close to the surface, you can use the geometry shown in Fig. 2-15. Distance  $d$  is a curved path along the surface of the earth. But because the earth's radius  $r_o$  is about 4000 statute miles, and it is thus very much larger than a practical antenna height  $h$ , you can simplify the model in Fig. 2-15A to that in Fig. 2-15B. The underlying assumption is that the earth has a radio radius equal to about  $\frac{4}{3}$  ( $K = 1.33$ ) of the actual physical radius of its surface.



2-15A Geometry for calculating radio distances.



2-15B Simplified geometry.

The value of distance  $d$  is found from the expression

$$d = \sqrt{2r_e h_{ft}} \quad [2.24]$$

where

$d$  is the distance to the radio horizon in statute miles

$r_e$  is the normalized radius of the earth

$h_{ft}$  is the antenna height in feet

Accounting for all constant factors, the expression reduces to

$$d = 1.414 \sqrt{h_{ft}} \quad [2.25]$$

all factors being the same as defined above.

**Example 2-1** A radio tower has a UHF radio antenna that is mounted 150 feet above the surface of the earth. Calculate the radio horizon (in statute miles) for this system.

**Solution:**

$$\begin{aligned} d &= 1.414 (h_{ft})^{1/2} \\ &= (1.414)(150 \text{ ft})^{1/2} \\ &= (1.414)(12.25) \\ &= 17.32 \text{ mi} \end{aligned}$$

For other units of measurement:

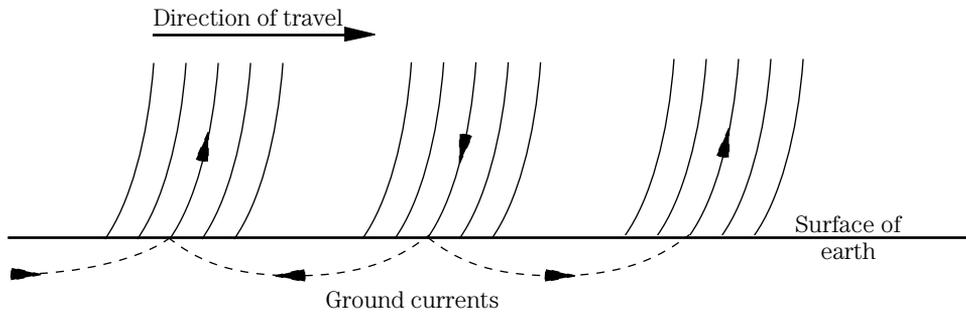
$$d_{\text{nmi}} = 1.23 \sqrt{h_{ft}} \quad \text{nmi} \quad [2.26]$$

and

$$d_{\text{km}} = 1.30 \sqrt{h_{ft}} \quad \text{km} \quad [2.27]$$

**Surface-wave communications** The surface wave travels in direct contact with the earth's surface, and it suffers a severe frequency-dependent attenuation because of absorption by the ground.

The surface wave extends to considerable heights above the ground level, although its intensity drops off rapidly at the upper end. The surface wave is subject to the same attenuation factors as the space wave, but in addition, it also suffers ground losses. These losses are caused by ohmic resistive losses in the conductive earth, and to the dielectric properties of the earth. In other words, the signal heats up the ground. Horizontally polarized waves are not often used for surface wave communications because the earth tends to short-circuit the E-field component. On vertically polarized waves, however, the earth offers electrical resistance to the E-field and returns currents to following waves (Fig. 2-16). The conductivity of the soil determines how much energy is returned. Table 2-1 shows the typical conductivity values for several different forms of soil.



**2-16** Distortion of vertically polarized electric field by lossy ground resistance.

**Table 2-1. Sample soil conductivity values**

Type of soil	Dielectric constant	Conductivity (siemens/meter)	Relative quality
Salt water	81	5	Best
Fresh water	80	0.001	Very poor
Pastoral hills	14–20	0.03–0.01	Very good
Marshy, wooded	12	0.0075	Average/poor
Rocky hills	12–14	10	Poor
Sandy	10	0.002	Poor
Cities	3–5	0.001	Very poor

The wavefront of a surface wave is tilted because of the losses in the ground that tend to retard the wavefront's bottom (also in Fig. 2-16). The tilt angle is a function of the frequency, as shown in Table 2-2.

**Table 2-2. Tilt angle as a function of frequency**

Frequency (kHz)	Tilt angle ratio	Earth/seawater (degrees)
20	2.07	4.3/2.08
200	104	13.4/0.13
2,000	64	32.3/0.5
20,000	25	35/1.38

Surface-wave attenuation is a function of frequency, and increases rapidly as frequency increases. For both forms of the ground wave, reception is affected by these factors: *wavelength*, *height* of both the receiving and transmitting antennas, *distance* between antennas, and both *terrain* and *weather* along the transmission

path. In addition, the surface wave is affected by the ground losses in Table 2-1. Because of the ground loss effects, the surface wave is attenuated at a much faster rate than the inverse square law.

**Ground-wave propagation frequency effects** The frequency of a radio signal in large measure determines its surface-wave behavior. In the very low-frequency (VLF) band (<300 kHz), ground losses are small for vertically polarized signals, so medium-distance communications (up to several hundred miles) are possible. In the medium-wave band (300 to 3000 kHz, including the AM broadcast band), distances of 1000 mi are possible with regularity—especially at night. In the high-frequency (HF) band, ground losses are more considerable, so the surface-wave distance reduces drastically. It is possible, in the upper end of the HF band (3000 to 30,000 kHz) for surface-wave signals to die out within a few dozen miles. This phenomenon is often seen in the 15- and 10-m amateur radio bands, as well as the 11-m (27-MHz) Citizen’s Band. Stations only 20 mi apart cannot communicate, but both can talk to a third station across the continent via ionospheric skip. Thus, the two stations close together must have a station more than 2000 mi away relay messages between them.

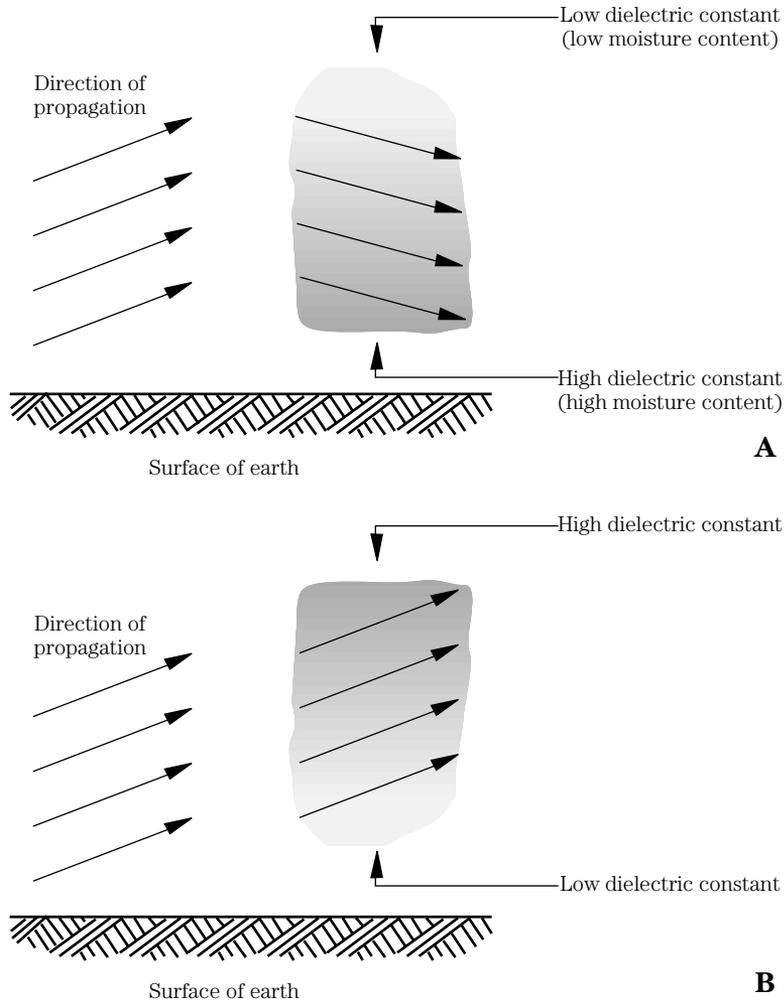
**Tropospheric propagation** The troposphere is the portion of the atmosphere between the surface of the earth and the stratosphere (or about 4 to 7 mi above the surface). Some older texts group tropospheric propagation with ground-wave propagation, but modern practice requires separate treatment. The older grouping overlooks certain common propagation phenomena that simply don’t happen with space or surface waves.

Refraction is the mechanism for most tropospheric propagation phenomena. The dielectric properties of the air, which are set mostly by the moisture content (Fig. 2-17), are a primary factor in tropospheric refraction. Recall that refraction occurs in both light or radio-wave systems when the wave passes between mediums of differing density. Under that situation, the wave path will bend an amount proportional to the difference in density.

Two general situations are typically found—especially at UHF and microwave frequencies. First, because air density normally decreases with altitude, the top of a beam of radio waves typically travels slightly faster than the lower portion of the beam. As a result, those signals refract a small amount. Such propagation provides slightly longer surface distances than are normally expected from calculating the distance to the radio horizon. This phenomenon is called *simple refraction*, and was discussed in a preceding section.

A special case of refraction called *superrefraction* occurs in areas of the world where warmed land air goes out over a cooler sea (Fig. 2-18). Examples of such areas have deserts that are adjacent to a large body of water: the Gulf of Aden, the southern Mediterranean, and the Pacific Ocean off the coast of Baja California are examples. VHF/UHF/microwave communications up to 200 mi are reported in such areas.

The second form of refraction is weather-related. Called *ducting*, this form of propagation (Fig. 2-19) is actually a special case of superrefraction. Evaporation of sea water causes temperature inversion regions to form in the atmosphere—that is, layered air masses in which the air temperature is greater than in the layers below it. (*Note:* air temperature normally decreases with altitude, but at the boundary with an

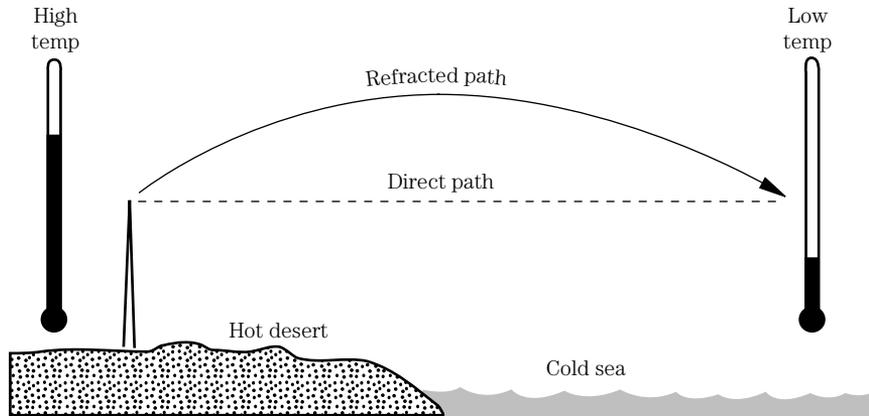


2-17 Refraction in the troposphere.

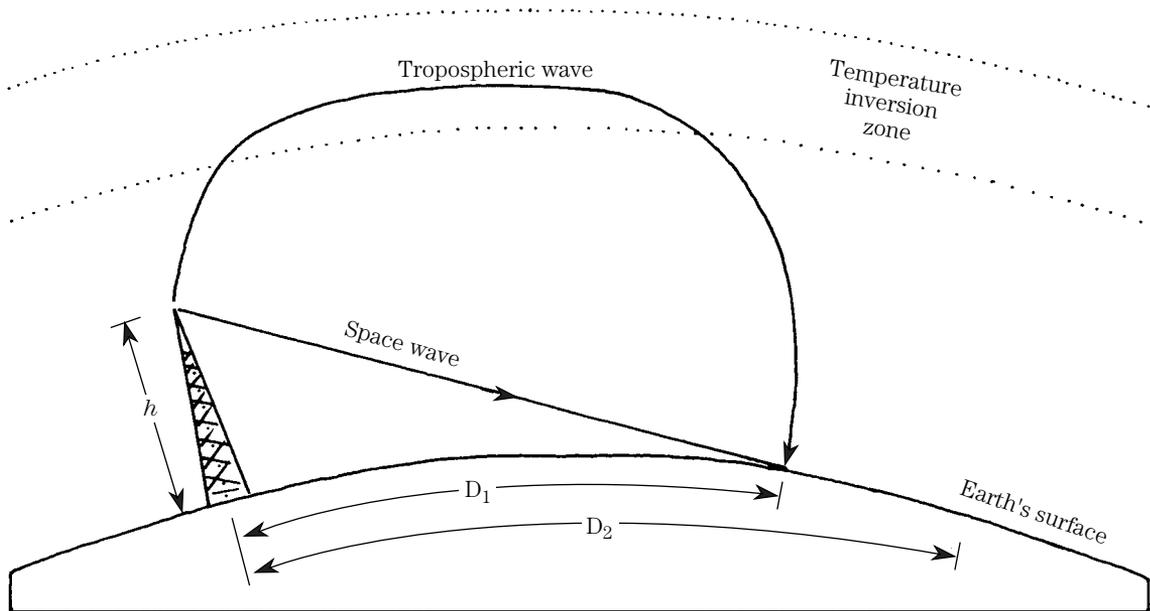
inversion region, it increases.) The inversion layer forms a “duct” that acts like a waveguide. In Fig. 2-19, the distance  $D_1$  is the normal “radio horizon” distance, and  $D_2$  is the distance over which duct communications can occur.

Ducting allows long-distance communications from lower VHF through microwave frequencies, with 50 MHz being a lower practical limit and 10 GHz being an ill-defined upper limit. Airborne operators of radio, radar, and other electronic equipment can sometimes note ducting at even higher microwave frequencies.

Antenna placement is critical for ducting propagation. Both the receiving and transmitting antennas must be either (1) inside the duct physically (as in airborne cases) or (2) able to propagate at an angle such that the signal gets trapped inside the duct. The latter is a function of antenna radiation angle. Distances up to 2500 mi



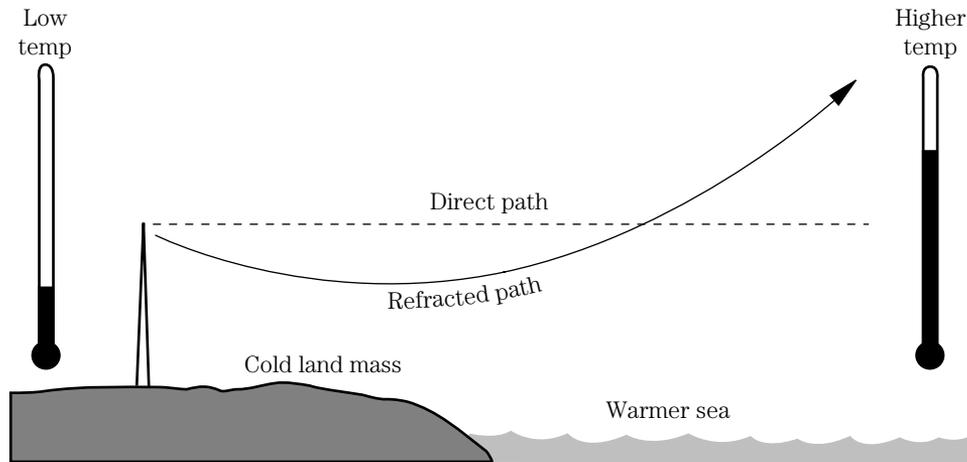
2-18 Superrefraction phenomena.



2-19 Ducting phenomenon.

or so are possible through ducting. Certain paths where frequent ducting occurs have been identified: the Great Lakes to the Atlantic seaboard; Newfoundland to the Canary Islands; across the Gulf of Mexico from Florida to Texas; Newfoundland to the Carolinas; California to Hawaii; and Ascension Island to Brazil.

Another condition is noted in the polar regions, where colder air from the land mass flows out over warmer seas (Fig. 2-20). Called *subrefraction*, this phenomena bends EM waves away from the earth's surface—thereby *reducing* the radio horizon by about 30 to 40 percent.



2-20 Subrefraction phenomena.

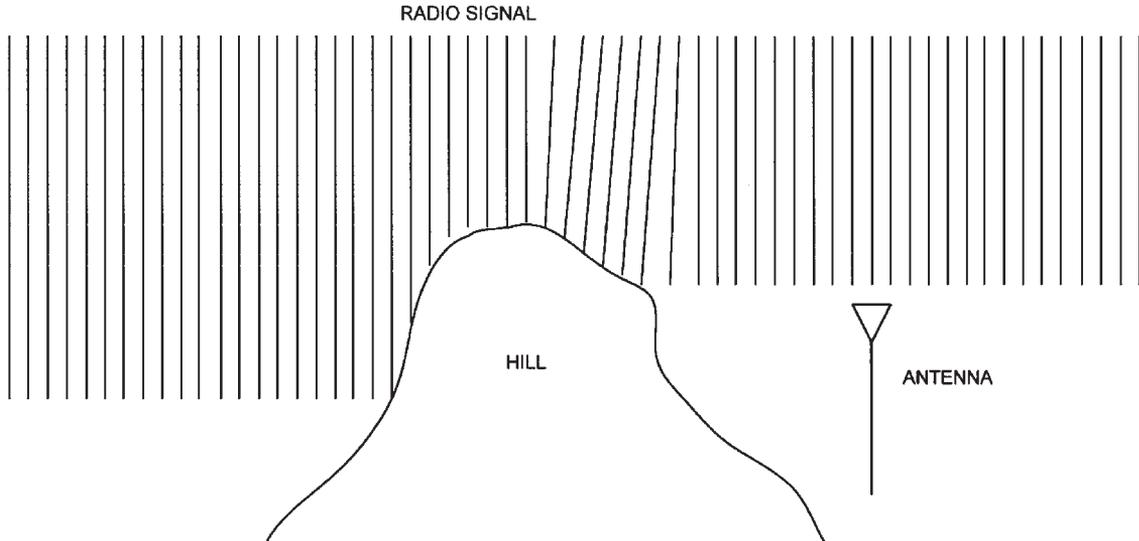
All tropospheric propagation that depends upon air-mass temperatures and humidity, shows diurnal (i.e., over the course of the day) variation caused by the local rising and setting of the sun. Distant signals may vary 20 dB in strength over a 24-hour period. These tropospheric phenomena explain how TV, FM broadcast, and other VHF signals can propagate great distances, especially along seacoast paths, at some times while being weak or nonexistent at others.

**Diffraction phenomena** Electromagnetic *waves* diffract when they encounter a radio-opaque object. The degree of diffraction, and the harm it causes, is frequency-related. Above 3 GHz, wavelengths are so small (approximately 10 cm) compared to object sizes that large attenuation of the signal occurs. In addition, beamwidths (a function of antenna size compared with wavelength) tend to be small enough above 3 GHz that blockage of propagation by obstacles is much greater.

Earlier in this chapter, large-scale diffraction around structures (such as buildings) was discussed. The view presented was from above, so it represented the horizontal plane. But there is also a diffraction phenomenon in the vertical plane. Terrain, or man-made objects, intervening in the path between UHF microwave stations (Fig. 2-21A) cause diffraction, and some signal attenuation. There is a minimum clearance required to prevent severe attenuation (up to 20 to 30 dB) from diffraction. Calculation of the required clearance comes from Huygens-Fresnel wave theory.

Consider Fig. 2-21B. A wave source  $A$ , which might be a transmitter antenna, transmits a wavefront to a destination  $C$  (receiver antenna). At any point along path  $A$ - $C$ , you can look at the wavefront as a partial spherical surface ( $B_1$ - $B_2$ ) on which all wave rays have the same phase. This plane can be called an *isophase plane*. You can assume that the  $d_n/d_h$  refraction gradient over the height extent of the wavefront is small enough to be considered negligible.

Using ray tracing we see rays  $r_a$  incoming to plane  $[B_1$ - $B_2]$ , and rays  $r_b$  outgoing from plane  $[B_1$ - $B_2]$ . The signal seen at  $C$  is the algebraic sum of all rays  $r_b$ . The signal pattern will have the form of an optical interference pattern with wave cancellation occurring between  $r_b$  waves that are a half-wavelength apart on  $[B_1$ - $B_2]$ . The ray in-



2-21A Terrain masking of VHF and higher-frequency signals.

points on plane  $[B_1-B_2]$  form radii  $R_n$  called *Fresnel zones*. The lengths of the radii are a function of the frequency and the ratio of the distances  $D_1$  and  $D_2$  (see Fig. 2-21B). The general expression is

$$R_n = M \sqrt{\frac{N}{F_{\text{GHz}}} \left( \frac{D_1 D_2}{D_1 + D_2} \right)} \quad [2.28]$$

where

$R_m$  is the radius of the  $n$ th Fresnel zone

$F_{\text{GHz}}$  is the frequency in GHz

$D_1$  is the distance from the source to plane  $B_1-B_2$

$D_2$  is the distance from destination to plane  $B_1-B_2$

$N$  is an integer (1, 2, 3, . . .)

$M$  a constant of proportionality equal to

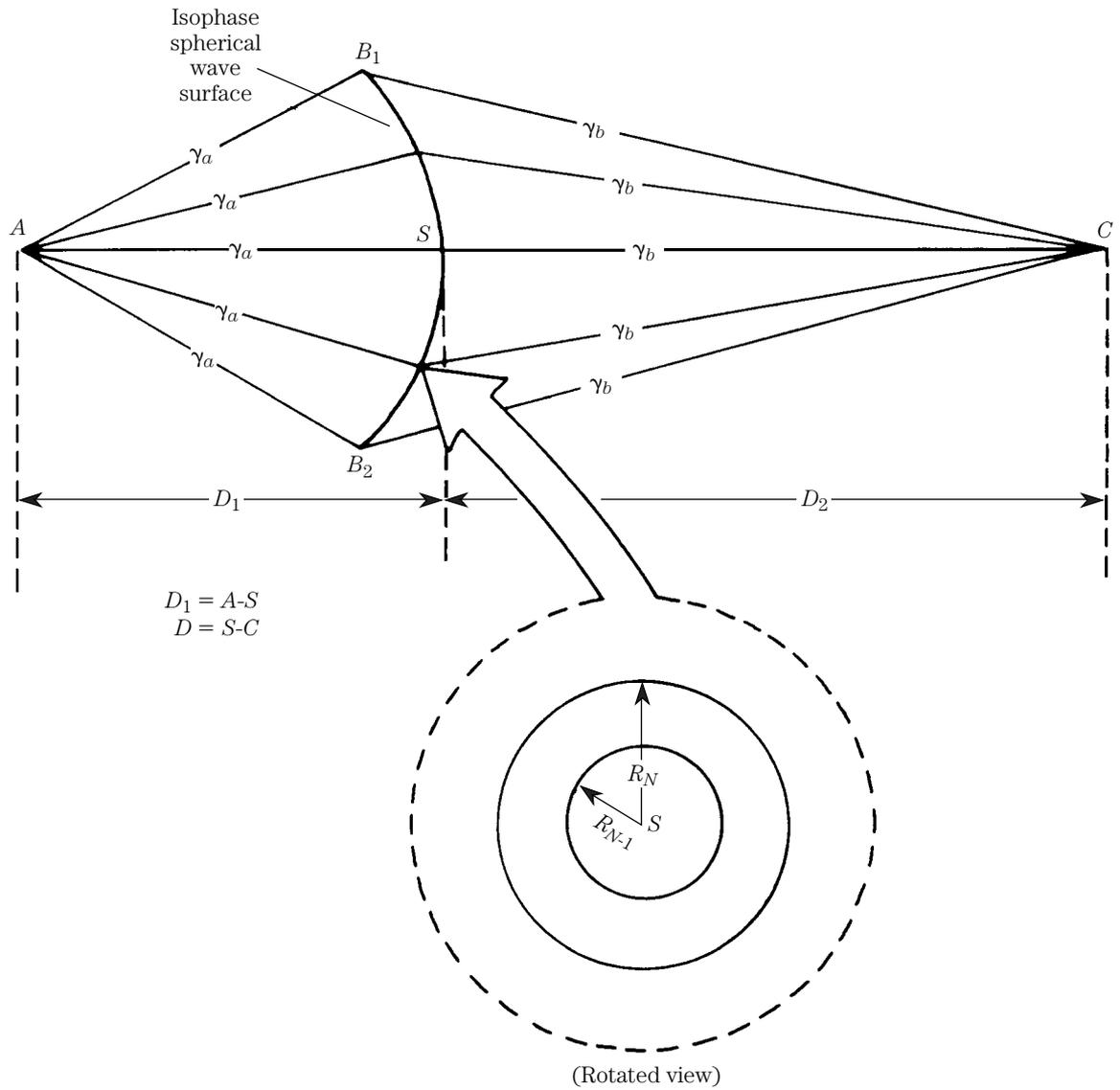
17.3 if  $R_m$  is in meters and  $D_1, D_2$  are in kilometers and

72.1 if  $R_m$  is in feet and  $D_1, D_2$  are in statute miles

If you first calculate the radius of the critical first Fresnel zone ( $R_1$ ), then you can calculate the  $n$ th Fresnel zone from

$$R_n = R_1 \sqrt{n} \quad [2.29]$$

**Example 2-2** Calculate the radius of the first Fresnel zone for a 2.5-GHz signal at a point that is 12 km from the source and 18 km from the destination.



2-21B Fresnel zone geometry.

**Solution:**

$$R_1 = M \sqrt{\frac{N}{F_{\text{GHz}}} \left( \frac{D_1 D_2}{D_1 + D_2} \right)}$$

$$R_1 = (17.3) \sqrt{\frac{1}{2.5} \left( \frac{12 \times 18}{12 + 18} \right)}$$

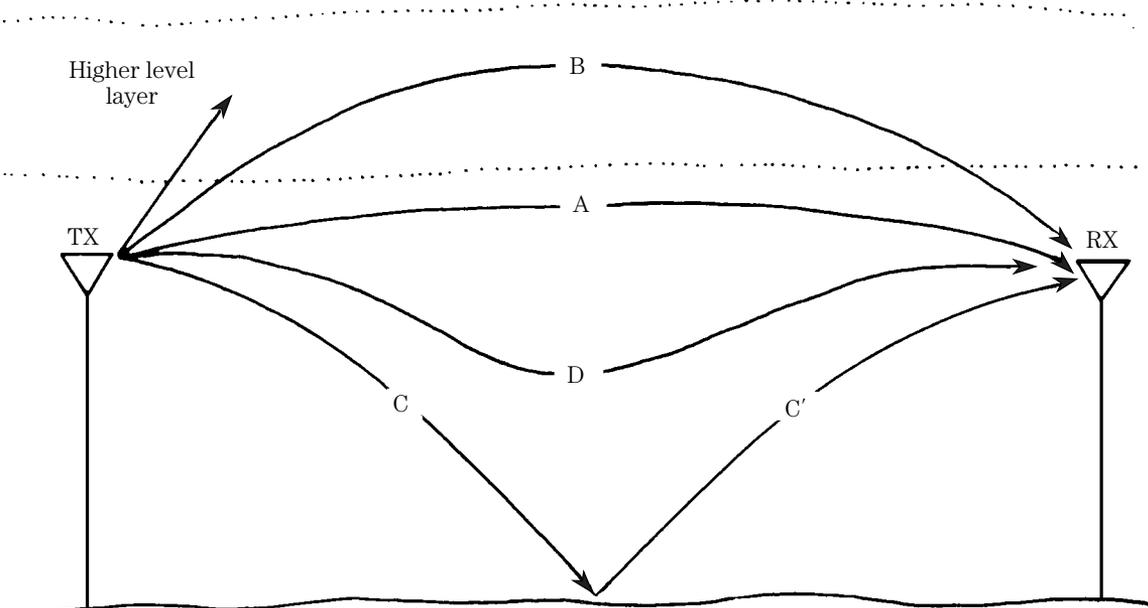
$$R_t = (17.3) \sqrt{(0.4) (7.2)}$$

$$R_1 = (17.3) (2.88) = 29.4 \text{ m}$$

For most terrestrial microwave systems an obstacle clearance of  $0.6R_1$  is required to prevent diffraction attenuation under most normal conditions. However, there are conditions in which the clearance zone should be more than one Fresnel zone.

**Fading mechanisms** Fading is defined as a reduction in amplitude caused by reduced received signal power, changes in phase or polarization, wave cancellation, or other related problems that are not caused by a change in the output power level or other parameters associated with either the transmitter or its antenna system. You would not ordinarily think that line-of-sight radio relay links would experience fading, but that is not true. Fading does, in fact, occur, and it can reach levels of 30 dB in some cases (20 dB is relatively common). In addition, fading phenomena in the VHF-and-up range can last several hours, with some periods being reported of several days in duration (although very rare). There are several mechanisms of fading, and these will be dealt with in this section. HF fading caused by ionospheric mechanisms will be covered later.

Any or all of the mechanisms shown in Fig. 2-22 can occur in a given system. In all cases, two or more signals arrive at the receiver antenna ( $R_r$ ). Ray *A* represents the direct path signal that is, ideally, the only signal to reach the destination. But it is also possible that a signal, in an elevated layer or other atmospheric anomaly, will cause refraction or subrefraction of the wave creating a second component, *B*. If this second signal arrives out of phase with *A*, then fading will occur (signal reinforcement—



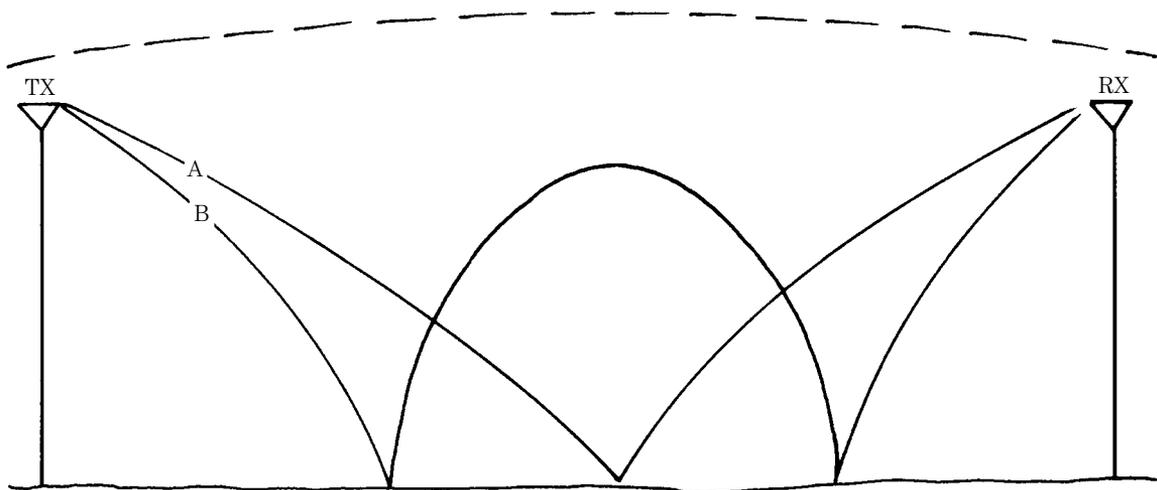
2-22 Multiple paths for signal to take between transmitter and receiver.

in phase—can also occur). It is also possible to see subrefraction fading, as in D. The classical multipath situation represented by ray C and its reflected component C' is also a source of fading.

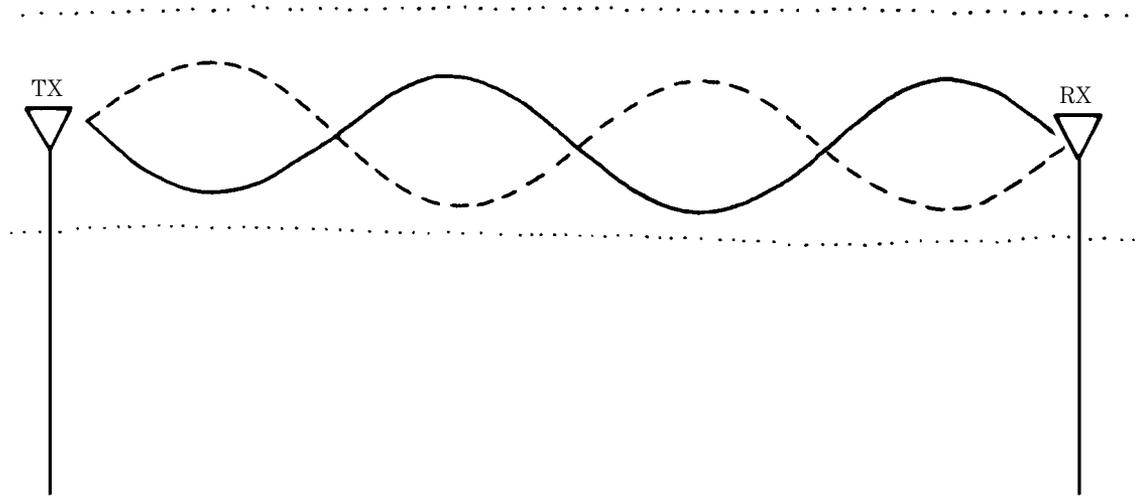
These mechanisms are frequency-sensitive, so a possible countermeasure is to use *frequency diversity*. Hopping over a 5 percent frequency change will help eliminate fading in many cases. In cases where either system constraints, or local spectrum usage prevents a 5 percent delta (change), then try for at least 2 or 3 percent.

Over ocean areas or other large bodies of water there is a possibility of encountering fair weather surface ducting as a cause of fading. These ducts form in the mid latitudes, starting about 2 to 3 km from shore, up to heights of 10 to 20 mi; wind velocities are found in the 10 to 60 km/h range. The cause of the problem is a combination of power fading, due to the presence of the duct, and surface reflections (see Fig. 2-23). Power fading alone can occur when there is a superrefractive duct elevated above the surface. The duct has a tendency to act as a waveguide and focus the signal (Fig. 2-24). Although the duct shown is superrefractive, it is also possible to have a subrefractive duct.

**Attenuation in weather** Microwave communications above about 10 GHz suffer an increasingly severe attenuation because of water vapor and oxygen in the atmosphere. Figure 2-25 shows the standard attenuation in dB/km for microwave frequencies. Note that there are several strong peaks in an ever-increasing curve. Setting a system frequency in these regions will cause poor communications or will require a combination of more transmit power, better receiver sensitivity, and better antennas on receiving and transmitting locations. The curves shown in Fig. 2-25 assume certain standardizing conditions. Rain and other weather conditions can severely increase the attenuation of signals. In addition to attenuation, radar exhibits severe clutter problems when signals backscatter from rain cells.



2-23 Multihop interference.



2-24 Wave interference.

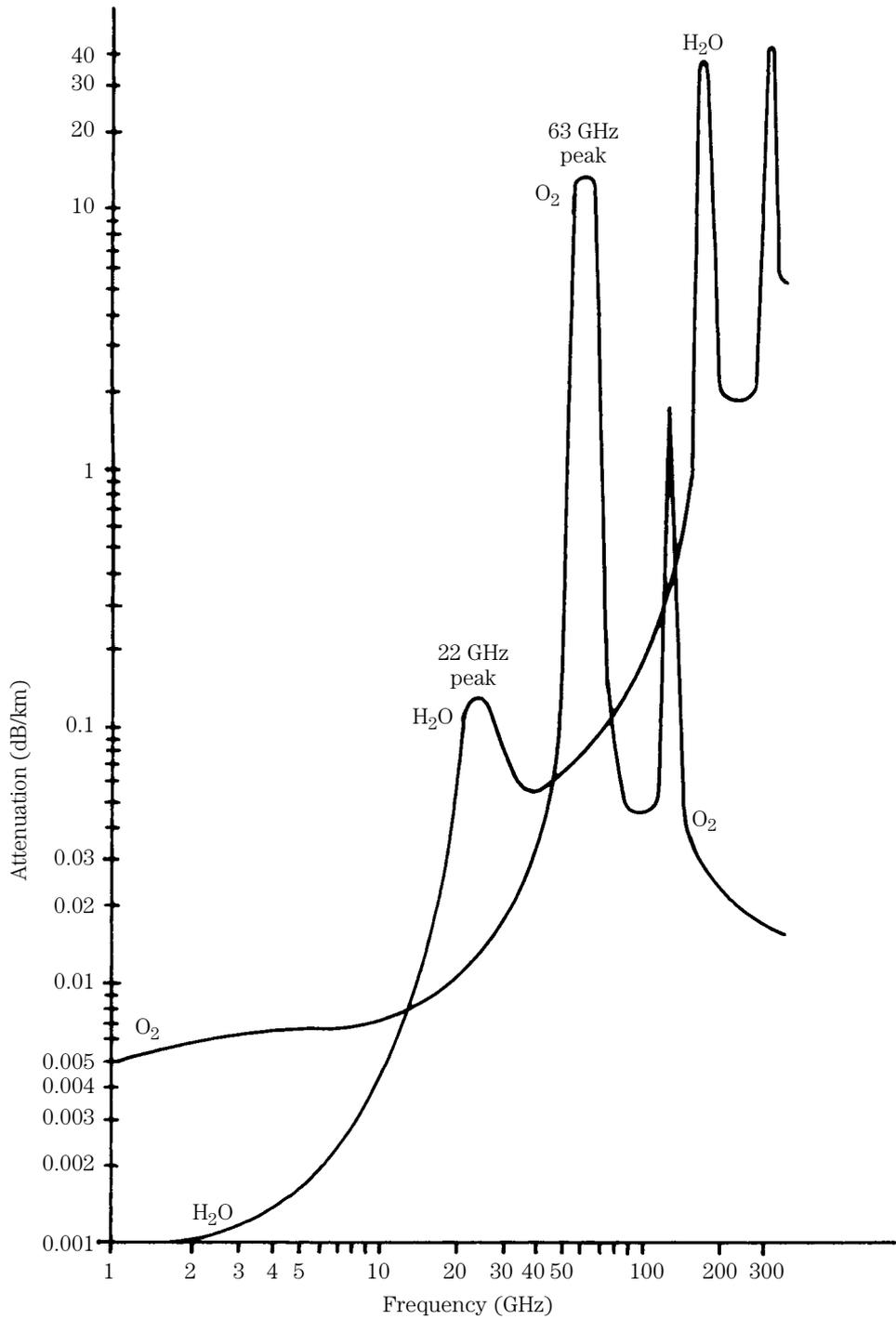
### Ionospheric propagation

Now let's turn your attention to the phenomenon of skip communications. Ionospheric propagation is responsible for the ability to do intercontinental broadcasting and communications. Long-distance radio transmission is carried out on the high-frequency (HF) bands (3 to 30 MHz), also called the *shortwave* bands. These frequencies are used because of the phenomenon called *skip*. Under this type of propagation, the earth's ionosphere acts as though it is a "radio mirror." Although the actual phenomenon is based on refraction (not reflection, as is frequently believed), the appearance to the casual observer is that shortwave and low-VHF radio signals are reflected from the ionosphere. The actual situation is a little different.

The key lies in the fact that this radio mirror is produced by ionization of the upper atmosphere. The upper portion of the atmosphere is called the *ionosphere* because it tends to be easily ionized by solar and cosmic radiation phenomena. The reason for the ease with which that region (30 to 300 mi above the surface) ionizes, is that the air density is very low. Energy from the sun strips away electrons from the outer shells of oxygen and nitrogen molecules. The electrons become negative ions, while the remaining portion of the atom forms positive ions. Because the air is so rarified at those altitudes, those ions can travel great distances before recombining to form electrically neutral atoms again. As a result, the average ionization level remains high in that region.

Several sources of energy will cause ionization of the upper atmosphere. Cosmic radiation from outer space causes some degree of ionization, but the majority of ionization is caused by solar energy.

The role of cosmic radiation was noticed during World War II, when British radar operators discovered that the distance at which their equipment could detect German aircraft was dependent upon whether or not the Milky Way was above the horizon. Intergalactic radiation raised the background microwave noise level, thereby adversely affecting the signal-to-noise ratio.



2-25 Atmospheric absorption of radio signals at microwave frequencies.

Events on the surface of the sun sometimes cause the radio mirror to seem to be almost perfect, and this situation makes spectacular propagation possible. At other times, however, solar disturbances (Fig. 2-26A) disrupt radio communications for days at a time.

There are two principal forms of solar energy that affect shortwave communications: *electromagnetic radiation* and *charged solar particles*. Most of the radiation is above the visible spectrum, in the ultraviolet and x-ray/gamma-ray region of the spectrum. Because electromagnetic radiation travels at the speed of light, solar events that release radiation cause changes to the ionosphere about 8 minutes later. Charged particles, on the other hand, having a finite mass must travel at a considerably slower velocity. They require two or three days to reach earth.

Various sources of both radiation and particles exist on the sun. Solar flares can release huge amounts of both radiation and particles. These events are unpredictable and sporadic. Solar radiation also varies over an approximately 27-day period, which is the rotational period of the sun. The same source of radiation will face the earth once every 27 days, and so events tend to be somewhat repetitive.

Solar and galactic noise affect the reception of weak signals. Solar noise can also affect radio propagation and act as a harbinger of changes in propagation patterns. Solar noise can be demonstrated by using an ordinary radio receiver and a directional antenna, preferably operating in the VHF/UHF regions of the spectrum (150 to 152 MHz frequently is used). Aim the antenna at the sun on the horizon at either sunset or sunrise. A dramatic change in background noise will be noted as the sun slides across the horizon.



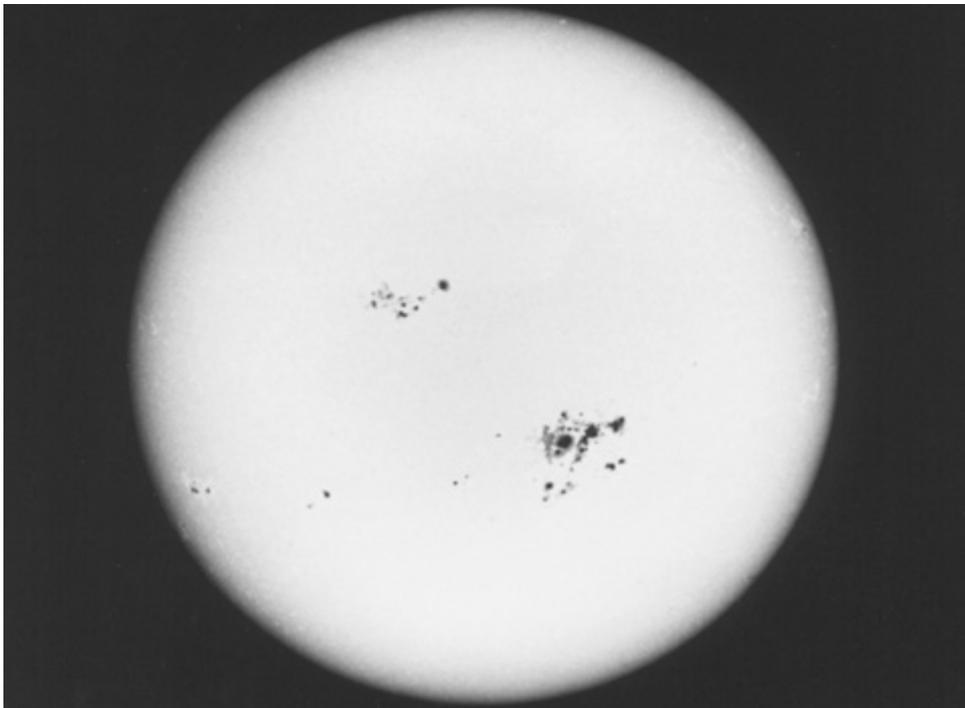
**2-26A** Solar event that can affect radio propagation on earth.

**Sunspots** A principal source of solar radiation, especially the periodic forms, is sunspots (Fig. 2-26B). Sunspots can be as large as 70,000 to 80,000 miles in diameter, and generally occur in clusters. The number of sunspots varies over a period of approximately 11 years, although the actual periods since 1750 (when records were first kept) have varied from 9 to 14 years. The sunspot number is reported daily as the statistically massaged *Zurich Smoothed Sunspot Number*, or *Wolf Number*. The number of sunspots greatly affects radio propagation via the ionosphere. The low was in the range of 60 (in 1907), and the high was about 200 (1958).

Another indicator of ionospheric propagation potential is the *solar flux index* (SFI). This measure is taken in the microwave region (wavelength of 10.2 cm, or 2.8 GHz), at 1700 UTC at Ottawa, Canada. The SFI is reported by the National Institute for Standards and Technology (NIST) radio stations WWV (Fort Collins, CO) and WWVH (Maui, Hawaii).

The ionosphere offers different properties that affect radio propagation at different times. Variations occur not only over the 11 year sunspot cycle, but also diurnally and seasonally. Obviously, if the sun affects propagation in a significant way, then differences between nighttime and daytime, and between summer and winter, must cause variations in the propagation phenomena observed.

The ionosphere is divided, for purposes of radio propagation studies, into various layers that have somewhat different properties. These layers are only well defined



**2-26B** Sunspots.

in textbooks. However, even there you find a variation in the precise altitudes of the layers above the earth's surface. In addition, the real physical situation is such that layers don't have sharply defined boundaries, but rather fade one into another instead. Thus, the division into layers is somewhat arbitrary. These layers (Fig. 2-27) are designated D, E, and F (with F being further subdivided into F1 and F2 sublayers).

**D layer** The D layer is the lowest layer in the ionosphere, and exists from approximately 30 to 50 mi above the surface. This layer is not ionized as much as the higher layers, because all forms of solar energy that cause ionization are severely attenuated by the higher layers above the D layer. The reason for this is that the D layer is much more dense than the E and F layers, and that density of air molecules allows ions to recombine to form electroneutral atoms very quickly.

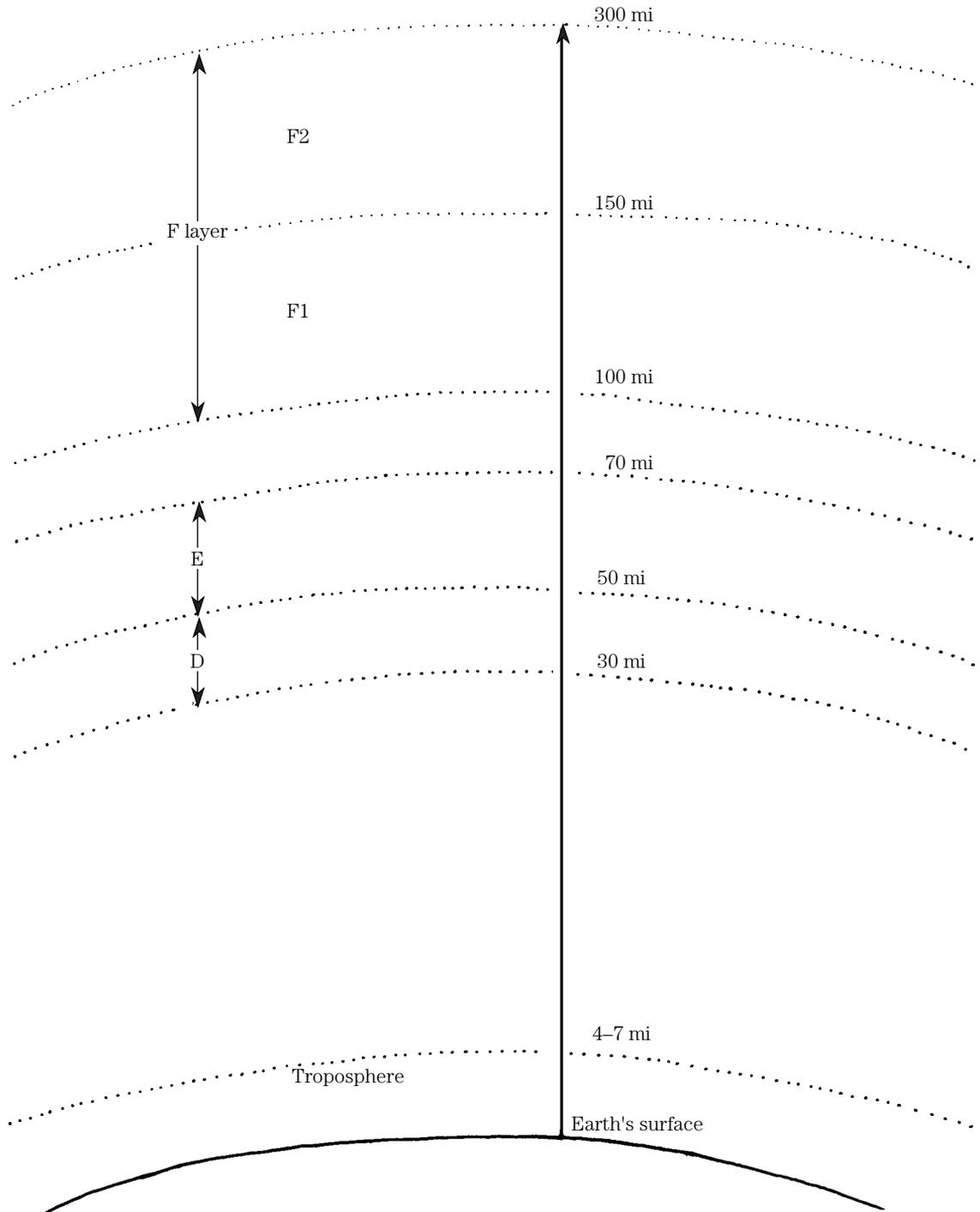
The extent of D layer ionization is proportional to the elevation of the sun, so it will achieve maximum intensity at midday. The D layer exists mostly during the warmer months of the year because of both the greater height of the sun above the horizon and the longer hours of daylight. The D layer almost completely disappears after local sunset. Some observers have reported sporadic incidents of D layer activity for a considerable time past sunset. The D layer exhibits a large amount of absorption of medium-wave and shortwave signals (to such an extent that signals below 4 to 6 MHz are completely absorbed by the D layer).

**E layer** The E layer exists from approximately 50 to 70 mi above the earth's surface, and it is considered the lowest region of the ionosphere that is important to radio communications. Like the D layer, this region is ionized only during the daylight hours, with ionization levels peaking at midday. The ionization level drops off sharply in the late afternoon, and almost completely disappears after local sunset.

During most of the year, the E layer is absorptive and it will not reflect radio signals. During the summer months, however, E layer propagation does occur. A phenomenon called "short skip" (i.e., less than 100 mi for medium wave and 1000 mi for shortwave signals) occurs in the E layer during the summer months, and in equatorial regions at other times.

A propagation phenomenon associated with the E layer is called *sporadic E propagation*. This phenomenon is caused by scattered zones of intense ionization in the E layer region of the ionosphere. The sporadic E phenomenon varies seasonally, and it is believed to be caused by the bombardment of solar particles. Sporadic E propagation affects the upper HF and lower VHF region. It is observed most frequently in the lower VHF spectrum (50 to 150 MHz), but it is also sometimes observed at higher frequencies. The VHF bands occasionally experience sporadic E propagation. Skip distances on VHF can reach 500 to 1500 miles on one hop—especially in the lower VHF region (including the 6-m band).

**F layer** The F layer of the ionosphere is the region that is the principal cause of long-distance shortwave communications. This layer is located from about 100 to 300 mi above the earth's surface. Unlike the lower layers, the air density in the F layer is low enough that ionization levels remain high all day, and decay slowly after local sunset. Minimum levels are reached just prior to local sunrise. Propagation in the F layer is capable of skip distances up to 2500 mi on a single hop. During the day there are actually two identifiable and distinct sublayers in the F layer region, and these are designated the F1 and F2 layers. The F1 layer is found approximately



2-27 Classification of the earth's atmosphere for radio propagation.

100 to 150 miles above the earth's surface, and the F2 layer is above the F1 extending up to the 270- to 300-mi limit. Beginning at local sundown, however, the lower regions of the F1 layer begin to deionize because of recombination of positive and negative ions. At some time after local sunset, the F1 and F2 layers have effectively merged to become a single reduced layer beginning at about 175 mi.

The height and degree of ionization of the F2 layer varies over the course of the day, with the season of the year, and with the 27-day sunspot cycle. The F2 layer begins to form shortly after local sunrise and reaches maximum shortly before noon. During the afternoon, the F2 layer ionization begins to decay in an exponential manner until, for purposes of radio propagation, it disappears sometime after local sunset. There is some evidence that ionization in the F layer does not completely disappear, but its importance to HF radio communication does disappear.

### Measures of ionospheric propagation

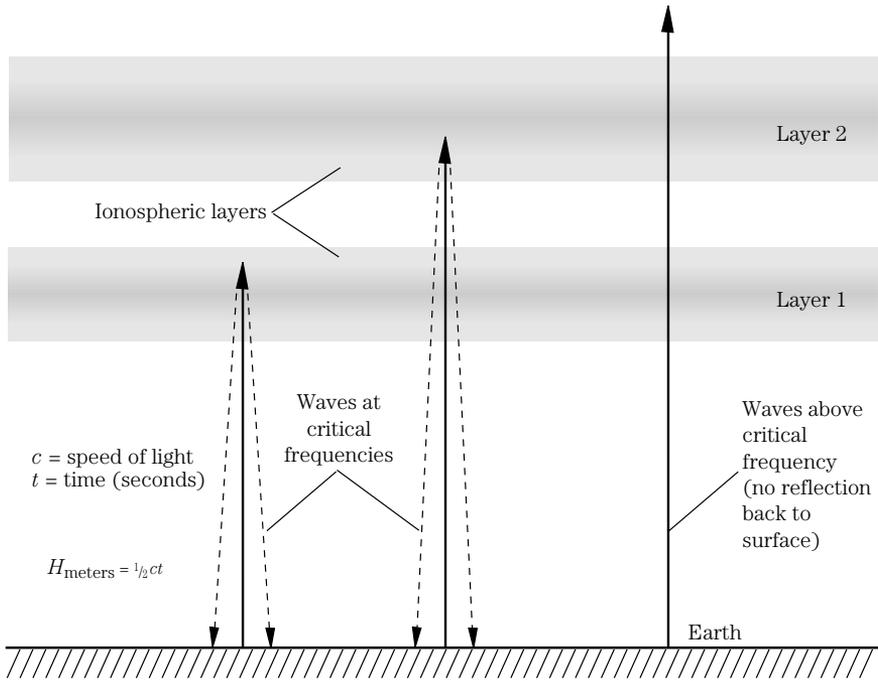
There are several different measures by which the ionosphere is characterized at any given time. These measures are used in making predictions of radio activity and long-distance propagation.

The *critical frequency* and *maximum usable frequency* (MUF) are indices that tell us something of the state of ionization and communications ability. These frequencies increase rapidly after sunrise and international communications usually begin within 30 minutes.

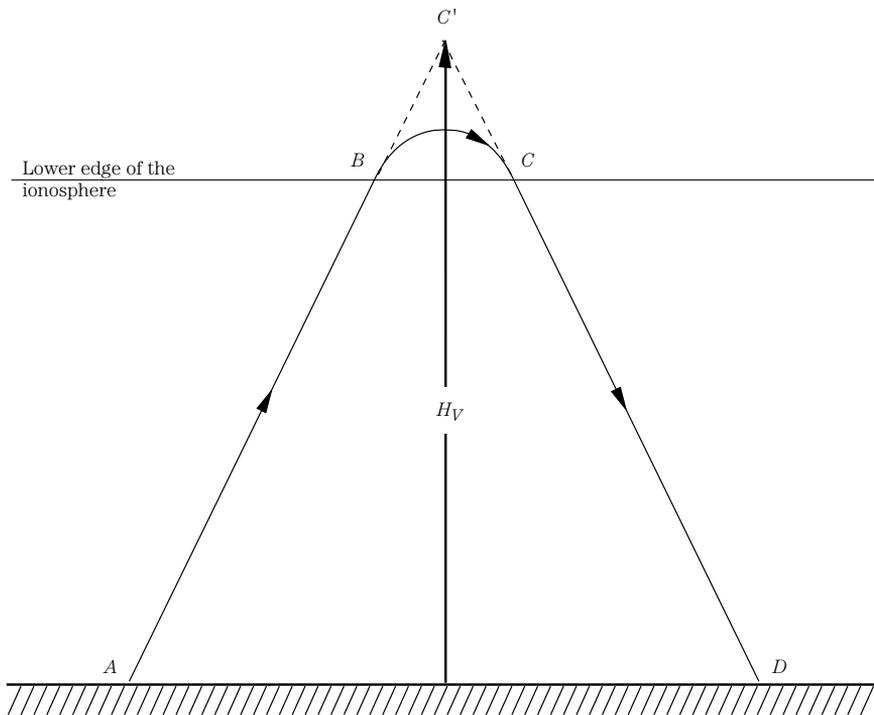
**Critical frequency  $F_c$**  The *critical frequency*, designated by  $F_c$ , is the highest frequency that can be reflected when a signal strikes the ionosphere as a vertical ( $90^\circ$  with respect to the surface) incident wave. The critical frequency is determined from an *ionogram*, which is a cathode-ray tube (CRT) oscilloscope display of the height of the ionosphere as a function of frequency. The ionogram is made by firing a pulse vertically (Fig. 2-28) at the ionosphere from the transmitting station. The critical frequency is that frequency that is just sufficient to be reflected back to the transmitter site. Values of  $F_c$  can be as low as 3 MHz during the nighttime hours, and as high as 10 to 15 MHz during the day.

**Virtual height** Radio waves are refracted in the ionosphere, and those above a certain critical frequency are refracted so much that they return to earth. Such waves appear to have been reflected from an invisible radio "mirror." An observer on the earth's surface could easily assume the existence of such a mirror by noting the return of the "reflected" signal. The height of this apparent "mirror" is called the *virtual height* of the ionosphere. Figure 2-29 shows the refraction phenomenon by which a radio wave is bent sufficiently to return to earth. Virtual height is determined by measuring the time interval required for an ionosonde pulse (similar to that used to measure critical frequency) to travel between the transmitting station and a receiving station (Fig. 2-30). A radio signal travels at a velocity of 300,000,000 m/s (the speed of light). By observing the time between transmitting the pulse and receiving it, you can calculate the virtual height of the ionosphere.

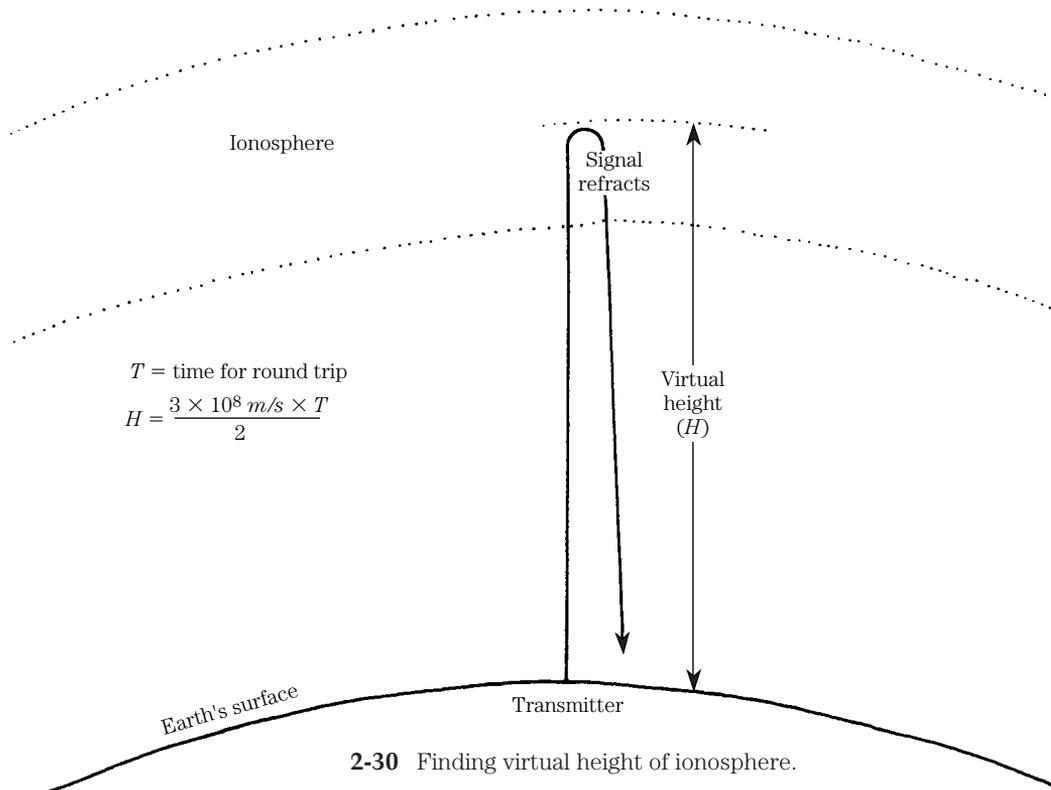
**Maximum usable frequency (MUF)** The *maximum usable frequency* is the highest frequency at which communications can take place via the ionosphere over a given path. The MUF between a fixed transmitter site and two different, widely separated, receivers need not be the same. Generally, however, the MUF is approximately



2-28 Finding critical height of the ionosphere.



2-29 Finding virtual height of the ionosphere.



$$T = \text{time for round trip}$$

$$H = \frac{3 \times 10^8 \text{ m/s} \times T}{2}$$

3 times higher than the critical frequency. Both the MUF and the critical frequency vary geographically, and they become higher at latitudes close to the equator.

It is a general rule that the best propagation occurs at frequencies just below the MUF. In fact, there is a so-called *frequency of optimum traffic* (FOT) that is approximately 85 percent of the MUF. Both noise levels and signal strengths are improved at frequencies near the FOT.

**Lowest usable frequency (LUF)** At certain low frequencies, the combination of ionospheric absorption, atmospheric noise, miscellaneous static, and/or receiver signal-to-noise ratio requirements conspire to reduce radio communications. The lowest frequency that can be used for communications, despite these factors, is the *lowest usable frequency*.

Unlike the MUF, the LUF is not totally dependent on atmospheric physics. The LUF of a system can be varied by controlling the signal-to-noise ratio (SNR). Although certain factors that contribute to SNR are beyond our control, the effective radiated power (ERP) of the transmitter can be changed; a 2-MHz decrease in LUF is available for every 10-dB increase in the ERP of the transmitter.

### Ionospheric variation and disturbances

The ionosphere is an extremely dynamic region of the atmosphere, especially from a radio operator's point of view, because it significantly alters radio propagation. The

dynamics of the ionosphere are conveniently divided into two general classes: *regular variation* and *disturbances*. This section covers both types of ionospheric change.

**Ionospheric variation** There are several different forms of variation seen on a regular basis in the ionosphere: *diurnal*, *27-day* (monthly), *seasonal*, and *11-year cycles*.

**Diurnal (daily) variation** The sun rises and falls on a 24-hour cycle, and because it is the principal source of ionization of the upper atmosphere, you can expect diurnal variation. During daylight hours the E and D levels exist, but these disappear at night. The height of the F2 layer increases until midday, and then it decreases until evening, when it disappears or merges with other layers. As a result of higher absorption in the E and D layers, lower frequencies are not useful during daylight hours. On the other hand, the F layers reflect higher frequencies during the day. In the 1- to 30-MHz region, the higher frequencies (>11 MHz) are used during daylight hours, and the lower frequencies (<11 MHz) at night.

**27-day cycle** Approximately monthly, this variation is caused by the rotational period of the sun. Sunspots are localized on the surface of the sun, so they will face the earth only during a portion of the month. As new sunspots are formed, they do not show up on the earthside face until their region of the sun rotates earthside.

**Seasonal cycle** The earth's tilt varies the exposure of the planet to the sun on a seasonal basis. In addition, the earth's yearly orbit is not circular; it is elliptical. As a result, the intensity of the sun's energy that ionizes the upper atmosphere varies with the seasons of the year. In general, the E, D, and F layers are affected—although the F2 layer is only minimally affected. Ion density in the F2 layer tends to be highest in winter, and less in summer. During the summer, the distinction between F1 and F2 layers is less obvious.

**11-year cycle** The number of sunspots, statistically averaged, varies on an approximately 11-year cycle. As a result, the ionospheric effects that affect radio propagation also vary on an 11-year cycle. Radio propagation, in the shortwave bands, is best when the average number of sunspots is at its highest.

**Disturbances** Disturbances in the ionosphere can have a profound effect on radio communications—and most of them (but not all) are bad. This section will briefly examine some of the more common forms.

**Sporadic E layer** A reflective cloud of ionization sometimes appears in the E layer of the ionosphere; this layer is sometimes called the  $E_s$  layer. It is believed that the  $E_s$  layer forms from the effects of wind shear between masses of air moving in opposite directions. This action appears to redistribute ions into a thin layer that is radio-reflective.

Sporadic E propagation is normally thought of as a VHF phenomenon, with most activity between 30 and 100 MHz, and decreasing activity up to about 200 MHz. However, about 25 to 50 percent of the time, sporadic E propagation is possible on frequencies down to 10 or 15 MHz. Reception over paths of 1400 to 2600 mi are possible in the 50-MHz region when sporadic E is present. In the northern hemisphere, the months of June and July are the most prevalent sporadic E months. On most days when sporadic E is present, it lasts only a few hours.

**Sudden ionospheric disturbances (SIDs)** The SID, or *Dellinger fade*, mechanism occurs suddenly, and rarely gives any warning. The SID can last from a few minutes to many hours. It is known that SIDs often occur in correlation with solar

flares, or “bright solar eruptions,” that produce an immense amount of ultraviolet radiation that impinges the upper atmosphere. The SID causes a tremendous increase in D layer ionization, which accounts for the radio propagation effects. The ionization is so intense that all receiver operators on the sunny side of the earth experience profound loss of signal strength above about 3 MHz. It is not uncommon for receiver owners to think their receivers are malfunctioning when this occurs. The sudden loss of signals on sunny side receivers is called *Dellinger fade*. The SID is often accompanied by variations in terrestrial electrical currents and magnetism levels.

**Ionospheric storms** The *ionospheric storm* appears to be produced by an abnormally large rain of atomic particles in the upper atmosphere, and is often preceded by SIDs 18 to 24 hours earlier. These storms tend to last from several hours, to a week or more, and are often preceded by two days or so by an abnormally large collection of sunspots crossing the solar disk. They occur, most frequently, and with greatest severity, in the higher latitudes, decreasing toward the equator. When the ionospheric storm commences, shortwave radio signals may begin to flutter rapidly and then drop out altogether. The upper ionosphere becomes chaotic; turbulence increases and the normal stratification into layers, or zones, diminishes.

Radio propagation may come and go over the course of the storm, but it is mostly dead. The ionospheric storm, unlike the SID, which affects the sunny side of the earth, is worldwide. It is noted that the MUF and critical frequency tend to reduce rapidly as the storm commences.

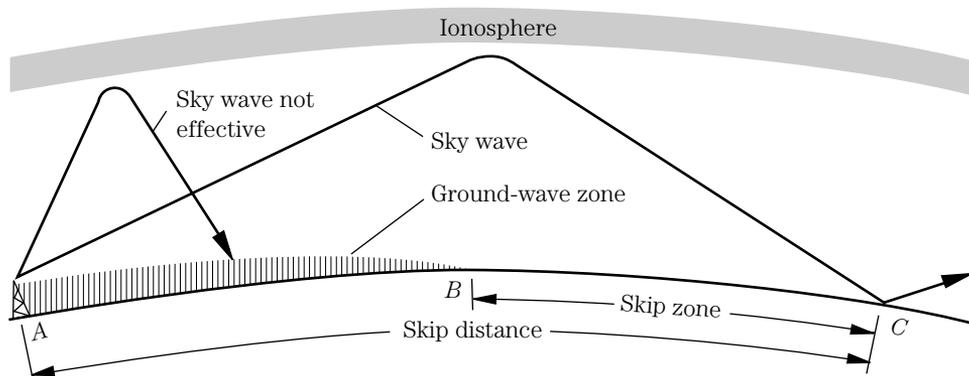
An ionospheric disturbance observed in November 1960 was preceded by about 30 minutes of extremely good, but abnormal, propagation. At 1500 hours EST, European stations were noted with S9+ signal strengths in the 7000- to 7300-kHz region of the spectrum, which is an extremely rare occurrence. After about 30 minutes, the bottom dropped out and even AM broadcast band skip (later that evening) was nonexistent. At the time, the National Bureau of Standards\* radio station, WWV, was broadcasting a “W2” propagation prediction at 19 and 49 minutes after each hour (which is terrible!). It was difficult to hear even the 5-MHz WWV frequency in the early hours of the disturbance, and it disappeared altogether for the next 48 hours. This signal’s fade-out occurred during the weekend of the annual ARRL “Sweepstakes” contest.

## Ionospheric sky-wave propagation

*Sky-wave propagation* occurs because signals in the ionosphere are refracted so much that they are bent back toward the earth’s surface. To observers on the surface, it looks like the signal was reflected from a radio mirror at the virtual height of the ionosphere. The *skip distance* is the surface distance between the transmitter point (*A* in Fig. 2-31) and the point where it returns to earth (point *C* in Fig. 2-31). The *ground-wave zone* is the distance from the transmitter site (*A* in Fig. 2-31) to where the ground wave fades to a low level, below usefulness (point *B* in Fig. 2-31). The *skip zone* is the distance from the outer edge of the ground-wave zone to the skip distance, or the distance from *B* to *C* in Fig. 2-31.

It is possible for the sky wave and the ground wave to interfere with each other at some frequencies, under some circumstances. When this happens, the sky wave has a relative phase that depends on its path length (among other

\* Now the National Institute of Standards and Technology.



2-31 Relationship between ground wave and different cases of sky wave.

things), so it will arrive at some seemingly random phase relative to the ground wave. Thus, the sky wave can selectively strengthen or cancel the ground wave, giving rise to a type of fading.

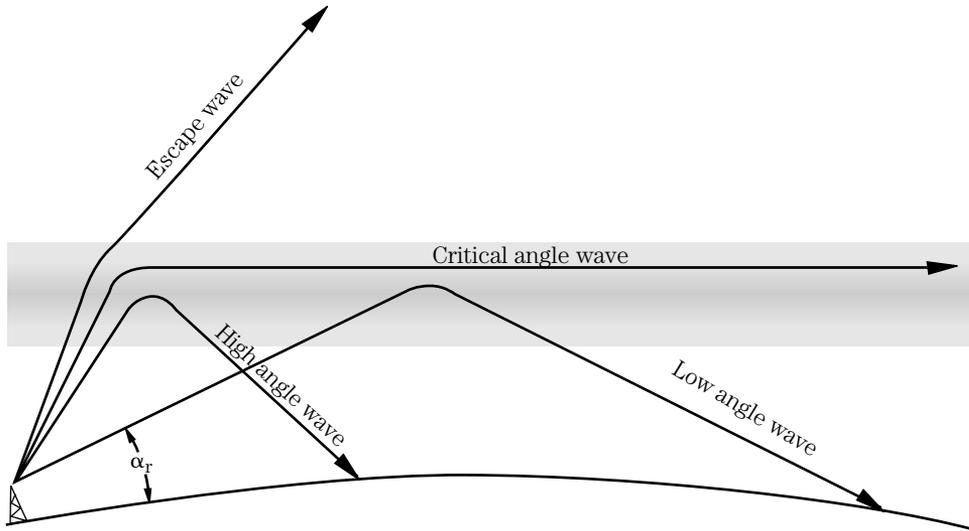
**Incident angle** One of the factors that affects the length of the skip distance is the incident angle of the radio wave. This angle is partially a function of the frequency, and partially a function of the natural radiation angle of the antenna ( $\alpha_r$ ). The frequency effects are seen as a function of the ionization level and how much a given frequency is refracted.

The antenna radiation angle is the angle of the main vertical lobe with respect to the earth's surface; it is partially a function of its design, and partially of its installation configuration. For example, a  $\frac{3}{8}$ -wavelength vertical antenna tends to have a lower angle of radiation than a  $\frac{1}{4}$ -wave vertical antenna. Similarly, the dipole's angle of radiation is a function of its height above ground.

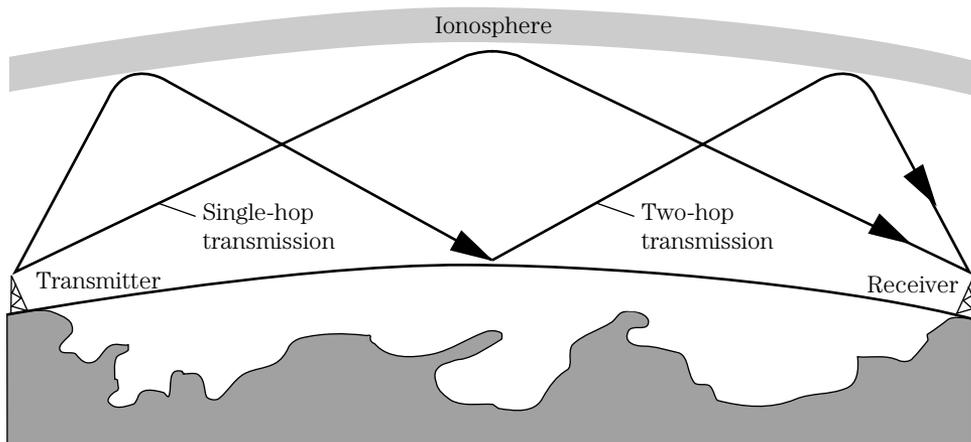
Figure 2-32 shows how the angle of radiation affects skip distance. Low-angle-of-radiation signals tend to travel farther with respect to the earth's surface before refracting, so they produce the longest skip distances. Higher angles of radiation have shorter skip distances because they tend to return to earth more rapidly. In summertime, some high-frequency bands (e.g., the 11-m—27 MHz—Citizens Band) offer high-angle “short skip” during the summer, and longer skip during the other months. At higher angles of radiation, there will be a *critical angle* wave and *escape angle* waves that are not returned to earth. These waves are not used for terrestrial communications or broadcasting. Figure 2-33 shows the difference between single-hop skip and multihop skip, as a function of incident angle. It is generally true that a multihop transmission is more subject to fading, and is weaker, than a single-hop transmission.

## Using the ionosphere

The refraction of high-frequency and some medium-wave radio signals back to earth via the ionosphere gives rise to intercontinental HF radio communications. This phenomenon becomes possible during daylight hours and for a while after sunset when the ionosphere is ionized. Figure 2-34 reiterates the mechanism of long-distance skip communications. The transmitter is located at point *T*, while receiving stations

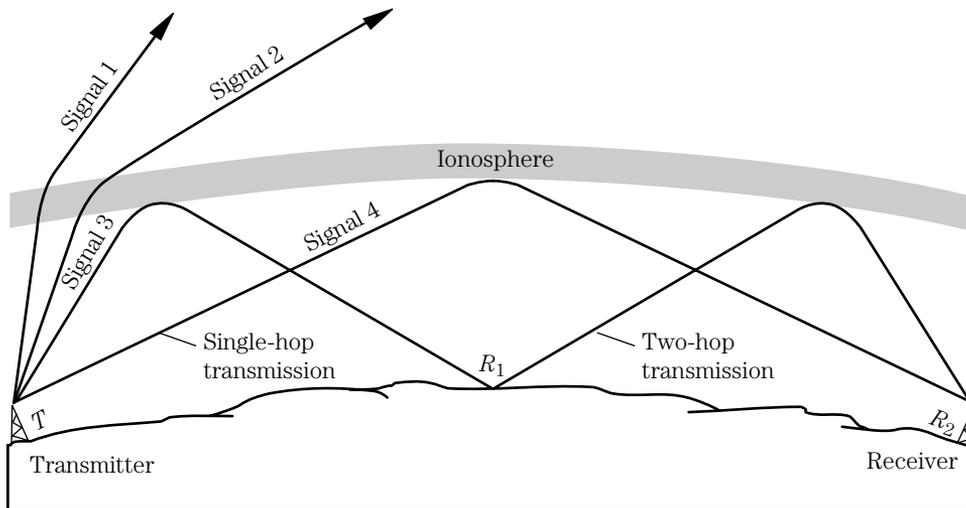


2-32 Sky-wave propagation as a function of antenna radiation angle.



2-33 Single-skip and multihop-skip communications.

are located at sites  $R_1$  and  $R_2$ . Signals 1 and 2 are not refracted sufficiently to be refracted back to earth, so they are lost in space. Signal 3, however, is refracted enough to return to earth, so it is heard at station  $R_1$ . The skip distance for signal 3 is the distance from  $T$  to  $R_1$ . At points between  $T$  and  $R_1$ , signal 3 is inaudible, except within ground-wave distance of the transmitter site ( $T$ ). This is the reason why two stations 40 mi apart hear each other only weakly, or not at all, while both stations can communicate with a third station 2000 mi away. In amateur radio circles, it is common for South American stations to relay between two U.S. stations only a few miles apart. For an example of this problem, listen to the Inter-American and Halo Missionary



2-34 Effects of radiation angle on distance.

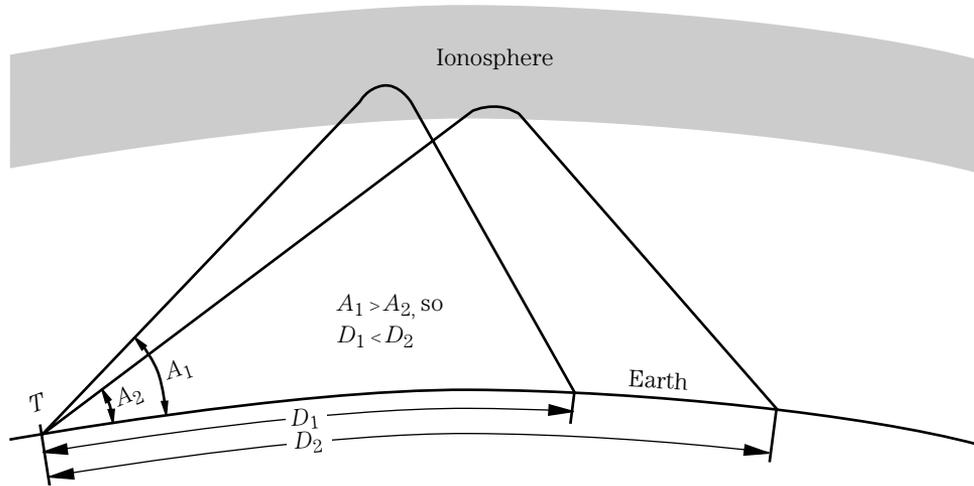
Nets on 21.290 MHz (15 m) daily from about 1700Z to 2100Z (ending time dependent upon traffic).

Multihop skip is responsible for the reception of the signal from transmitter  $T$  at site  $R_3$ . The signal *reflects* (not refracts) from the surface at  $R_1$ , and is retransmitted into the ionosphere, where it is again refracted back to earth.

Figure 2-35 shows a situation where skip signals are received at different distances depending upon the *angle of radiation* of the transmitting antenna. A high angle of radiation causes a shorter skip zone, but a lower angle of radiation results in a longer skip zone. Communication between any particular locations on any given frequency requires adjustment of the antenna radiation angle. Some international shortwave stations have multiple antennas with different radiation angles to ensure that the correct skip distances are available.

**Great circle paths** A *great circle* is a line between two points on the surface of a sphere, such that it lays on a plane through the earth's center and includes the two points. When translated to "radiospeak," a great circle is the shortest path on the surface of the earth between two points. Navigators and radio operators use the great circle for similar, but different, reasons. The navigator's reason is in order to get from here to there, and the radio operator's is to get a transmission path from here to there.

The heading of a directional antenna is normally aimed at the receiving station along its great circle path. Unfortunately, many people do not understand the concept well enough, for they typically aim the antenna in the wrong direction. For example, I live near Washington, D.C., which is on approximately the same latitude as Lisbon, Portugal. If I catch a lift on Superman's back, and he flies due East, we'll have dinner in Lisbon, right? Wrong. If you head due east from Washington, DC, across the Atlantic, the first landfall would be west Africa, somewhere near Zaire or Angola. Why? Because the great circle bearing  $90^\circ$  takes us far south. The *geometry of spheres*, not flat planes, governs the case.



2-35 Skip propagation.

Figure 2-36 shows a great circle map centered on the Washington, D.C. area. These maps, or computer tabulations of the same data, can often be purchased for your own location by supplying your latitude and longitude to the service company that does the job. By drawing a line from your location at the center of the chart to the area you want to hear, and then extending it to the edge of the chart, you will obtain the beam heading required.

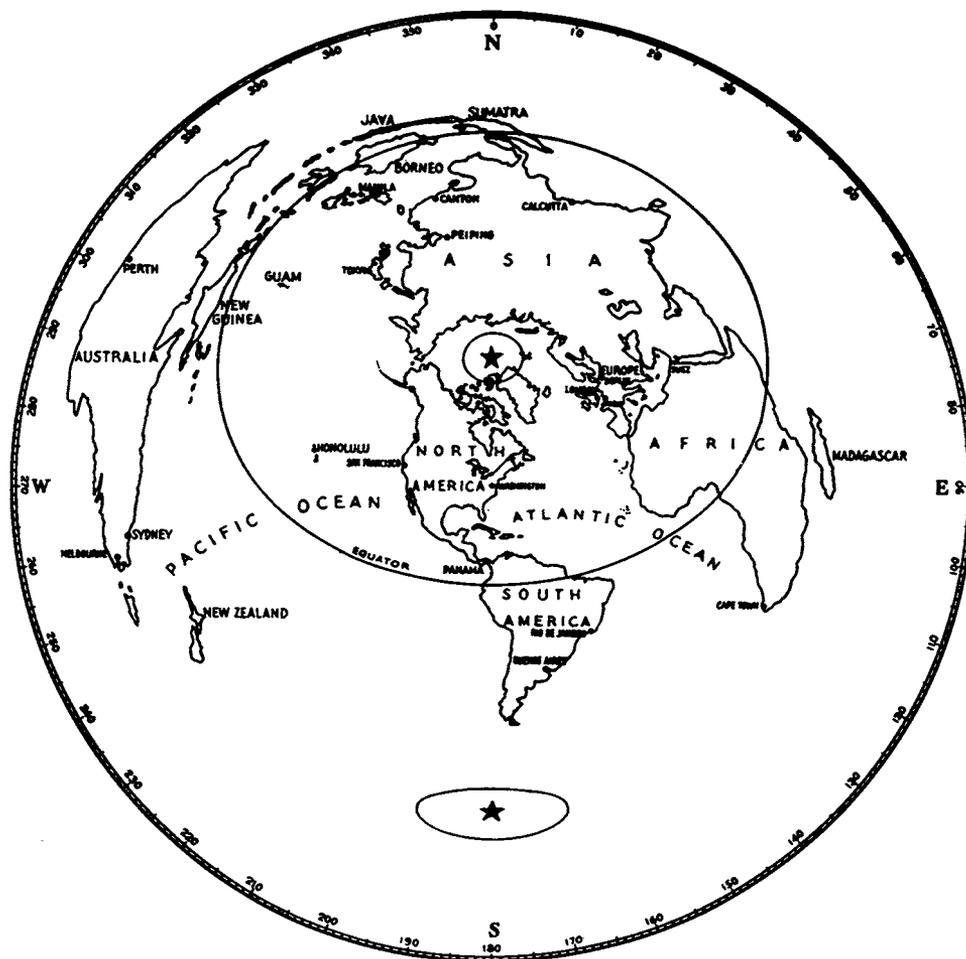
**Long path versus short path** The earth is a sphere (or more precisely, an oblate spheroid), so from any given point to any other point there are two great circle paths: the *long path* (major arc) and the *short path* (minor arc). In general, the best reception occurs along the short path. In addition, short-path propagation is more nearly “textbook,” compared with long path reception. However, there are times when long path is better, or is the only path that will deliver a signal to a specific location from the geographic location in question.

**Gray line propagation** The *gray line* is the twilight zone between the nighttime and daytime halves of the earth. This zone is also called the *planetary terminator* (Fig. 2-37). It varies up to  $+23^\circ$  either side of the north-south longitudinal lines, depending on the season of the year (it runs directly north-south only at the vernal and autumnal equinoxes). The D layer of the ionosphere absorbs signals in the HF region. This layer disappears almost completely at night, but it builds up during the day. Along the gray line, the D layer is rapidly decaying west of the line, and has not quite built up east of the line.

Brief periods of abnormal propagation occur along the gray line. Stations on either side of the line can be heard from regions, and at distances, that would otherwise be impossible on any given frequency. For this reason, radio operators often prefer to listen at dawn and dusk for this effect.

### Scatter propagation modes

*Ionospheric scatter propagation* occurs when clouds of ions exist in the atmosphere. These clouds can exist in both the ionosphere and the troposphere, although

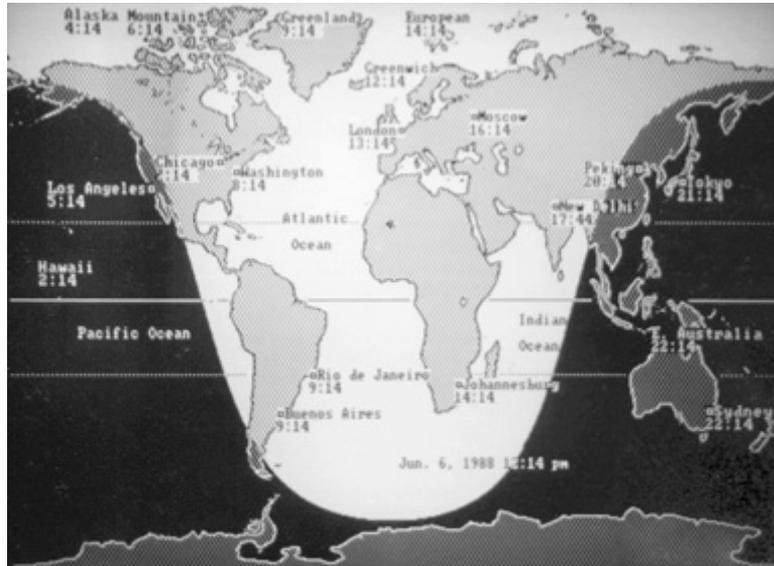


2-36 Azimuthal map centered on Washington, D.C. (Courtesy of *The ARRL Antenna Book*.)

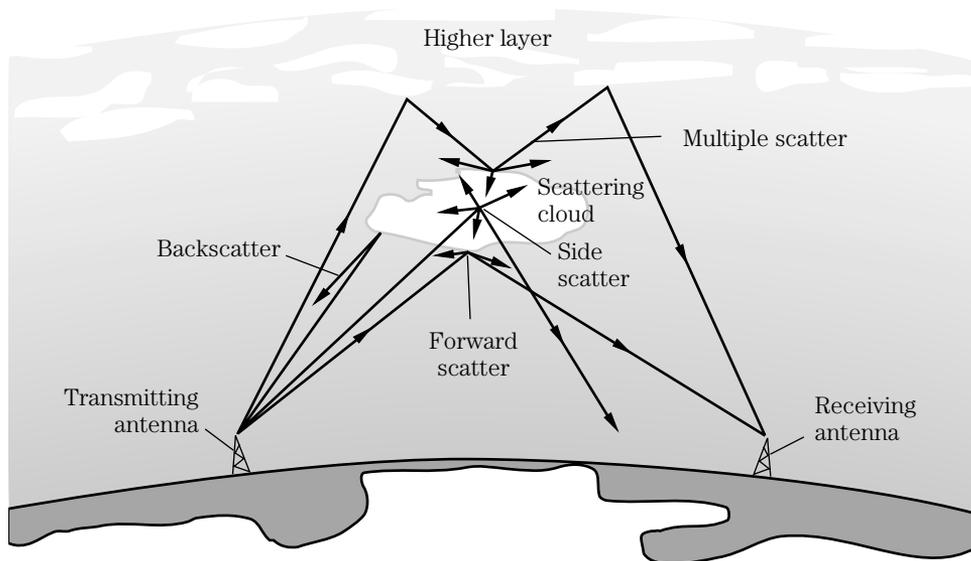
the tropospheric model is more reliable for communications. Figure 2-38 shows the mechanism for scatter propagation. Radio signals from the transmitter are reflected from the cloud of ions to a receiver location that otherwise might not receive it. Scatter propagation occurs mostly in the VHF region, and it allows communications over extended paths that are not normally available.

There are at least three different modes of scatter from ionized clouds: *backscatter*, *side scatter*, and *forward scatter*. The backscatter mode is a bit like radar, in that the signal is returned back to the transmitter site, or to regions close to the transmitter. Forward scatter occurs when the reflected signal continues in the same azimuthal direction (with respect to the transmitter), but is redirected toward the earth's surface. Side scatter is similar to forward scatter, but the azimuthal direction might change.

Unfortunately, there are often multiple reflections from the ionized cloud, and these are shown as "multiple scatter" in Fig. 2-38. When these reflections are able to



2-37 The planetary terminator (“gray line”) provides some unusual propagation effects. (Courtesy of MFJ Enterprises.)



2-38 Various modes of scatter propagation.

reach the receiving site, the result is a rapid, fluttery fading that can be of quite profound depths.

### **Auroral propagation**

The *auroral effect* produces a luminescence in the upper atmosphere resulting from bursts of particles released from the sun 18 to 48 hours earlier. The light emitted is called the *northern lights* and the *southern lights*. The ionized regions of the atmosphere that create the lights form a radio reflection shield, especially at VHF and above, although 15 to 20 MHz effects are known. Auroral propagation effects are normally seen in the higher latitudes, although listeners in the southern tier of states in the United States are often treated to the reception of signals from the north being reflected from auroral clouds.

### **Meteor scatter propagation**

When meteors enter the earth's atmosphere, they do more than simply heat up to the point of burning. The burning meteor leaves a wide, but very short duration, transient cloud of ionized particles in its path. These ions act as a radio mirror that permits short bursts of reception between sites correctly situated. Meteor scatter reception is not terribly reliable, although at least two companies offer meteor scatter communications services for users in the higher latitudes.

### **Other propagation anomalies**

The ionosphere is a physically complex place, and even the extensive coverage in this chapter is not sufficient to do it justice. Indeed, entire books are available on the subject, and it is a valid engineering subspecialty. It is therefore not surprising that a number of propagation anomalies are known.

**Nonreciprocal direction** If you listen to an amateur band receiver on the East Coast of the United States, you will sometimes hear European stations—especially in the late afternoon. But when you try to work those stations there is no reply whatsoever. They simply don't hear you! This propagation anomaly causes the radio wave to travel different paths dependent on which direction it travels; i.e., an east→west signal is not necessarily the reciprocal of a west→east signal. This anomaly can occur when a radio signal travels through a heavily ionized medium in the presence of a magnetic field, which is exactly the situation when the signal travels through the ionosphere in the presence of the earth's magnetic field.

Another anomaly seen in the radio literature of the 1930s is the *Radio Luxembourg effect*. It is named after the radio station where it was first noticed. In a nonlinear ionosphere, it is sometimes noted that the modulation of superpower (i.e., > 500,000 W) shortwave broadcasters will be transferred to the carrier of a weaker signal in the same or nearby band. The interchange noted in the 1930s when this phenomenon was first discovered was between Radio Luxembourg and Britain's British Broadcasting Corporation (BBC) overseas outlets.

## Propagation predictions

Propagation predictions for the VLF through low VHF bands are published each month in several magazines. Ham radio operators often use those in *QST* magazine (Fig. 2-39), while SWLs and others tend to prefer those published in magazines such as *Monitoring Times*. These charts relate the time of day in Universal Coordinated Time (UTC, formerly, GMT) and the frequency for transmission to different parts of the world.

## Fading

Skip communications are not without problems. One phenomenon is fading (i.e., a variation in signal strength as perceived at the receiver site). This problem can sometimes be overcome by using one of several diversity reception systems. Three forms of diversity technique are used: *frequency diversity*, *spatial diversity*, and *polarity diversity*.

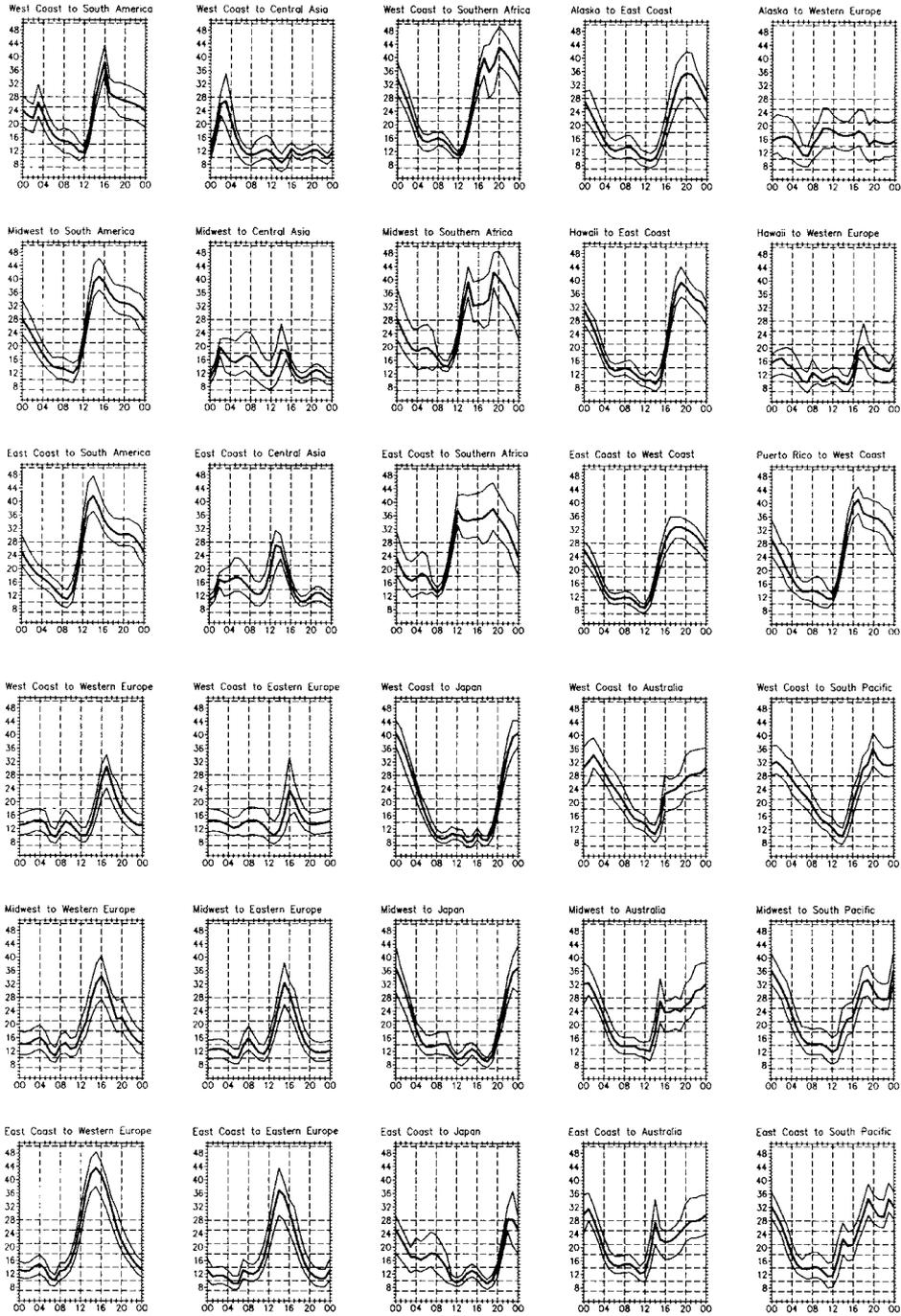
In the frequency diversity system (Fig. 2-40), the transmitter will send out two or more frequencies simultaneously with the same modulating information. Because the two frequencies will fade differentially, one will always be strong.

The spatial diversity system (Fig. 2-41) assumes that a single transmitter frequency is used. At the receiving site, two or more receiving antennas are used, spaced one-half wavelength apart. The theory is that the signal will fade at one antenna while it increases at the other. A three-antenna system is often used. Three separate, but identical, receivers, often tuned by the same master local oscillator, are connected to the three antennas. Audio mixing, based on the strongest signal, keeps the audio output constant while the radio-frequency (RF) signal fades.

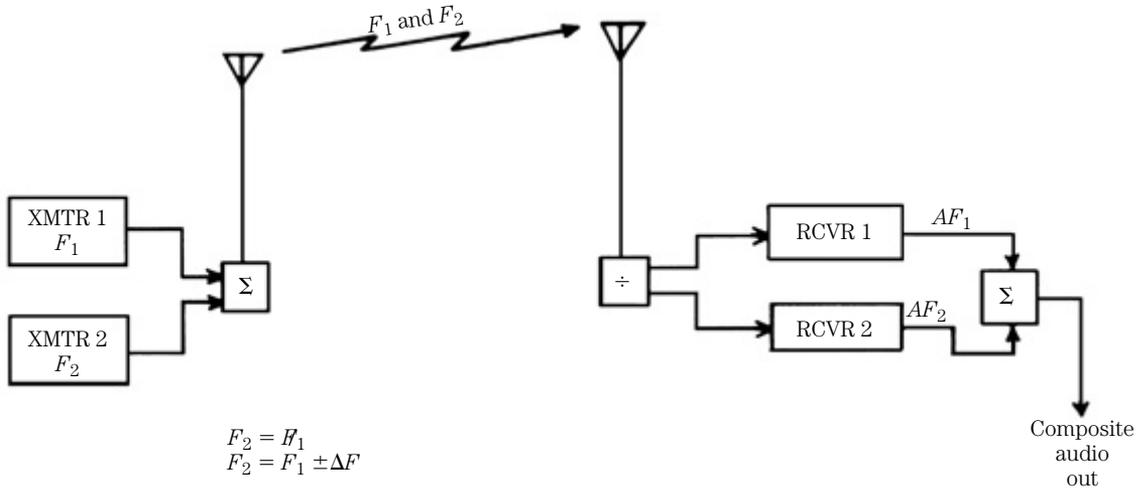
Polarity diversity reception (Fig. 2-42) uses both vertical and horizontal polarization antennas to receive the signal. As in the space diversity system, the outputs of the vertical and horizontal receivers are combined to produce a constant level output.

Another form of fading, *selective fading*, derives from the fact that fading is a function of frequency. The carrier and upper and lower sidebands of an AM signal have slightly different frequencies, so they arrive out of phase with each other. Although this type of fading is lessened by using single-sideband (SSB) transmission, that does not help AM users. In those systems, some people use a filtering system that eliminates the carrier and one sideband; it then reconstitutes the AM signal with a product detector.

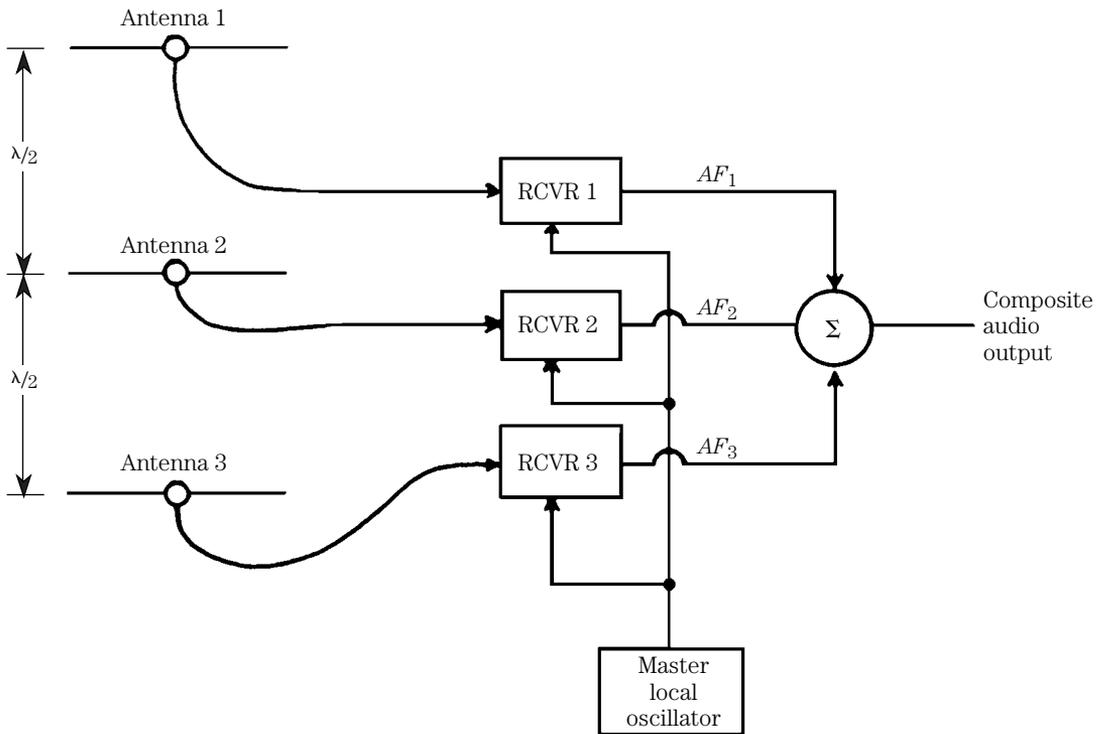
SSB receivers with stable local and product detector oscillators, and a sharp intermediate-frequency (IF) bandpass filter, can be used to reduce the effects of differential fading of AM signals because of the phasing of the lower sideband (LSB), upper sideband (USB), and carrier components. Carefully tune the receiver to only one sideband of the signal, and note when the heterodyne beatnote disappears. The correct point is characterized by the fact that you can then switch among USB, LSB, and continuous-wave (CW) modes without changing the received signal output.



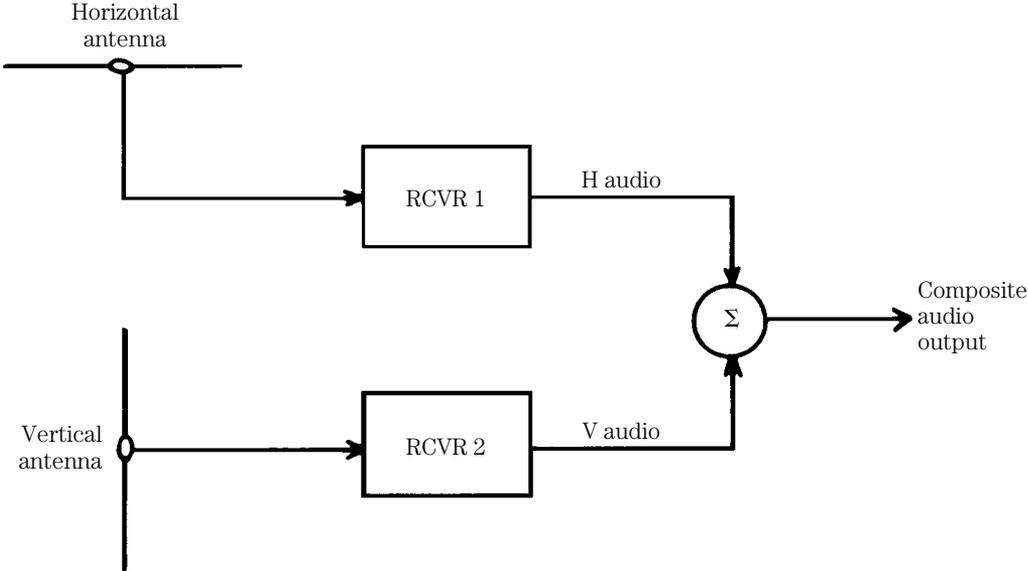
**2-39** Propagation prediction charts appear in magazines such as *QST* every month.  
 (Courtesy of the American Radio Relay League.)



2-40 Frequency diversity reduces fading.



2-41 Spatial diversity.



2-42 Polarization diversity.

## 3 CHAPTER

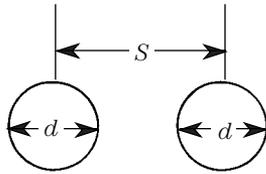
# Transmission lines

*TRANSMISSION LINES AND WAVEGUIDES* ARE CONDUITS FOR TRANSPORTING RF SIGNALS between elements of a system. For example, transmission lines are used between an exciter output and transmitter input, and between the transmitter input and its output, and between the transmitter output and the antenna. Although often erroneously characterized as a “length of shielded wire,” transmission lines are actually complex networks containing the equivalent of all the three basic electrical components: resistance, capacitance, and inductance. Because of this fact, transmission lines must be analyzed in terms of an *RLC* network.

## Parallel and coaxial lines

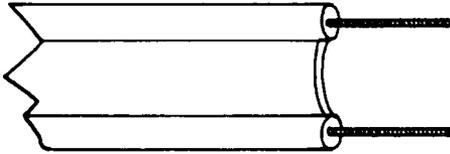
This chapter will consider several types of transmission lines. Both step-function and sine-wave ac responses will be studied. Because the subject is both conceptual and analytical, both analogy and mathematical approaches to the theory of transmission lines will be used.

Figure 3-1 shows several basic types of transmission line. Perhaps the oldest and simplest form is the *parallel line* shown in Figs. 3-1A through 3-1D. Figure 3-1A shows an end view of the parallel conductor transmission line. The two conductors, of diameter  $d$ , are separated by a dielectric (which might be air) by a spacing  $S$ . These designations will be used in calculations later. Figure 3-1B shows a type of parallel line called *twin lead*. This is the old-fashioned television antenna transmission line. It consists of a pair of parallel conductors separated by a plastic dielectric. TV-type twin lead has a characteristic impedance of  $300\ \Omega$ , while certain radio transmitting-antenna twin lead has an impedance of  $450\ \Omega$ . Another form of twin lead is *open line*, shown in Fig. 3-1C. In this case, the wire conductors are separated by an air dielectric, with support provided by stiff (usually ceramic) insulators. A tie wire (only one shown) is used to fasten each insulator end to the main conductor. Some users of open line prefer the form of insulator or supporter shown in Fig. 3-1D. This

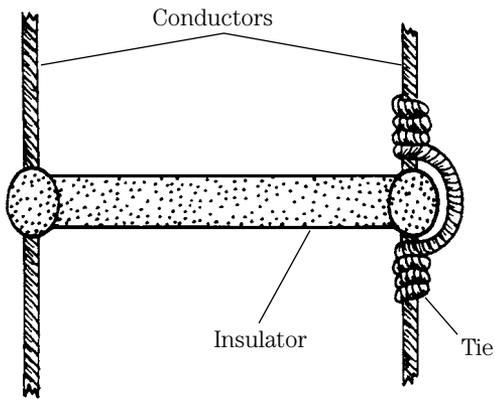


3-1A Parallel line transmission line (end view).

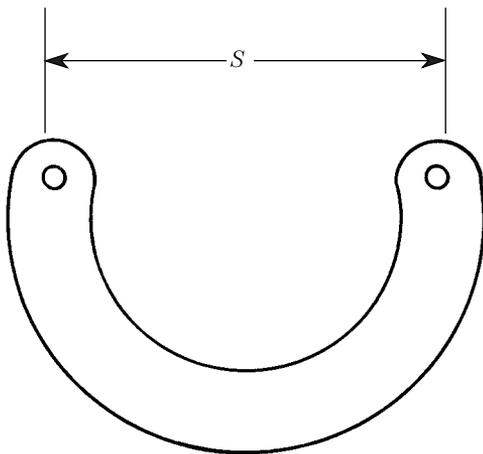
(End view)



3-1B Twin-lead transmission line.



3-1C Parallel line construction details.



3-1D Horseshoe parallel line spreader.

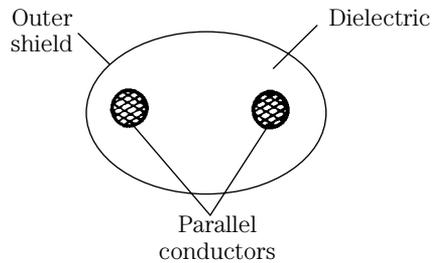
form of insulator is made of either plastic or ceramic, and is in the form of a U. The purpose of this shape is to reduce losses, especially in rainy weather, by increasing the leakage currents path relative to spacing  $S$ .

Parallel lines have been used at VLF, MW, and HF frequencies for decades. Even antennas into the low VHF are often found using parallel lines. The higher impedance of these lines (relative to coaxial cable) yields lower loss in high-power applications. For years, the VHF, UHF, and microwave application of parallel lines was limited to educational laboratories, where they are well suited to performing experiments (to about 2 GHz) with simple, low-cost instruments. Today, however, printed circuit and hybrid semiconductor packaging has given parallel lines a new lease on life, if not an overwhelming market presence.

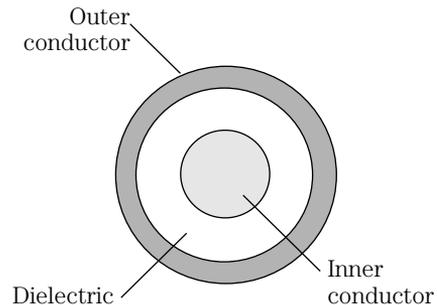
Figure 3-1E shows a form of parallel line called *shielded twin lead*. This type of line uses the same form of construction as TV-type twin lead, but it also has a braided shielding surrounding it. This feature makes it less susceptible to noise and other problems.

The second form of transmission line, which finds considerable application at microwave frequencies, is *coaxial cable* (Figs. 3-1F through 3-1L). This form of line consists of two cylindrical conductors sharing the same axis (hence “coaxial”), and separated by a dielectric (Fig. 3-1F). For low frequencies (in flexible cables) the dielectric may be polyethylene or polyethylene foam, but at higher frequencies *Teflon* and other materials are used. Also used, in some applications, are dry air and dry nitrogen.

3-1E Shielded twin-lead transmission line.



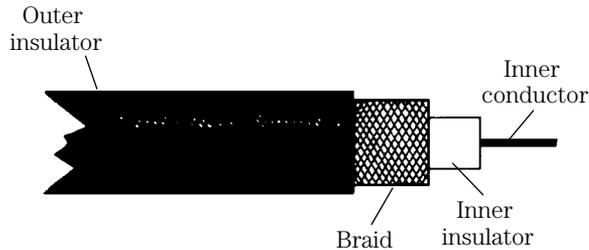
3-1F Coaxial cable (end view).



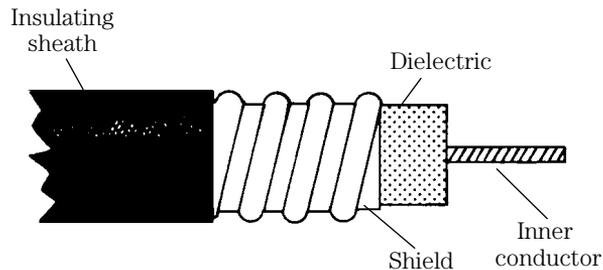
Several forms of coaxial line are available. Flexible coaxial cable is perhaps the most common form. The outer conductor in such cable is made of either braid or foil (Fig. 3-1G). Television broadcast receiver antennas provide an example of such cable from common experience. Another form of flexible or semiflexible coaxial line is *helical line* (Fig. 3-1H) in which the outer conductor is spiral wound. *Hardline* (Fig. 3-1I) is coaxial cable that uses a thin-wall pipe as the outer conductor. Some hardline coax used at microwave frequencies has a rigid outer conductor and a solid dielectric.

*Gas-filled line* is a special case of hardline that is hollow (Fig. 3-1J), the center conductor is supported by a series of thin ceramic or Teflon insulators. The dielectric is either anhydrous (i.e., dry) nitrogen or some other inert gas.

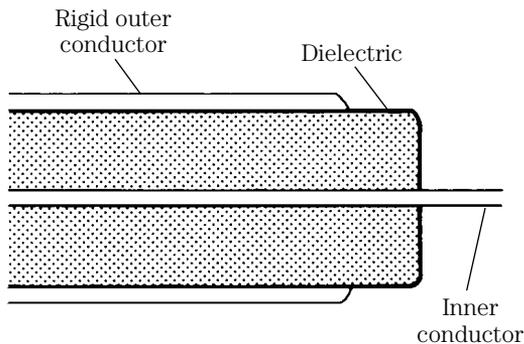
Some flexible microwave coaxial cable uses a solid “air-articulated” dielectric (Fig. 3-1K), in which the inner insulator is not continuous around the center conductor, but rather is ridged. Reduced dielectric losses increase the usefulness of the



**3-1G** Coaxial cable (side view).

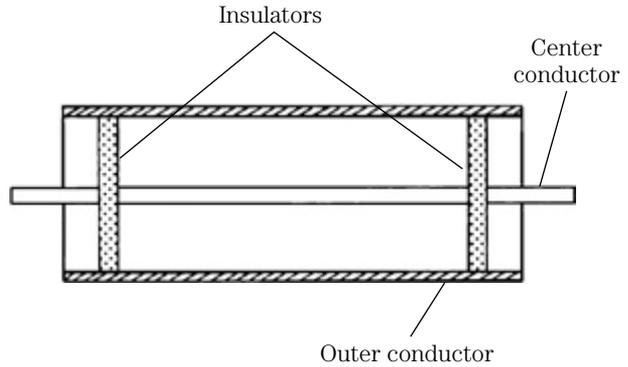


**3-1H** Coaxial hardline cable.

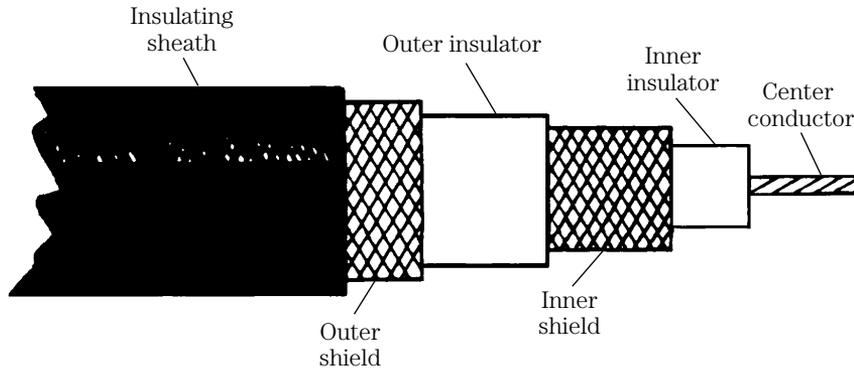
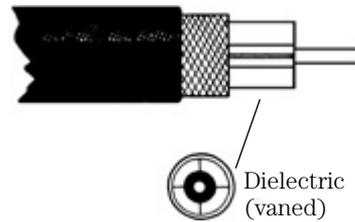


**3-1I** Rigid coaxial line.

3-1J Gas-filled hollow coaxial line.



3-1K Articulated coaxial line.



3-1L Double-shielded coaxial line.

cable at higher frequencies. Double-shielded coaxial cable (Fig. 3-1L) provides an extra measure of protection against radiation from the line, and EMI from outside sources, from getting into the system.

*Stripline*, also called *microstripline* (Fig. 3-1M), is a form of transmission line used at high UHF and microwave frequencies. The stripline consists of a critically sized conductor over a ground-plane conductor, and separated from it by a dielectric. Some striplines are sandwiched between two groundplanes and are separated from each by a dielectric.

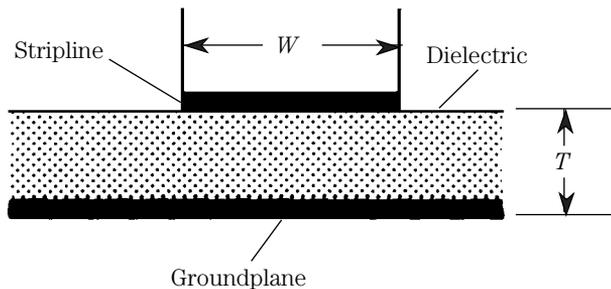
## Transmission line characteristic impedance ( $Z_o$ )

The transmission line is an  $RLC$  network (see Fig. 3-2), so it has a *characteristic impedance*  $Z_o$ , also sometimes called a *surge impedance*. Network analysis will show that  $Z_o$  is a function of the per unit of length parameters *resistance*  $R$ , *conductance*  $G$ , *inductance*  $L$ , and *capacitance*  $C$ , and is found from

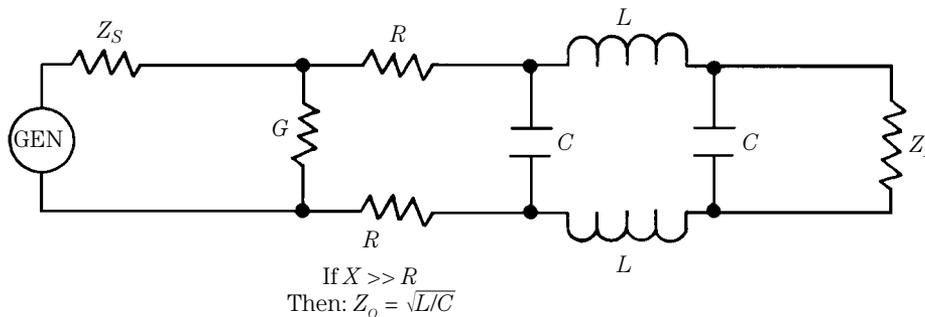
$$Z_o = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \quad [3.1]$$

where

- $Z_o$  is the characteristic impedance, in ohms
- $R$  is the resistance per unit length, in ohms
- $G$  is the conductance per unit length, in mhos
- $L$  is the inductance per unit length, in henrys
- $C$  is the capacitance per unit length, in farads
- $\omega$  is the angular frequency in radians per second ( $2\pi F$ )



3-1M Stripline transmission line.



3-2 Equivalent circuit of transmission line.

In microwave systems the resistances are typically very low compared with the reactances, so Eq. 3.1 can be reduced to the simplified form:

$$Z_o = \sqrt{\frac{L}{C}} \quad [3.2]$$

**Example 3-1** A nearly lossless transmission line ( $R$  is very small) has a unit length inductance of 3.75 nH and a unit length capacitance of 1.5 pF. Find the characteristic impedance  $Z_o$ .

**Solution:**

$$\begin{aligned} Z_o &= \sqrt{\frac{L}{C}} \\ &= \sqrt{\frac{\left(3.75 \text{ nH} \times \frac{1 \text{ H}}{10^9 \text{ nH}}\right)}{\left(1.5 \text{ pF} \times \frac{1 \text{ F}}{10^{12} \text{ pF}}\right)}} \\ &= \sqrt{\frac{3.75 \times 10^{-9} \text{ H}}{1.5 \times 10^{-12} \text{ F}}} \\ &= \sqrt{2.5 \times 10^3} = 50 \Omega \end{aligned}$$

The characteristic impedance for a specific type of line is a function of the conductor size, the conductor spacing, the conductor geometry (see again Fig. 3-1), and the dielectric constant of the insulating material used between the conductors. The dielectric constant  $e$  is equal to the reciprocal of the velocity (squared) of the wave when a specific medium is used:

$$e = \frac{1}{v^2} \quad [3.3]$$

where

$e$  is the dielectric constant (for a perfect vacuum  $e = 1.000$ )  
 $v$  is the velocity of the wave in the medium

(a) *Parallel line*

$$Z_o = \frac{276}{\sqrt{e}} \log \left( \frac{2S}{d} \right) \quad [3.4]$$

where

$Z_o$  is the characteristic impedance, in ohms  
 $e$  is the dielectric constant  
 $S$  is the center-to-center spacing of the conductors  
 $d$  is the diameter of the conductors

(b) *Coaxial line*

$$Z_o = \frac{138}{\sqrt{e}} \log \left( \frac{D}{d} \right) \quad [3.5]$$

where

$D$  is the diameter of the outer conductor  
 $d$  is the diameter of the inner conductor

(c) *Shielded parallel line*

$$Z_o = \frac{276}{\sqrt{e}} \log \left( 2A \frac{(1 - B^2)}{(1 + B^2)} \right) \quad [3.6]$$

where

$A = s/d$   
 $B = s/D$

(d) *Stripline*

$$Z_o = \frac{377}{\sqrt{e_t}} \left( \frac{T}{W} \right) \quad [3.7A]$$

where

$e_t$  is the relative dielectric constant of the printed wiring board (PWB)  
 $T$  is the thickness of the printed wiring board  
 $W$  is the width of the stripline conductor

The relative dielectric constant  $e_t$  used above differs from the normal dielectric constant of the material used in the PWB. The relative and normal dielectric constants move closer together for larger values of the ratio  $W/T$ .

**Example 3-2** A stripline transmission line is built on a 4-mm-thick printed wiring board that has a relative dielectric constant of 5.5. Calculate the characteristic impedance if the width of the strip is 2 mm.

**Solution:**

$$Z_o = \frac{377}{\sqrt{e_t}} \left( \frac{T}{W} \right)$$

$$\begin{aligned}
 &= \frac{377}{\sqrt{5.5}} \left( \frac{4 \text{ mm}}{2 \text{ mm}} \right) \\
 &= \frac{377}{2.35} (2) = 321 \ \Omega
 \end{aligned}$$

In practical situations, we usually don't need to calculate the characteristic impedance of a stripline, but rather design the line to fit a specific system impedance (e.g., 50  $\Omega$ ). We can make some choices of printed circuit material (hence dielectric constant) and thickness, but even these are usually limited in practice by the availability of standardized boards. Thus, stripline *width* is the variable parameter. Equation 3.2 can be arranged to the form:

$$W = \frac{377 T}{Z_o \sqrt{e}} \quad \text{[3.7B]}$$

The impedance of 50  $\Omega$  is accepted as standard for RF systems, except in the cable TV industry. The reason for this diversity is that power handling ability and low loss operation don't occur at the same characteristic impedance. For example, the maximum power handling ability for coaxial cables occurs at 30  $\Omega$ , while the lowest loss occurs at 77  $\Omega$ ; 50  $\Omega$  is therefore a reasonable tradeoff between the two points. In the cable TV industry, however, the RF power levels are minuscule, but lines are long. The tradeoff for TV is to use 75  $\Omega$  as the standard system impedance in order to take advantage of the reduced attenuation factor.

## Transmission line characteristics

### Velocity factor

In the section preceding this section, we discovered that the velocity of the wave (or signal) in the transmission line is less than the free-space velocity (i.e., less than the speed of light). Further, we discovered in Eq. 3.3 that velocity is related to the dielectric constant of the insulating material that separates the conductors in the transmission line. Velocity factor  $v$  is usually specified as a decimal fraction of  $c$ , the speed of light ( $3 \times 10^8$  m/s). For example, if the velocity factor of a transmission line is rated at "0.66," then the velocity of the wave is  $0.66c$ , or  $(0.66) (3 \times 10^8 \text{ m/s}) = 1.98 \times 10^8 \text{ m/s}$ .

Velocity factor becomes important when designing things like transmission line transformers, or any other device in which the length of the line is important. In most cases, the transmission line length is specified in terms of *electrical length*, which can be either an angular measurement (e.g., 180° or  $\pi$  radians), or a relative measure keyed to wavelength (e.g., one-half wavelength, which is the same as 180°). The *physical length* of the line is longer than the equivalent electrical length. For example, let's consider a 1-GHz half-wavelength transmission line.

A rule of thumb tells us that the length of a wave (in meters) in free space is  $0.30/F$ , where frequency  $F$  is expressed in gigahertz; therefore, a half-wavelength line is  $0.15/F$ .

At 1 GHz, the line must be  $0.15 \text{ m}/1 \text{ GHz} = 0.15 \text{ m}$ . If the velocity factor is 0.80, then the *physical length* of the transmission line that will achieve the desired *electrical length* is  $[(0.15 \text{ m}) (v)]/F = [(0.15 \text{ m}) (0.80)]/1 \text{ GHz} = 0.12 \text{ m}$ . The derivation of the rule of thumb is “left as an exercise for the student.” (*Hint*: It comes from the relationship between wavelength, frequency, and velocity of propagation for any form of wave.)

There are certain practical considerations regarding velocity factor that result from the fact that the physical and electrical lengths are not equal. For example, in a certain type of phased-array antenna design, radiating elements are spaced a half-wavelength apart, and must be fed  $180^\circ$  (half-wave) out of phase with each other. The simplest interconnect is to use a half-wave transmission line between the  $0^\circ$  element and the  $180^\circ$  element. According to the standard wisdom, the transmission line will create the  $180^\circ$  phase delay required for the correct operation of the antenna. Unfortunately, because of the velocity factor, the physical length for a one-half electrical wavelength cable is shorter than the free-space half-wave distance between elements. In other words, the cable will be too short to reach between the radiating elements by the amount of the velocity factor!

Clearly, velocity factor is a topic that must be understood before transmission lines can be used in practical situations. Table 3-1 shows the velocity factors for several types of popular transmission line. Because these are nominal values, the actual velocity factor for any given line should be measured.

**Table 3-1. Transmission line characteristics**

Type of line	$Z_o$ (ohms)	Velocity factor
		$v$
½-in. TV parallel line (air dielectric)	300	0.95
1-in. TV parallel line (air dielectric)	450	0.95
TV twin lead	300	0.82
UHF TV twin lead	300	0.80
Polyethylene coaxial cable	*	0.66
Polyethylene foam coaxial cable	*	0.79
Air-space polyethylene foam coaxial cable	*	0.86
Teflon	*	0.70

\* Various impedances depending upon cable type.

### Transmission line noise

Transmission lines are capable of generating noise and spurious voltages that are seen by the system as valid signals. Several such sources exist. One source is the coupling between noise currents flowing in the outer conductor and the inner conductor. Such currents are induced by nearby electromagnetic interference and other sources (e.g., connection to a noisy groundplane). Although coaxial design reduces noise pickup, compared with parallel line, the potential for EMI exists. Selection of high-grade line, with a high degree of shielding, reduces the problem.

Another source of noise is thermal noises in the resistances and conductances. This type of noise is proportional to resistance and temperature.

There is also noise created by mechanical movement of the cable. One species results from the movement of the dielectric against the two conductors. This form of noise is caused by electrostatic discharges in much the same manner as the spark created by rubbing a piece of plastic against woolen cloth.

A second species of mechanically generated noise is *piezoelectricity* in the dielectric. Although more common in cheap cables, one should be aware of it. Mechanical deformation of the dielectric causes electrical potentials to be generated.

Both species of mechanically generated noise can be reduced or eliminated by proper mounting of the cable. Although rarely a problem at lower frequencies, such noise can be significant at microwave frequencies when signals are low.

### Coaxial cable capacitance

A coaxial transmission line possesses a certain capacitance per unit of length. This capacitance is defined by

$$C = \frac{24e}{\log (D/d)} \frac{\text{pF}}{\text{m}} \quad [3.8A]$$

A long run of coaxial cable can build up a large capacitance. For example, a common type of coax is rated at 65 pF/m. A 150-m roll thus has a capacitance of 65 pF/m  $\times$  (150 m), or 9750 pF. When charged with a high voltage, as is done in breakdown voltage tests at the factory, the cable *acts like a charged high-voltage capacitor*. Although rarely (if ever) lethal to humans, the stored voltage in new cable can deliver a nasty electrical shock and can irreparably damage electronic components.

### Coaxial cable cutoff frequency $F_c$

The normal mode in which a coaxial cable propagates a signal is as a transverse electromagnetic (TEM) wave, but others are possible—and usually undesirable. There is a maximum frequency above which TEM propagation becomes a problem, and higher modes dominate. Coaxial cable should *not* be used above a frequency of

$$F = \frac{6.75}{(D + d) \sqrt{e}} \text{ GHz} \quad [3.8B]$$

where

- $F$  is the TEM-mode cutoff frequency
- $D$  is the diameter of the outer conductor, in inches
- $d$  is the diameter of the inner conductor, in inches
- $e$  is the dielectric constant

When maximum operating frequencies for cable are listed, it is the TEM mode that is cited. Beware of attenuation, however, when making selections for microwave frequencies. A particular cable may have a sufficiently high TEM-mode frequency, but still exhibit a high attenuation per unit length at X or Ku bands.

## Transmission line responses

In order to understand the operation of transmission lines, we need to consider two cases: *step-function response* and the *steady-state ac response*. The step-function case involves a single event when a voltage at the input of the line snaps from zero (or a steady value) to a new (or nonzero) value, and remains there until all action dies out. This response tells us something of the behavior of pulses in the line, and in fact is used to describe the response to a single-pulse stimulus. The steady-state ac response tells us something of the behavior of the line under stimulation by a sinusoidal RF signal.

### Step-function response of a transmission line

Figure 3-3 shows a parallel transmission line with characteristic impedance  $Z_o$  connected to a load impedance  $Z_L$ . The generator at the input of the line consists of a voltage source  $V$  in series with a source impedance  $Z_s$  and a switch  $S_1$ . Assume for the present that all impedances are pure resistances (i.e.,  $R + j0$ ). Also, assume that  $Z_s = Z_o$ .

When the switch is closed at time  $T_o$  (Fig. 3-4A), the voltage at the input of the line ( $V_{in}$ ) jumps to  $V/2$ . In Fig. 3-2, you may have noticed that the  $LC$  circuit resembles a delay line circuit. As might be expected, therefore, the voltage wavefront propagates along the line at a velocity  $v$  of:

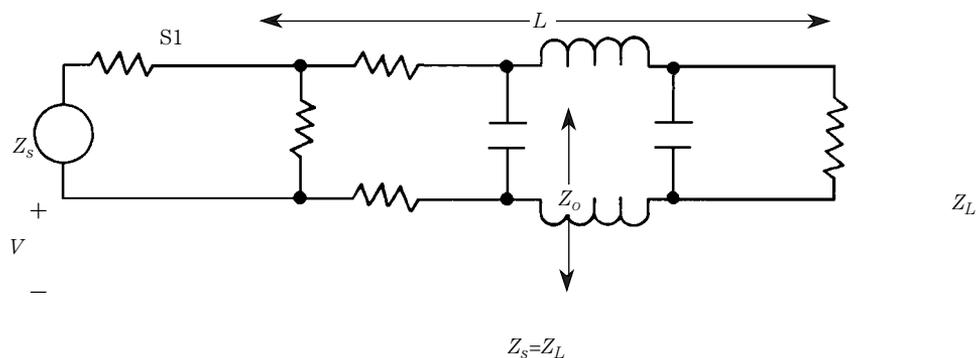
$$v = \frac{1}{\sqrt{LC}} \quad [3.9]$$

where

$v$  is the velocity, in meters per second

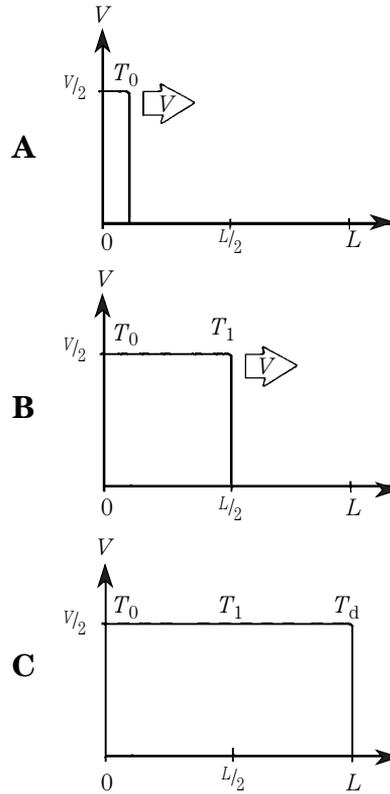
$L$  is the inductance, in henrys

$C$  is the capacitance, in farads



3-3 Schematic example of transmission line.

3-4 Step-function propagation along transmission line at three points.



At time  $T_1$  (Fig. 3-4B), the wavefront has propagated one-half the distance  $L$ , and by  $T_d$  it has propagated the entire length of the cable (Fig. 3-4C).

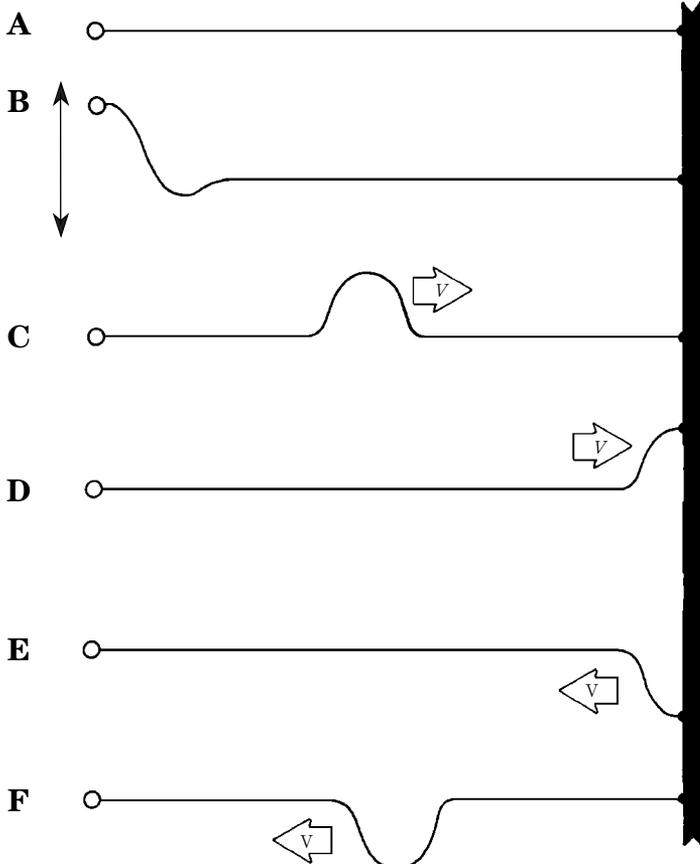
If the load is perfectly matched (i.e.,  $Z_L = Z_o$ ), then the load absorbs the wave and no component is reflected. But in a mismatched system ( $Z_L$  is not equal to  $Z_o$ ), a portion of the wave is reflected back down the line toward the generator.

Figure 3-5 shows the rope analogy for reflected pulses in a transmission line. A taut rope (Fig. 3-5A) is tied to a rigid wall that does not absorb any of the energy in the pulse propagated down the rope. When the free end of the rope is given a vertical displacement (Fig. 3-5B), a wave is propagated down the rope at velocity  $v$  (Fig. 3-5C). When the pulse hits the wall (Fig. 3-5D), it is reflected (Fig. 3-5E) and propagates back down the rope toward the free end (Fig. 3-5F).

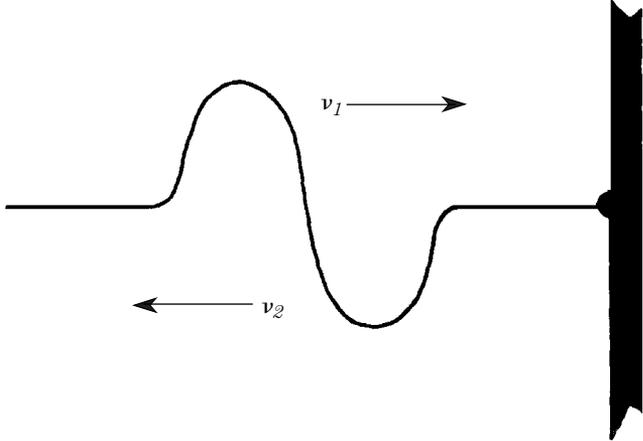
If a second pulse is propagated down the line before the first pulse dies out, then there will be two pulses on the line at the same time (Fig. 3-6A). When the two pulses interfere, the resultant will be the algebraic sum of the two. In the event that a pulse train is applied to the line, the interference pattern will set up *standing waves*, an example of which is shown in Fig. 3-6B.

### Reflection coefficient

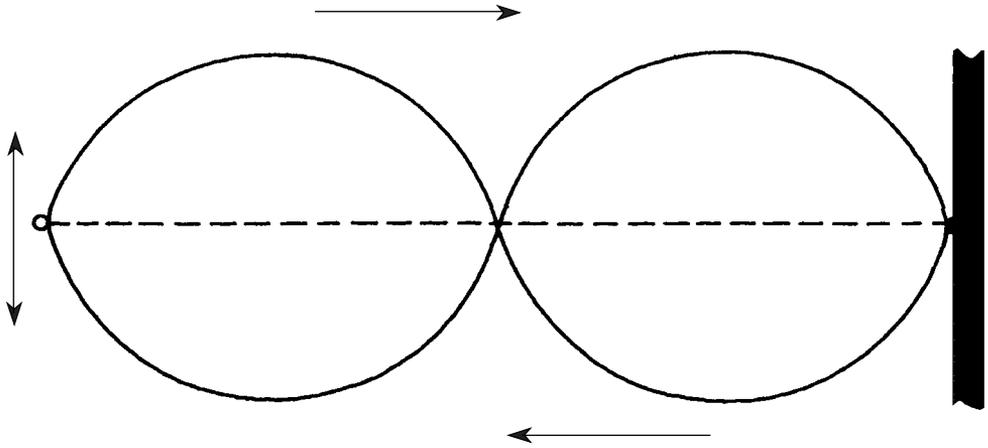
The *reflection coefficient*  $\Gamma$  of a circuit containing a transmission line and load impedance is a measure of how well the system is matched. The absolute value of the re-



3-5 Rope analogy to transmission line.



3-6A Interfering opposite waves.



3-6B Standing waves.

reflection coefficient varies from  $-1$  to  $+1$ , depending upon the magnitude of reflection;  $\Gamma = 0$  indicates a perfect match with no reflection, while  $-1$  indicates a short-circuited load, and  $+1$  indicates an open circuit. To understand the reflection coefficient, let's start with a basic definition of the resistive load impedance  $Z = R + j0$ :

$$Z_L = \frac{V}{I} \quad [3.10]$$

where

$Z_L$  is the load impedance  $R + j0$   
 $V$  is the voltage across the load  
 $I$  is the current flowing in the load

Because there are both reflected and incident waves, we find that  $V$  and  $I$  are actually the sum of incident and reflected voltages and currents, respectively. Therefore:

$$Z_L = \frac{V}{I} \quad [3.11A]$$

$$Z_L = \frac{V_{\text{inc}} + V_{\text{ref}}}{I_{\text{inc}} + I_{\text{ref}}} \quad [3.11B]$$

where

$V_{\text{inc}}$  is the incident (i.e., forward) voltage  
 $V_{\text{ref}}$  is the reflected voltage  
 $I_{\text{inc}}$  is the incident current  
 $I_{\text{ref}}$  is the reflected current

## 74 *Transmission lines*

Because of Ohm's law, you can define the currents in terms of voltage, current, and the characteristic impedance of the line:

$$I_{\text{inc}} = \frac{V_{\text{inc}}}{Z_o} \quad [3.12]$$

and

$$I_{\text{inc}} = \frac{-V_{\text{ref}}}{Z_o} \quad [3.13]$$

(The minus sign in Eq. 3.13 indicates that a direction reversal has taken place.)

The two expressions for current (Eqs. 3.12 and 3.13) may be substituted into Eq. 3.11 to yield

$$Z_L = \frac{V_{\text{inc}} + V_{\text{ref}}}{\frac{V_{\text{inc}}}{Z_o} - \frac{V_{\text{ref}}}{Z_o}} \quad [3.14]$$

The reflection coefficient  $\Gamma$  is defined as the ratio of reflected voltage to incident voltage:

$$\Gamma = \frac{V_{\text{ref}}}{V_{\text{inc}}} \quad [3.15]$$

Using this ratio in Eq. 3.14 gives

$$\Gamma = \frac{Z_L - Z_o}{Z_L + Z_o} \quad [3.16]$$

**Example 3-3** A 50- $\Omega$  transmission line is connected to a 30- $\Omega$  resistive load. Calculate the reflection coefficient  $\Gamma$ .

**Solution:**

$$\begin{aligned} \Gamma &= \frac{Z_L - Z_o}{Z_L + Z_o} \\ &= \frac{(50 \Omega) - (30 \Omega)}{(50 \Omega) + (30 \Omega)} \\ &= \frac{20}{80} = 0.25 \end{aligned}$$

**Example 3-4** In Example 3-3, the incident voltage is 3 V rms. Calculate the reflected voltage.

**Solution:**

If

$$\Gamma = \frac{V_{\text{ref}}}{V_{\text{inc}}}$$

then

$$\begin{aligned} V_{\text{ref}} &= \Gamma V_{\text{inc}} \\ &= (0.25) (3 \text{ V}) = 0.75 \text{ V} \end{aligned}$$

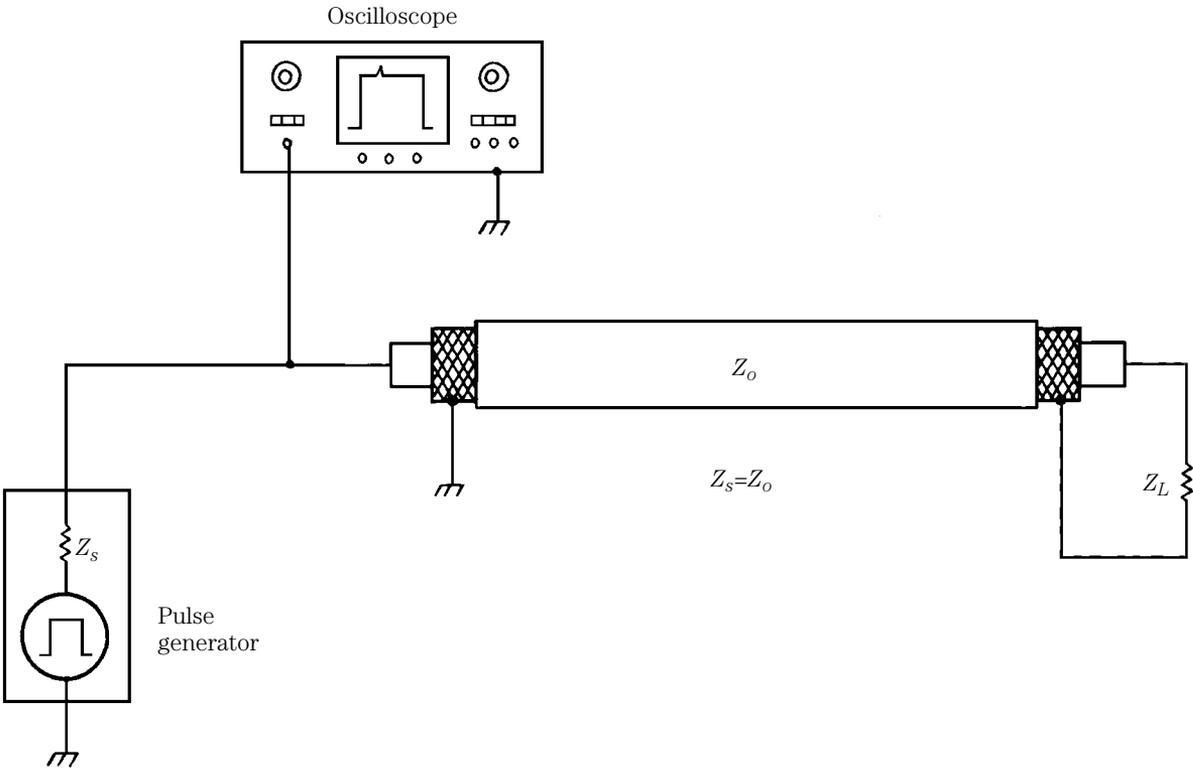
The phase of the reflected signal is determined by the relationship of load impedance and transmission line characteristic impedance. For resistive loads ( $Z = R + j0$ ): if the ratio  $Z_L/Z_o$  is 1.0, then there is no reflection; if  $Z_L/Z_o$  is less than 1.0, then the reflected signal is  $180^\circ$  out of phase with the incident signal; if the ratio  $Z_L/Z_o$  is greater than 1.0 then the reflected signal is in phase with the incident signal. In summary:

Ratio	Angle of reflection
$Z_L/Z_o = 1$	No reflection
$Z_L/Z_o < 1$	$180^\circ$
$Z_L/Z_o > 1$	$0^\circ$

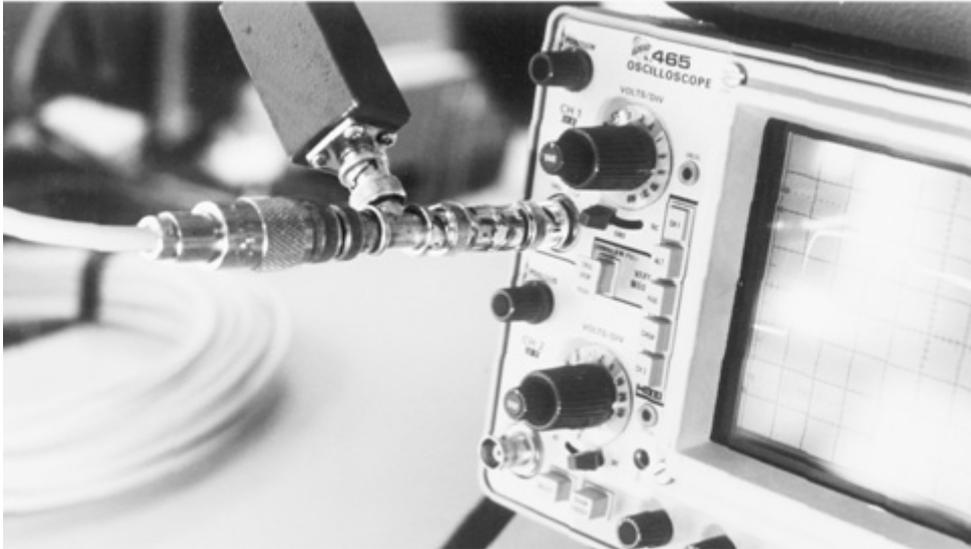
The step-function (or pulse) response of the transmission line leads to a powerful means of analyzing the line, and its load, on an oscilloscope. Figure 3-7A shows (in schematic form) the test set-up for *time domain reflectometry* (TDR) measurements. An oscilloscope and a pulse (or square-wave) generator are connected in parallel across the input end of the transmission line. Figure 3-7B shows a pulse test jig built by the author for testing lines at HF. The small shielded box contains a TTL square-wave oscillator circuit. Although a crystal oscillator can be used, an  $RC$  timed circuit running close to 1000 kHz is sufficient. In Fig. 3-7B, you can see the test pulse generator box is connected in parallel with the cable under test and the input of the oscilloscope. A closer look is seen in Fig. 3-7C. A BNC “tee” connector and a double male BNC adapter are used to interconnect the box with the ‘scope.

If a periodic waveform is supplied by the generator, then the display on the oscilloscope will represent the sum of reflected and incident pulses. The duration of the pulse (i.e., pulse width), or one-half the period of the square wave, is adjusted so that the returning reflected pulse arrives approximately in the center of the incident pulse.

Figure 3-8 shows a TDR display under several circumstances. Approximately 30 m of coaxial cable, with a velocity factor of 0.66, was used in a test setup similar to Fig. 3-7. The pulse width was approximately 0.9 microseconds ( $\mu\text{s}$ ). The hori-

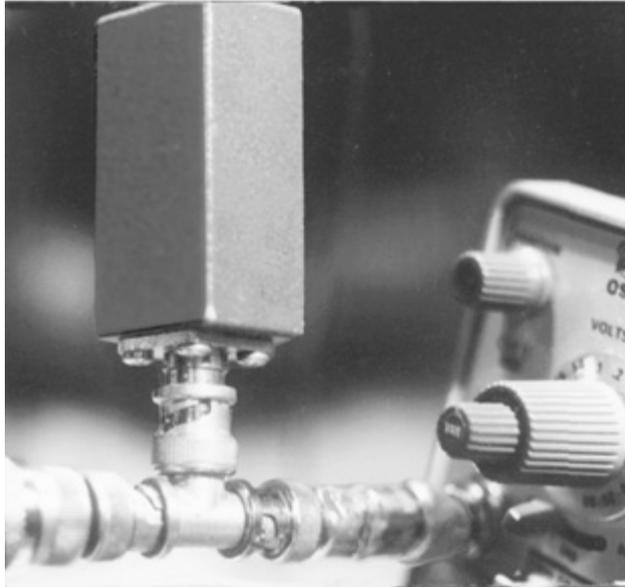


3-7A Time domain reflectometry setup.

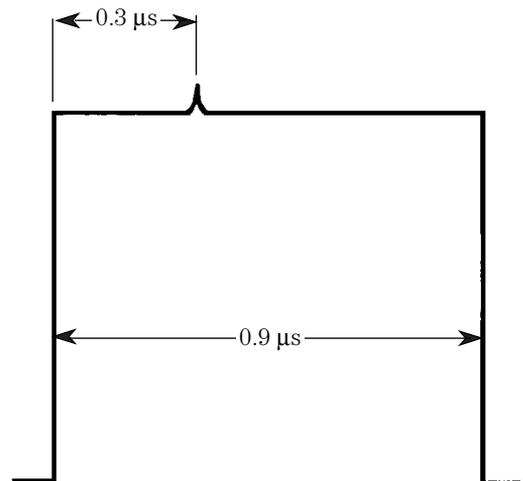


3-7B Test setup for impromptu time domain reflectometry.

**3-7C** Close-up of RF connections.

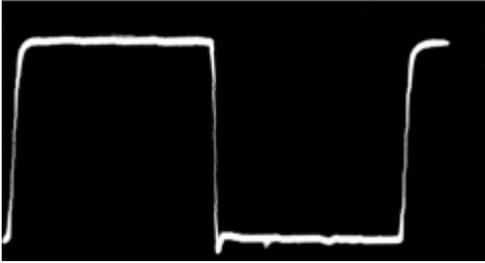


**3-8A** Idealized TDR pulse. Small “pip” on top is reflected signal interfering with forward pulse.



horizontal sweep time on the 'scope was adjusted to show only one pulse—which, in this case, represented one-half of a 550-kHz square wave (Fig. 3-8B).

The displayed trace in Fig. 3-8B shows the pattern when the load is matched to the line ( $Z_L = Z_o$ ). A slight discontinuity exists on the high side of the pulse, and this represents a small reflected wave. Even though the load and line were supposedly matched, the connectors at the end of the line presented a slight impedance discontinuity that shows up on the 'scope as a reflected wave. In general, any discontinuity in the line, any damage to the line, any too-sharp bend, or other anomaly, causes a slight impedance variation, and hence a reflection.



**3-8B** TDR pulse with no significant reflection.

Notice that the anomaly occurs approximately one-third of the 0.9- $\mu\text{s}$  duration (or 0.3  $\mu\text{s}$ ) after the onset of the pulse. This fact tells us that the reflected wave arrives back at the source 0.3  $\mu\text{s}$  after the incident wave leaves. Because this time period represents a round-trip, you can conclude that the wave required 0.3  $\mu\text{s}/2$ , or 0.15  $\mu\text{s}$  to propagate the length of the line. Knowing that the velocity factor is 0.66 for that type of line, you can calculate its approximate length:

$$\begin{aligned} \text{Length} &= cvT && \text{[3.17]} \\ &= \frac{(3 \times 10^8 \text{ m/s})}{s} \times (0.66) \times (1.5 \times 10^{-7} \text{ s}) \\ &= 29.7 \text{ m} \end{aligned}$$

which agrees “within experimental accuracy” with the 30 m actual length prepared for the experiment ahead of time. Thus, the TDR setup (or a TDR instrument) can be used to measure the length of a transmission line. A general equation is

$$L_{\text{meters}} = \frac{cvT_d}{2} \quad \text{[3.18]}$$

where

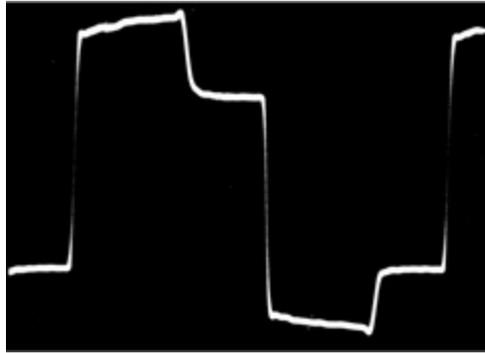
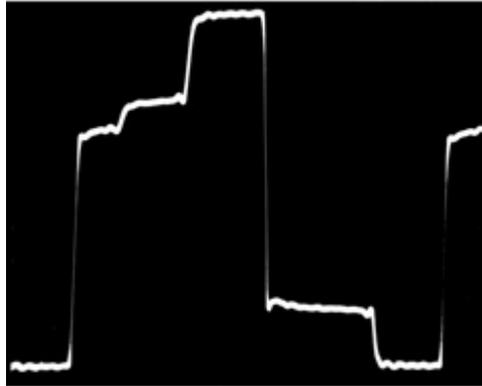
$L$  is the length in meters

$c$  is the velocity of light ( $3 \times 10^8$  m/s)

$v$  is the velocity factor of the transmission line

$T_d$  is the round-trip time between the onset of the pulse, and the first reflection

Figures 3-8C through 3-8H show the behavior of the transmission line to the stepfunction when the load impedance is mismatched to the transmission line ( $Z_L$  not equal to  $Z_o$ ). Figure 3-8C shows what happens when the load impedance is less than the line impedance (in this case,  $0.5Z_o$ ). The reflected wave is inverted, and

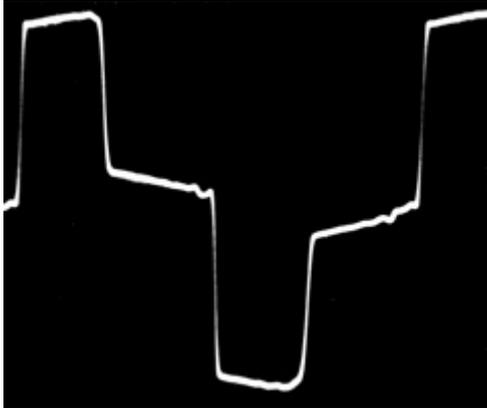
3-8C  $Z_L < Z_o$ .3-8D  $Z_L > Z_o$ .

sums with the incident wave along the top of the pulse. The reflection coefficient can be determined by examining the relative amplitudes of the two waves.

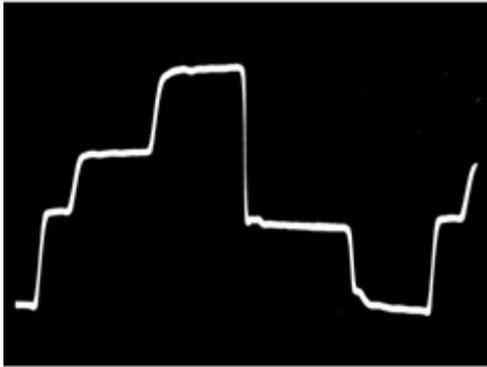
The opposite situation, in which  $Z_L$  is  $2Z_o$ , is shown in Fig. 3-8D. In this case, the reflected wave is in phase with the incident wave, so it adds to the incident wave as shown. The cases for a short-circuited load and an open-circuited load are shown in Figs. 3-8E and 3-8F, respectively. The cases of reactive loads are shown in Figs. 3-8G and 3-8H. The waveform in Fig. 3-8G resulted from a capacitance in series with a  $50\text{-}\Omega$  (matched) resistance; the waveform in Fig. 3-8H resulted from a  $50\text{-}\Omega$  resistance in series with an inductance.

### The ac response of the transmission line

When a CW RF signal is applied to a transmission line, the excitation is sinusoidal (Fig. 3-9), so it becomes useful for us to investigate the steady-state ac response of the line. The term *steady-state* implies a sine wave of constant amplitude, phase, and



3-8E  $Z_L = 0$ .

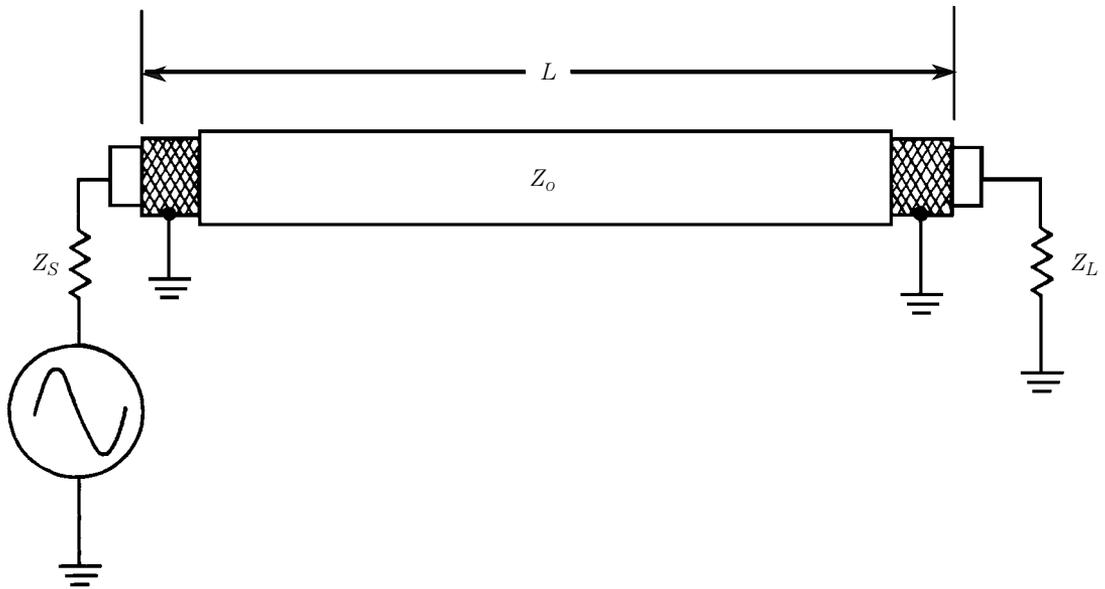
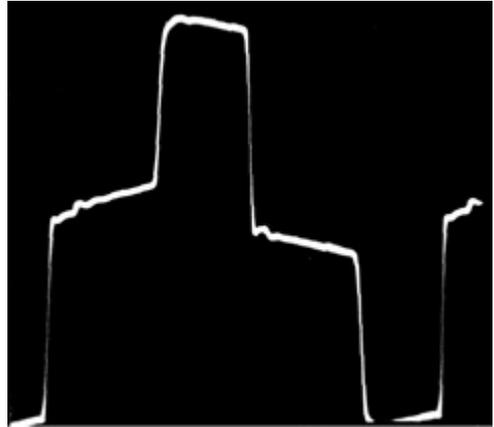


3-8F  $Z_L = \infty$ .



3-8G  $Z_L = 50 - jX_C$ .

3-8H  $Z_L = 50 + jX_L$



3-9 AC-excited transmission line.

frequency. When ac is applied to the input of the line, it propagates along the line at a given velocity. The ac signal amplitude and phase will decay exponentially:

$$V_R = Ve^{-\gamma l} \quad [3.19]$$

where

$V_R$  is the voltage received at the far end of the line

$V$  is the applied voltage

$l$  is the length of the line

$y$  is the *propagation constant* of the line

The propagation constant  $y$  is defined in various equivalent ways, each of which serves to illustrate its nature. For example, the propagation constant is proportional to the product of impedance and admittance characteristics of the line:

$$y = \sqrt{ZY} \quad [3.20]$$

or, since  $Z = R + j\omega L$  and  $Y = G + j\omega C$ , we may write:

$$y = \sqrt{(R + j\omega l)(G + j\omega C)} \quad [3.21]$$

You can also write an expression for the propagation constant in terms of the *line attenuation constant*  $a$  and *phase constant*  $B$ :

$$y = a + jB \quad [3.22]$$

If you can assume that susceptance dominates conductance in the admittance term, and reactance dominates resistance in the impedance term (both usually true at microwave frequencies), then we may neglect the  $R$  and  $G$  terms altogether and write:

$$y = j\omega \sqrt{LC} \quad [3.23]$$

We may also reduce the phase constant  $B$  to

$$B = \omega \sqrt{LC} \quad [3.24]$$

or

$$B = \omega Z_0 C \quad \text{rad/m} \quad [3.25]$$

and, of course, the characteristic impedance remains:

$$Z_0 = \sqrt{LC} \quad [3.26]$$

### Special cases

The impedance “looking into” a transmission line ( $Z$ ) is the impedance presented to the source by the combination of load impedance and transmission line characteristic impedance. Below are presented equations that define the looking-in impedance seen by a generator (or source) driving a transmission line.

The case where the load impedance, and line characteristic impedance, are matched is defined by

$$Z_L = R_L + j0 = Z_o$$

In other words, the load impedance is resistive and equal to the characteristic impedance of the transmission line. In this case, the line and load are matched, and the impedance looking in will be a simple  $Z = Z_L = Z_o$ . In other cases, however, we find different situations where  $Z_L$  is not equal to  $Z_o$ .

**1.  $Z_L$  is not equal to  $Z_o$  in a random-length lossy line:**

$$Z = (Z_o) \left[ \frac{Z_L + Z_o \tanh (yl)}{Z_o + Z_L \tanh (yl)} \right] \quad [3.27]$$

where

$Z$  is the impedance looking in, in ohms

$Z_L$  is the load impedance, in ohms

$Z_o$  is the line characteristic impedance, in ohms

$l$  is the length of the line, in meters

$y$  is the propagation constant

**2.  $Z_L$  not equal to  $Z_o$  in a lossless, or very low loss, random-length line:**

$$Z = (Z_o) \left[ \frac{Z_L + jZ_o \tan (Bl)}{Z_o + jZ_L \tan (Bl)} \right] \quad [3.28]$$

Equations 3.27 and 3.28 serve for lines of any random length. For lines that are either integer multiples of a half-wavelength, or odd-integer (i.e., 1, 3, 5, 7, . . . , etc.) multiples of a quarter-wavelength, special solutions for these equations are found—and some of these solutions are very useful in practical situations. For example, consider . . .

**3. Half-wavelength lossy lines:**

$$Z = (Z_o) \left[ \frac{Z_L + Z_o \tanh (al)}{Z_o + Z_L \tanh (al)} \right] \quad [3.29]$$

**Example 3-5** A lossless 50- $\Omega$  ( $Z_o$ ) transmission line is exactly one-half wavelength long and is terminated in a load impedance of  $Z = 30 + j0$ . Calculate the input impedance looking into the line. (*Note:* in a lossless line  $a = 0$ .)

**Solution:**

$$Z = (Z_o) \left[ \frac{Z_L + Z_o \tanh (al)}{Z_o + Z_L \tanh (al)} \right]$$

$$\begin{aligned}
Z &= (50 \Omega) \left\{ \frac{30 + 50 \tanh [(0) (\pi)]}{50 + 30 \tanh [(0) (\pi)]} \right\} \\
&= (50 \Omega) \left[ \frac{30 + 50 \tanh (0)}{50 + 30 \tanh (0)} \right] \\
&= (50 \Omega) \left( \frac{30 + 0}{50 + 0} \right) \\
&= (50 \Omega) (30/50) = 30 \Omega
\end{aligned}$$

In Example 3-5 we discovered that the impedance looking into a lossless (or very low loss) half-wavelength transmission line is the load impedance:

$$Z = Z_L \quad \text{[3.30]}$$

The fact that line input impedance equals load impedance, is very useful in certain practical situations. For example, a resistive impedance is not changed by the line length. Therefore, when an impedance is inaccessible for measurement purposes, the impedance can be measured through a transmission line that is an integer multiple of a half wavelength.

Our next special case involves a quarter-wavelength transmission line, and those that are *odd* integer multiples of quarter-wavelengths (of course, *even* integer multiples of a quarter wavelength obey the half-wavelength criteria).

#### 4. Quarter-wavelength lossy lines:

$$Z = (Z_o) \left[ \frac{Z_L + Z_o \coth (al)}{Z_o + Z_L \coth (al)} \right] \quad \text{[3.31]}$$

and

#### 5. Quarter-wavelength lossless or very low loss lines:

$$Z = \frac{(Z_o)^2}{Z_L} \quad \text{[3.32]}$$

From Eq. 3.32, you can discover an interesting property of the quarter-wavelength transmission line. First, divide each side of the equation by  $Z_o$ :

$$\frac{Z}{Z_o} = \frac{[Z_o]^2}{Z_L Z_o} \quad \text{[3.33]}$$

$$= \frac{Z_o}{Z_L} \quad \text{[3.34]}$$

The ratio  $Z/Z_o$  shows an inversion of the load impedance ratio  $Z_L/Z_o$ , or, stated another way,

$$\frac{Z}{Z_o} = \frac{1}{Z_L/Z_o} \quad [3.35]$$

Again, from Eq. 3.32, you can deduce another truth about quarter-wavelength transmission lines:

If

$$Z = \frac{\sqrt{Z_o}}{Z_L} \quad [3.36]$$

then

$$ZZ_L = \sqrt{Z_o} \quad [3.37]$$

which means

$$Z_o = \sqrt{ZZ_L} \quad [3.38]$$

Equation 3.38 shows that a quarter-wavelength transmission line can be used as an *impedance matching network*. Called a *Q section*, the quarter-wavelength transmission line used for impedance matching requires a characteristic impedance  $Z_o$ —if  $Z$  is the source impedance and  $Z_L$  is the load impedance.

**Example 3-6** A  $50\text{-}\Omega$  source must be matched to a load impedance of  $36\ \Omega$ . Find the characteristic impedance required of a Q section matching network.

**Solution:**

$$\begin{aligned} Z &= \sqrt{ZZ_L} \\ &= \sqrt{(50\ \Omega)(36\ \Omega)} \\ &= \sqrt{1800\ \Omega^2} = 42\ \Omega \end{aligned}$$

**6. Transmission line as a reactance:** Reconsider Eq. 3.28, which related impedance looking in to load impedance and line length:

$$Z = (Z_o) \left[ \frac{Z_L + jZ_o \tan(Bl)}{Z_o + jZ_L \tan(Bl)} \right] \quad [3.39]$$

Now, for the case of a shorted line (i.e.,  $Z_L = 0$ ), the solution is

$$Z = (Z_o) \left[ \frac{(0) + jZ_o \tan(Bl)}{Z_o + j(0) \tan(Bl)} \right] \quad [3.40]$$

$$Z = (Z_o) \left[ \frac{jZ_o \tan (Bl)}{Z_o} \right] \quad [3.41]$$

$$Z = j Z_o \tan (Bl) \quad [3.42]$$

Recall from Eq. 3.25 that

$$B = WZ_oC \quad [3.43]$$

Substituting Eq. 3.43 into Eq. 3.42 produces

$$Z = j Z_o \tan (\omega Z_o Cl) \quad [3.44]$$

or

$$Z = j Z_o \tan (2\pi FZ_o Cl) \quad [3.45]$$

Because the solutions to Eqs. 3.44 and 3.45 are multiplied by the  $j$  operator, the impedance is actually a reactance ( $Z = 0 + jX$ ). It is possible to achieve almost any possible reactance (within certain practical limitations) by adjusting the length of the transmission line and shorting the “load” end. This fact leads us to a practical method for impedance matching.

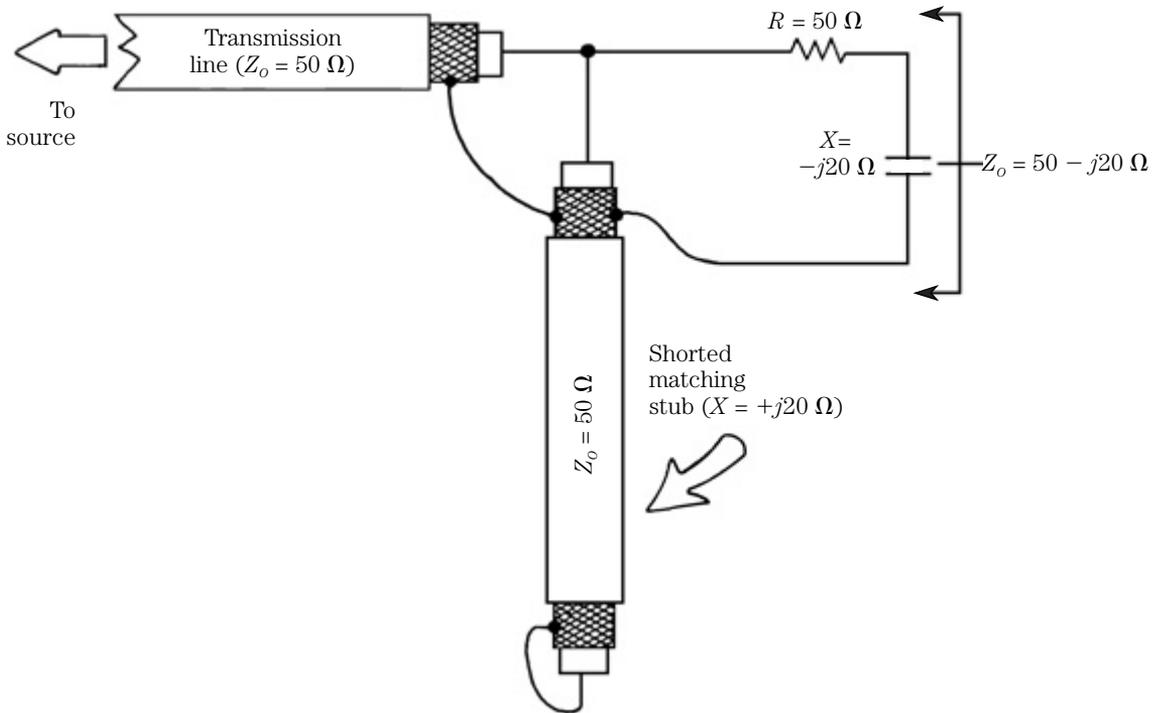
Figure 3-10A shows a circuit in which an unmatched load is connected to a transmission line with characteristic impedance  $Z_o$ . The load impedance  $Z_L$  is of the form  $Z = R \pm jX$ , and, in this case, it is equal to  $50 - j20$ .

A complex impedance load can be matched to its source by interposing the complex conjugate of the impedance. For example, in the case where  $Z = 50 - j20$ , the matching impedance network will require an impedance of  $50 + j20 \Omega$ . The two impedances combine to produce a result of  $50 \Omega$ . The situation of Fig. 3-10A shows a matching stub with a reactance equal in magnitude, but opposite in sign, with respect to the reactive component of the load impedance. In this case, the stub has a reactance of  $+j20 \Omega$  to cancel a reactance of  $-j20 \Omega$  in the load.

A *quarter-wavelength shorted stub* is a special case of the stub concept that finds particular application in microwave circuits. *Waveguides* (Chap. 19) are based on the properties of the quarter-wavelength shorted stub. Figure 3-10B shows a quarter-wave stub and its current distribution. The current is maximum across the short, but wave cancellation forces it to zero at the terminals. Because  $Z = V/I$ , when  $I$  goes to zero, the impedance becomes infinite. Thus, a quarter-wavelength stub has an infinite impedance at its resonant frequency, and redundant acts as an insulator. This concept may be hard to swallow, but the stub is a “metal insulator.”

## Standing wave ratio

The reflection phenomenon was noted earlier during the coverage of the step-function and single-pulse response of a transmission line; the same phenomenon also applies when the transmission line is excited with an ac signal. When a transmission



3-10A Stub matching system.

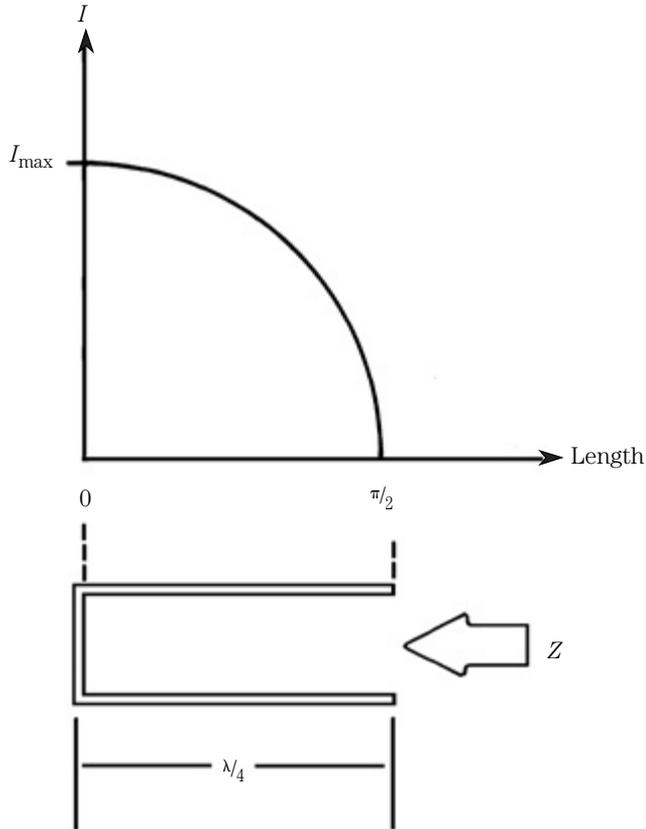
line is not matched to its load, some of the energy is absorbed by the load and some is reflected back down the line toward the source. The interference of incident (or “forward”) and reflected (or “reverse”) waves creates *standing waves* on the transmission line.

If the voltage or current is measured along the line, it will vary, depending on the load, according to Fig. 3-11. Figure 3-11A shows the voltage-versus-length curve for a matched line (i.e., where  $Z_L = Z_o$ ). The line is said to be “flat” because the voltage (and current) is constant all along the line. But now consider Figs. 3-11B and 3-11C.

Figure 3-11B shows the voltage distribution over the length of the line when the load end of the line is *shorted* (i.e.,  $Z_L = 0$ ). Of course, at the load end the voltage is zero, which results from zero impedance. The same impedance and voltage situation is repeated every half-wavelength down the line from the load end toward the generator. Voltage *minima* are called *nodes*, and voltage *maxima* are called *antinodes*.

The pattern in Fig. 3-11C results when the line is unterminated (open) (i.e.,  $Z_L = \infty$ ). Note that the pattern is the same shape as Fig. 3-11B (shorted line), but the phase is shifted  $90^\circ$ . In both cases, the reflection is 100 percent, but the phase of the reflected wave is opposite.

Figure 3-11D shows the situation in which  $Z_L$  is not equal to  $Z_o$ , but is neither zero nor infinite. In this case, the nodes represent some finite voltage,  $V_{\min}$ , rather

**3-10B** Quarter-wavelength stub.

than zero. The *standing wave ratio* (SWR) reveals the relationship between load and line.

If the current along the line is measured, the pattern will resemble the patterns of Fig. 3-11. The SWR is then called ISWR, to indicate the fact that it came from a current measurement. Similarly, if the SWR is derived from voltage measurements it is called VSWR. Perhaps because voltage is easier to measure, VSWR is the term most commonly used in most radio work.

VSWR can be specified in any of several equivalent ways:

1. From incident voltage ( $V_i$ ) and reflected voltage ( $V_r$ ):

$$\text{VSWR} = \frac{V_i + V_r}{V_i - V_r} \quad [3.46]$$

2. From transmission line voltage measurements (Fig. 3-11D):

$$\text{VSWR} = \frac{V_{\text{max}}}{V_{\text{min}}} \quad [3.47]$$

3. From load and line characteristic impedances:

$$(Z_L > Z_o) \text{ VSWR} = Z_L/Z_o \quad \text{[3.48]}$$

$$(Z_L < Z_o) \text{ VSWR} = Z_o/Z_L \quad \text{[3.49]}$$

4. From incident ( $P_i$ ) and reflected ( $P_r$ ) power:

$$\text{VSWR} = \frac{1 + \sqrt{P_r/P_i}}{1 - \sqrt{P_r/P_i}} \quad \text{[3.50]}$$

5. From reflection coefficient ( $\Gamma$ ):

$$\text{VSWR} = \frac{1 + \Gamma}{1 - \Gamma} \quad \text{[3.51]}$$

It is also possible to determine the reflection coefficient  $\Gamma$  from a knowledge of VSWR:

$$\Gamma = \frac{\text{VSWR} - 1}{\text{VSWR} + 1} \quad \text{[3.52]}$$

The relationship between reflection coefficient  $\Gamma$  and VSWR is shown in Fig. 3-11D.

VSWR is usually expressed as a *ratio*. For example, when  $Z_L$  is  $100 \Omega$  and  $Z_o$  is  $50 \Omega$ , the VSWR is  $Z_L/Z_o = 100 \Omega/50 \Omega = 2$ , which is usually expressed as  $\text{VSWR} = 2:1$ . VSWR can also be expressed in decibel form:

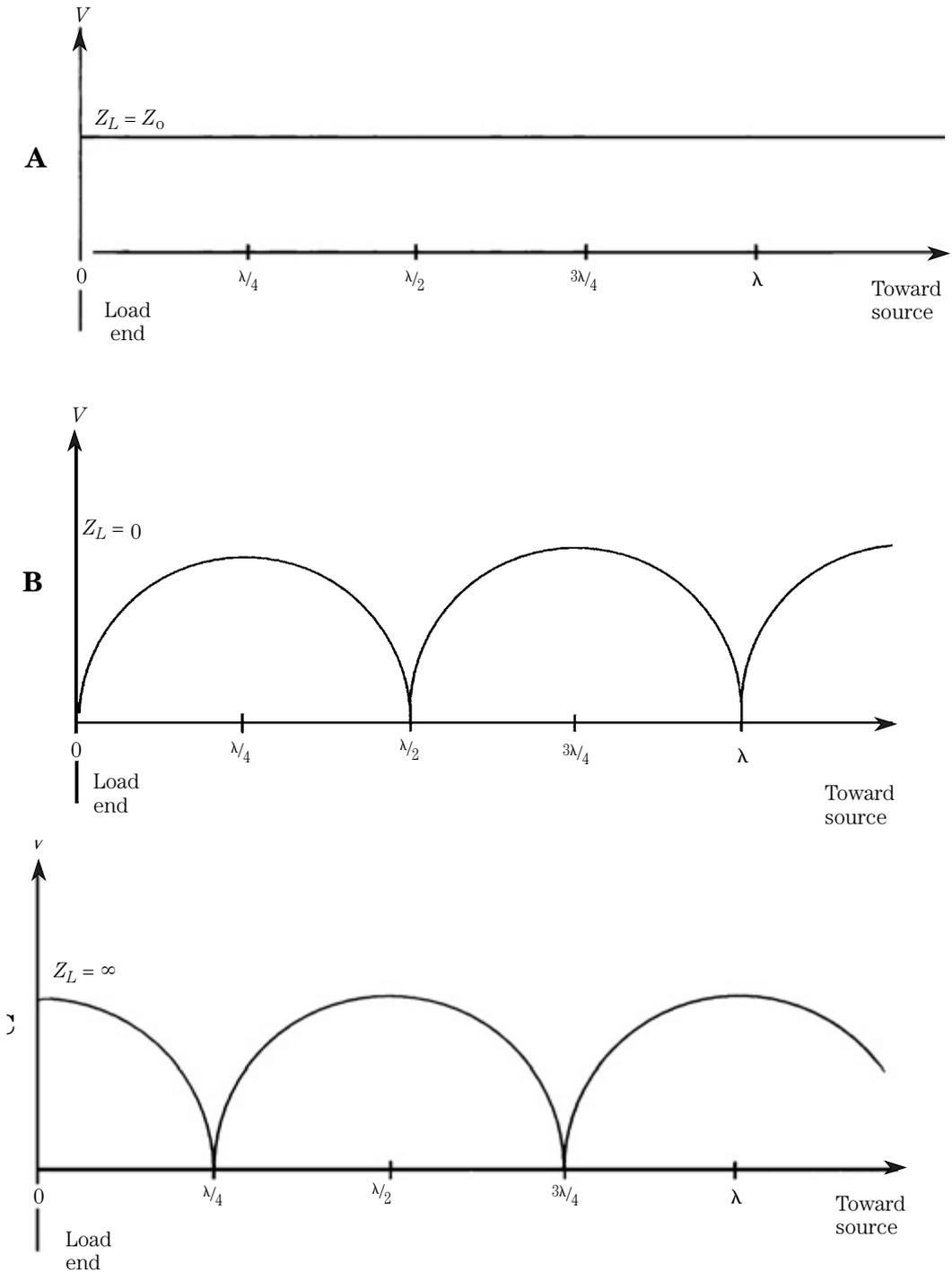
$$\text{VSWR} = 20 \log (\text{VSWR}) \quad \text{[3.53]}$$

**Example 3-7** A transmission line is connected to a mismatched load. Calculate both the VSWR and VSWR decibel equivalent if the reflection coefficient  $\Gamma$  is 0.25.

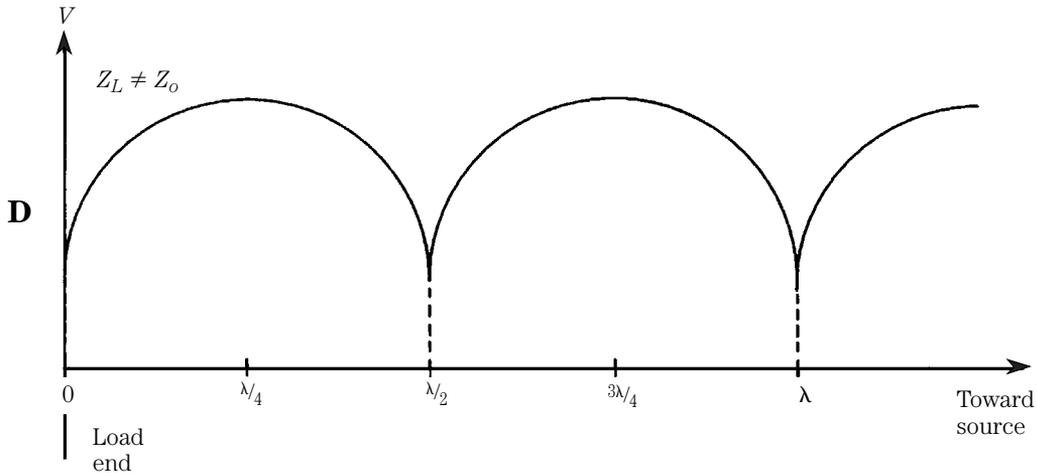
**Solution:**

$$\begin{aligned} (a) \text{ VSWR} &= \frac{1 + \Gamma}{1 - \Gamma} \\ &= \frac{1 + 0.25}{1 - 0.25} \\ &= \frac{1.25}{0.75} = 1.67:1 \end{aligned}$$

$$\begin{aligned} (b) \text{ VSWR}_{\text{dB}} &= 20 \log (\text{VSWR}) \\ &= (20) (\log 1.67) \\ &= (20) (0.22) = 4.3 \text{ dB} \end{aligned}$$



**3-11** Voltage versus electrical length: (A) Matched impedances, (B)  $Z_L = 0$ , (C)  $Z_L = \infty$ .



3-11 Continued: (D)  $Z_L$  not equal to  $Z_o$ .

The SWR is regarded as important in systems for several reasons. The base of these reasons is the fact that the reflected wave represents energy lost to the load. For example, in an antenna system, less power is radiated if some of its input power is reflected back down the transmission line, because the antenna feedpoint impedance does not match the transmission line characteristic impedance. The next section covers the problem of mismatch losses.

### Mismatch (VSWR) losses

The power reflected from a mismatched load represents a loss, and will have implications that range from negligible to profound, depending on the situation. For example, one result might be a slight loss of signal strength at a distant point from an antenna. A more serious problem can result in the destruction of the output device in a transmitter. The latter problem so plagued early solid-state transmitters that designers opted to include shutdown circuitry to sense high VSWR, and turn down output power proportionally.

In microwave measurements, VSWR on the transmission lines (that interconnect devices under test, instruments, and signal sources) can cause erroneous readings—and invalid measurements.

Determination of VSWR losses must take into account *two* VSWR situations. Figure 3-9 shows a transmission line of impedance  $Z_o$  interconnecting a load impedance  $Z_L$ , and a source with an output impedance  $Z_s$ . There is a potential for impedance mismatch at *both ends* of the line.

In the case where one end of the line is matched (either  $Z_s$  or  $Z_L$ ), the *mismatch loss* caused by SWR at the mismatched end is

$$ML = -10 \log \left[ 1 - \left( \frac{SWR - 1}{SWR + 1} \right)^2 \right] \quad [3.54]$$

which from Eq. 3.52 is

$$ML = -10 \log (1 - \Gamma^2) \quad \mathbf{[3.55]}$$

**Example 3-8** A coaxial transmission line with a characteristic impedance of 50  $\Omega$  is connected to the 50- $\Omega$  output ( $Z_o$ ) of a signal generator, and also to a 20-W load impedance  $Z_L$ . Calculate the mismatch loss.

**Solution:**

(a) First find the VSWR:

$$\begin{aligned} \text{VSWR} &= Z_o/Z_L \\ &= (50 \Omega)/(20 \Omega) = 2.5:1 \end{aligned}$$

(b) Mismatch loss:

$$\begin{aligned} ML &= -10 \log \left[ 1 - \left( \frac{\text{SWR} - 1}{\text{SWR} + 1} \right)^2 \right] \\ &= -10 \log \left[ 1 - \left( \frac{2.5 - 1}{2.5 + 1} \right)^2 \right] \\ &= -10 \log \left[ 1 - \left( \frac{1.5}{3.5} \right)^2 \right] \\ &= -10 \log [1 - (0.43)^2] \\ &= -10 \log [1 - 0.185] \\ &= -10 \log [0.815] \\ &= (-10) (-0.089) = 0.89 \end{aligned}$$

When both ends of the line are mismatched, a different equation is required:

$$ML = 20 \log [1 \pm (\Gamma_1 \times \Gamma_2)] \quad \mathbf{[3.56]}$$

where

$\Gamma_1$  is the reflection coefficient at the source end of the line,  
 $(\text{VSWR}_1 - 1)/(\text{VSWR}_1 + 1)$

$\Gamma_2$  is the reflection coefficient at the load end of the line,  
 $(\text{VSWR}_2 - 1)/(\text{VSWR}_2 + 1)$

Note that the solution to Eq. 3.56 has two values:  $[1 + (\Gamma_1 \Gamma_2)]$  and  $[1 - (\Gamma_1 \Gamma_2)]$ .

The equations reflect the mismatch loss solution for low-loss or “lossless” transmission lines. This is a close approximation in some situations; however, it is insufficient when the line is lossy. Although not very important at low frequencies, loss becomes higher at microwave frequencies. Interference between incident and re-

flected waves produces increased current at certain antinodes—which increases ohmic losses—and increased voltage at certain antinodes—which increases dielectric losses. It is the latter that increases with frequency. Equation 3.57 relates reflection coefficient and line losses to determine total loss on a given line.

$$\text{Loss} = 10 \log \left( \frac{n^2 - \Gamma^2}{n - n\Gamma^2} \right) \quad [3.57]$$

where

Loss is the total line loss in decibels

$\Gamma$  is the reflection coefficient

$n$  is the quantity  $10^{(A/10)}$

$A$  is the total attenuation presented by the line, in decibels, when the line is properly matched ( $Z_L = Z_o$ )

**Example 3-9** A 50- $\Omega$  transmission line is terminated in a 30- $\Omega$  resistive impedance. The line is rated at a loss of 3 dB/100 ft at 1 GHz. Calculate (a) loss in 5 ft of line, (b) reflection coefficient, and (c) total loss in a 5-ft line mismatched per above.

**Solution:**

$$(a) A = \frac{3 \text{ dB}}{100 \text{ ft}} \times 5 \text{ ft} = 0.15 \text{ dB}$$

$$(b) \Gamma = \frac{Z_L - Z_o}{Z_L + Z_o}$$

$$= \frac{50 - 30}{50 + 30}$$

$$= 20/80 = 0.25$$

$$(c) n = 10^{(A/10)}$$

$$= 10^{(0.15/10)}$$

$$= 10^{(0.015)} = 1.04$$

$$\begin{aligned} \text{Loss} &= 10 \log \left( \frac{n^2 - \Gamma^2}{n - n\Gamma^2} \right) \\ &= 10 \log \left[ \frac{(1.04)^2 - (0.25)^2}{1.04 - (1.04)(0.25)^2} \right] \\ &= 10 \log \left[ \frac{1.082 - 0.063}{1.04 - (1.04)(0.063)} \right] \\ &= 10 \log \left( \frac{1.019}{1.04 - 0.066} \right) \end{aligned}$$

$$\begin{aligned} &= 10 \log \left( \frac{1.019}{0.974} \right) \\ &= 10 \log (1.046) \\ &= (10) (0.02) = 0.2 \text{ dB} \end{aligned}$$

Compare the matched line loss ( $A = 0.15$  dB) with the total loss (Loss = 0.2 dB), which includes mismatch loss and line loss. The difference (i.e., Loss -  $A$ ) is only 0.05 dB. If the VSWR was considerably larger, however, the loss would rise.

## 4 CHAPTER

# The Smith chart

THE MATHEMATICS OF TRANSMISSION LINES, AND CERTAIN OTHER DEVICES, BECOMES cumbersome at times, especially in dealing with complex impedances and “nonstandard” situations. In 1939, Phillip H. Smith published a graphical device for solving these problems, and an improved version of the chart followed in 1945. That graphic aid, somewhat modified over time, is still in constant use in microwave electronics, and other fields where complex impedances and transmission line problems are found. The *Smith chart* is indeed a powerful tool for the RF designer.

## Smith chart components

The modern Smith chart is shown in Fig. 4-1. It consists of a series of overlapping orthogonal circles (i.e., circles that intersect each other at right angles). This chapter will dissect the Smith chart, so that the origin and use of these circles is apparent. The set of orthogonal circles makes up the basic structure of the Smith chart.

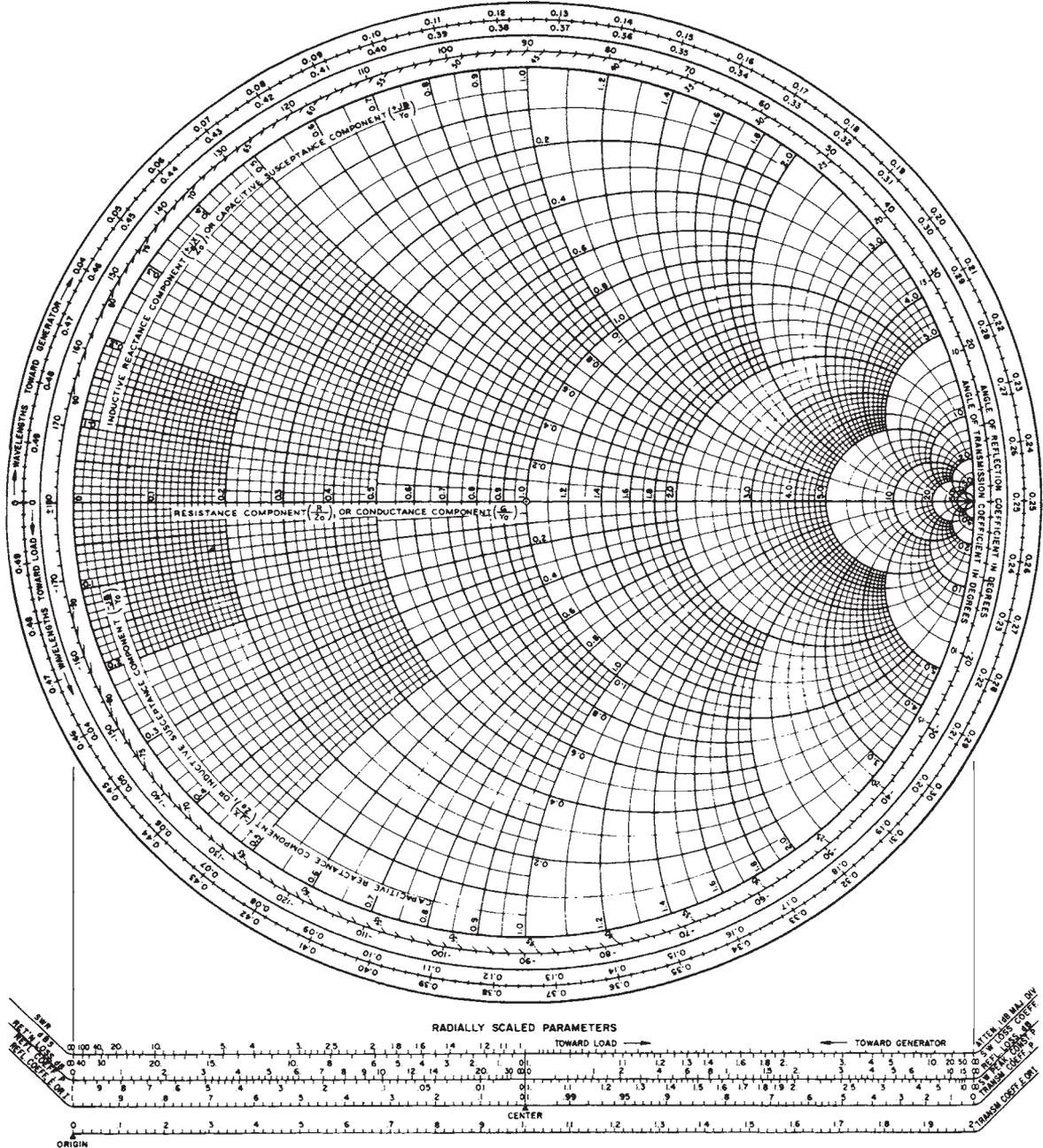
### The normalized impedance line

A baseline is highlighted in Fig. 4-2, and bisects the Smith chart outer circle. This line is called the *pure resistance line*, and forms the reference for measurements made on the chart. Recall that a complex impedance contains both resistance and reactance, and is expressed in the mathematical form:

$$Z = R \pm jX \quad [4.1]$$

NAME	TITLE	DWG. NO.
SMITH CHART FORM 82-BSPR(9-66)	KAY ELECTRIC COMPANY, PINE BROOK, N. J. © 1966. PRINTED IN U.S.A.	DATE

IMPEDANCE OR ADMITTANCE COORDINATES



4-1 The Smith chart. (Courtesy of Kay Elementrics)

where

$Z$  is the complex impedance

$R$  is the resistive component of the impedance

$X$  is the reactive component of the impedance\*

The pure resistance line represents the situation where  $X = 0$ , and the impedance is therefore equal to the resistive component only. In order to make the Smith chart universal, the impedances along the pure resistance line are *normalized* with reference to system impedance (e.g.,  $Z_o$  in transmission lines); for most microwave RF systems the system impedance is standardized at  $50\ \Omega$ . In order to normalize the actual impedance, divide it by the system impedance. For example, if the load impedance of a transmission line is  $Z_L$ , and the characteristic impedance of the line is  $Z_o$ , then  $Z = Z_L/Z_o$ . In other words,

$$Z = \frac{R \pm jX}{Z_o} \quad [4.2]$$

The pure resistance line is structured such that the system standard impedance is in the center of the chart, and has a normalized value of 1.0 (see point *A* in Fig. 4-2). This value derives from the fact that  $Z_o/Z_o = 1.0$ .

To the left of the 1.0 point are decimal fraction values used to denote impedances less than the system impedance. For example, in a  $50\text{-}\Omega$  transmission line system with a  $25\text{-}\Omega$  load impedance, the normalized value of impedance is  $25\ \Omega/50\ \Omega$  or 0.50 (“B” in Fig. 4-2). Similarly, points to the right of 1.0 are greater than 1 and denote impedances that are higher than the system impedance. For example, in a  $50\text{-}\Omega$  system connected to a  $100\text{-}\Omega$  resistive load, the normalized impedance is  $100\ \Omega/50\ \Omega$ , or 2.0; this value is shown as point *C* in Fig. 4-2. By using normalized impedances, you can use the Smith chart for almost any practical combination of system, and load and/or source, impedances, whether resistive, reactive, or complex.

Reconversion of the normalized impedance to actual impedance values is done by multiplying the normalized impedance by the system impedance. For example, if the resistive component of a normalized impedance is 0.45, then the actual impedance is

$$Z = (Z_{\text{normal}}) (Z_o) \quad [4.3]$$

$$= (0.45) (50\ \Omega) \quad [4.4]$$

$$= 22.5\ \Omega \quad [4.5]$$

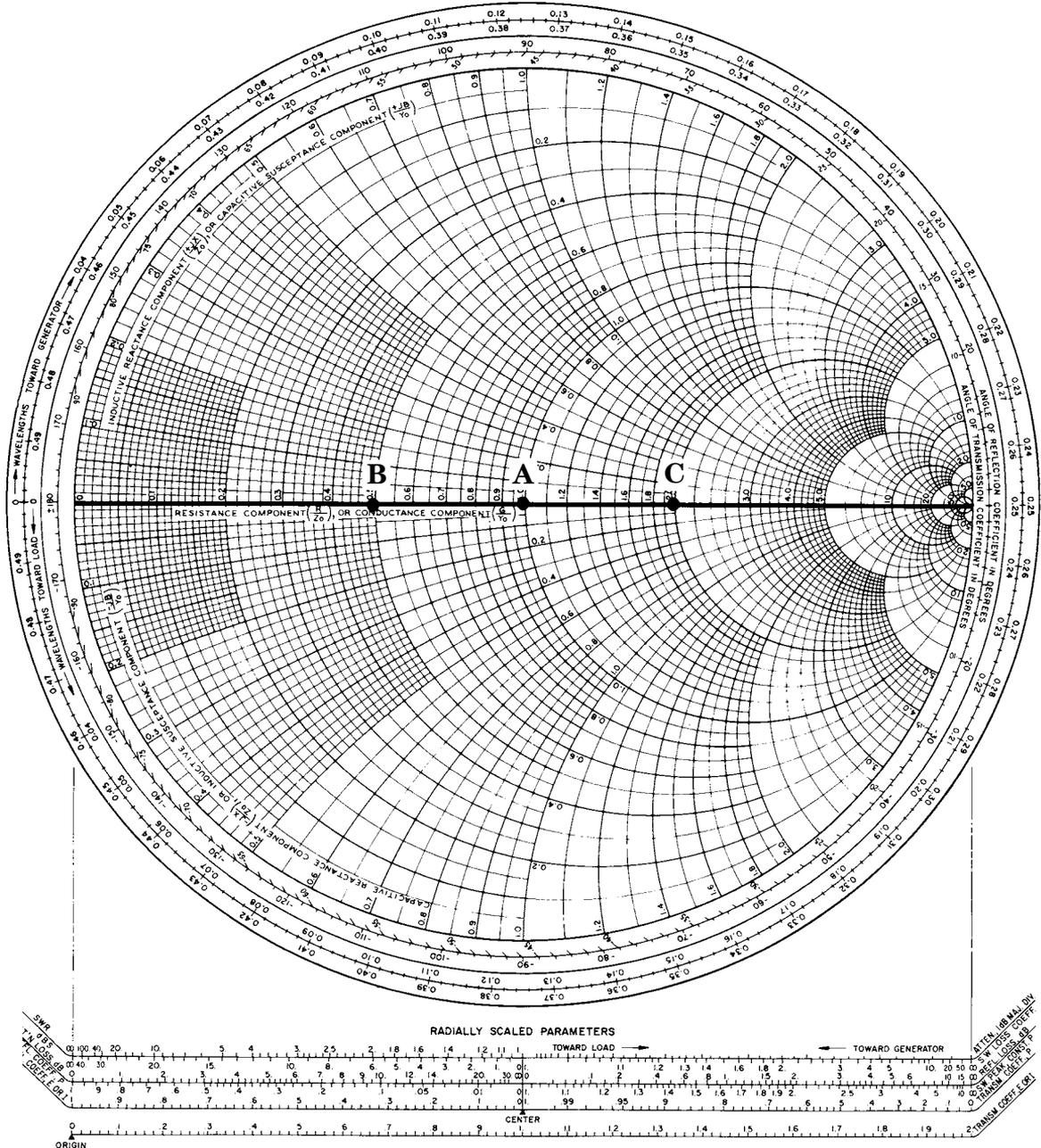
### The constant resistance circles

The *isoresistance circles*, also called the *constant resistance circles*, represent points of equal resistance. Several of these circles are shown highlighted in Fig. 4-3. These circles are all tangent to the point at the righthand extreme of the pure resis-

\*According to the standard sign convention the inductive reactance ( $X_L$ ) is positive (+) and the capacitive reactance ( $X_C$ ) is negative (-). The term  $X$  in Eq. 4.1 above is the difference between the two reactances ( $X = X_L - X_C$ ).

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SMITH CHART FORM 82-BSPR (9-66)	KAY ELECTRIC COMPANY, PINE BROOK, N. J. © 1966 PRINTED IN U.S.A.	DATE

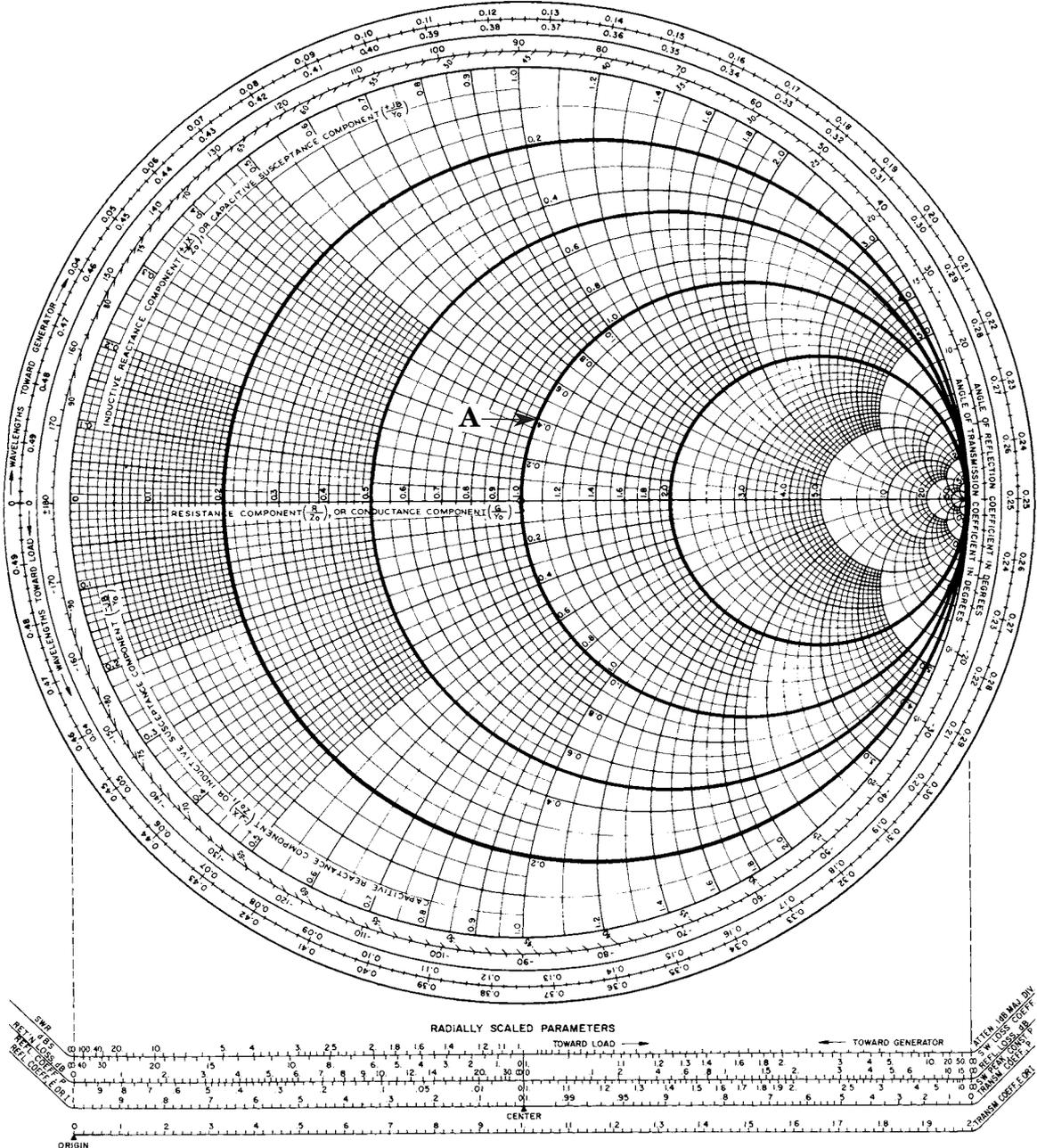
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4-2 Normalized impedance line. (Courtesy of Kay Elemtetrics)

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4-3 Constant resistance circles. (Courtesy of Kay Elementrics)

tance line, and are bisected by that line. When you construct complex impedances (for which  $X \neq 0$ ) on the Smith chart, the points on these circles will all have the same resistive component. Circle  $A$ , for example, passes through the center of the chart, so it has a normalized constant resistance of 1.0. Note that impedances that are pure resistances (i.e.,  $Z = R + j0$ ) will fall at the intersection of a constant resistance circle and the pure resistance line, and complex impedances (i.e.,  $X \neq 0$ ) will appear at any other points on the circle. In Fig. 4-3, circle  $A$  passes through the center of the chart, so it represents all points on the chart with a normalized resistance of 1.0. This particular circle is sometimes called the *unity resistance circle*.

### The constant reactance circles

Constant reactance circles are highlighted in Fig. 4-4. The circles (or circle segments) *above* the pure resistance line (Fig. 4-4A) represent the *inductive reactance* ( $+X$ ), and those circles (or segments) *below* the pure resistance line (Fig. 4-4B) represent *capacitive reactance* ( $-X$ ). In both cases, circle  $A$  represents a normalized reactance of 0.80.

One of the outer circles (i.e., circle  $A$  in Fig. 4-4C) is called the *pure reactance circle*. Points along circle  $A$  represent reactance only; in other words, an impedance of  $Z = 0 \pm jX$  ( $R = 0$ ).

Figure 4-4D shows how to plot impedance and admittance on the Smith chart. Consider an example in which system impedance  $Z_o$  is  $50 \Omega$ , and the load impedance is  $Z_L = 95 + j55 \Omega$ . This load impedance is normalized to

$$Z = \frac{Z_L}{Z_o} \quad [4.6]$$

$$= \frac{95 + j55 \Omega}{50 \Omega} \quad [4.7]$$

$$= 1.9 + j1.1 \quad [4.8]$$

An *impedance radius* is constructed by drawing a line from the point represented by the normalized load impedance,  $1.9 + j1.1$ , to the point represented by the normalized system impedance (1.0) in the center of the chart. A circle is constructed from this radius, and is called the *VSWR circle*.

*Admittance* is the reciprocal of impedance, so it is found from

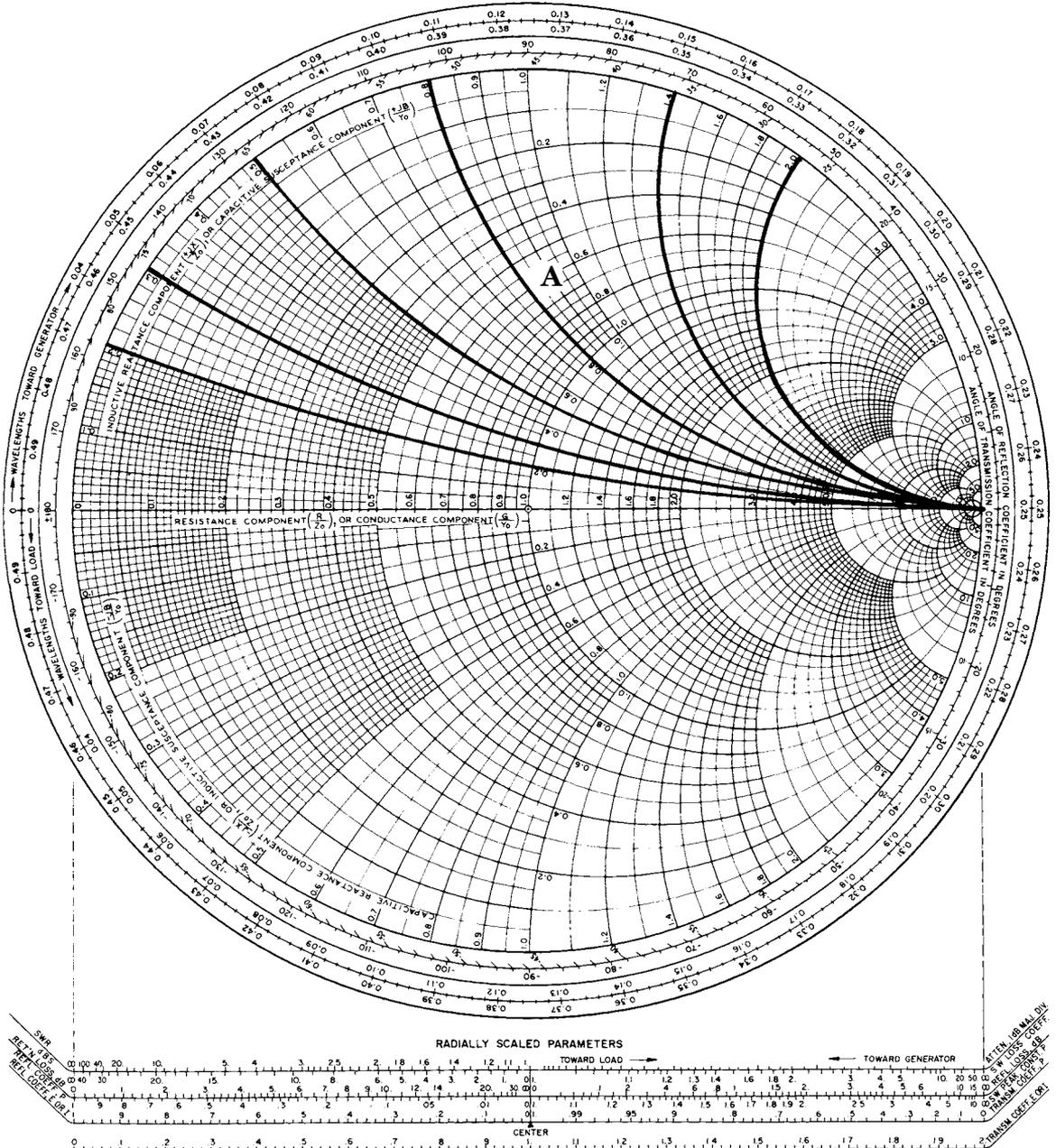
$$Y = \frac{1}{Z} \quad [4.9]$$

Because impedances in transmission lines are rarely pure resistive, but rather contain a reactive component also, impedances are expressed by complex notation:

$$Z = R \pm jX \quad [4.10]$$

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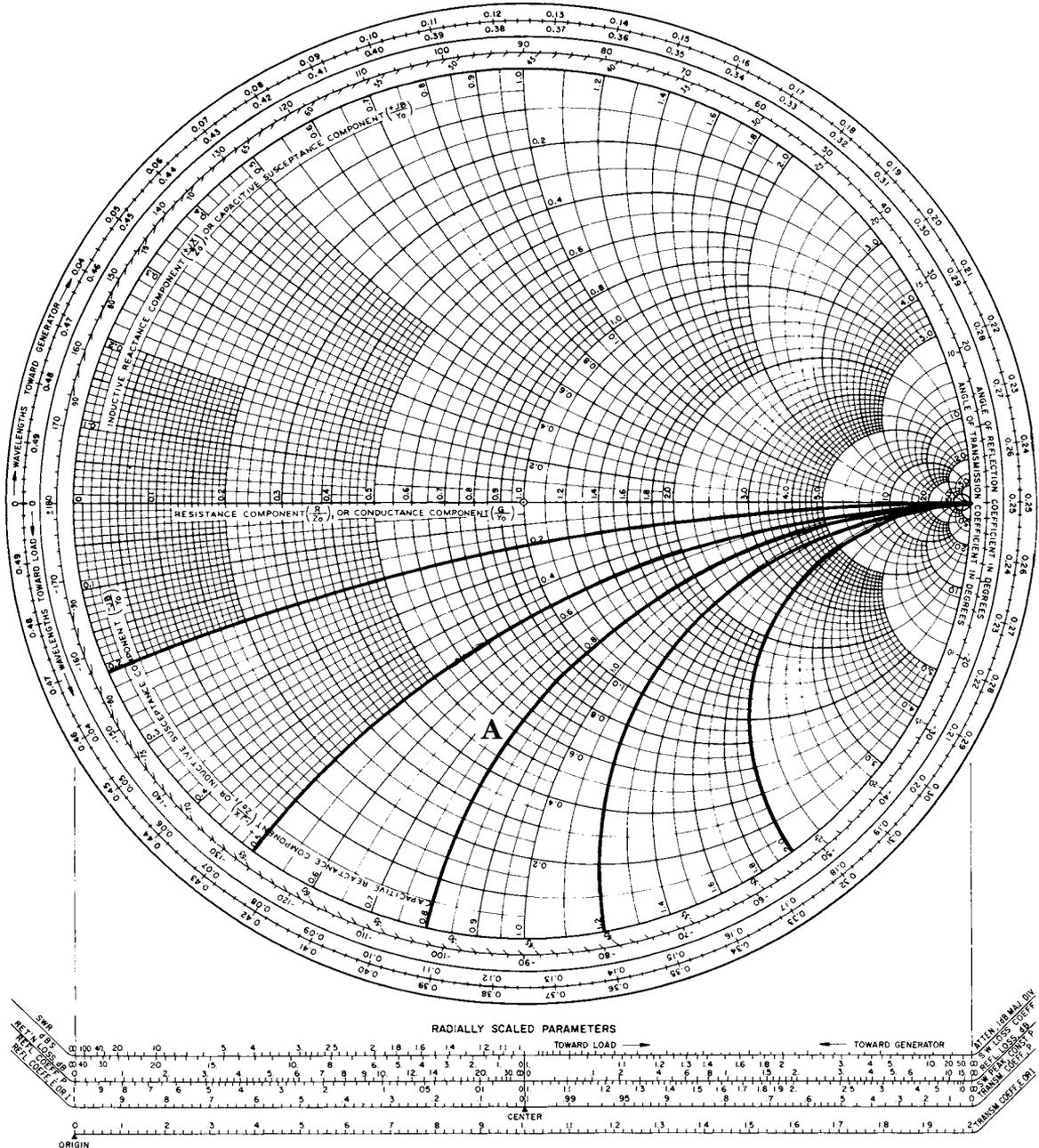
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4-4A Constant inductive reactance lines. (Courtesy of Kay Elementrics)

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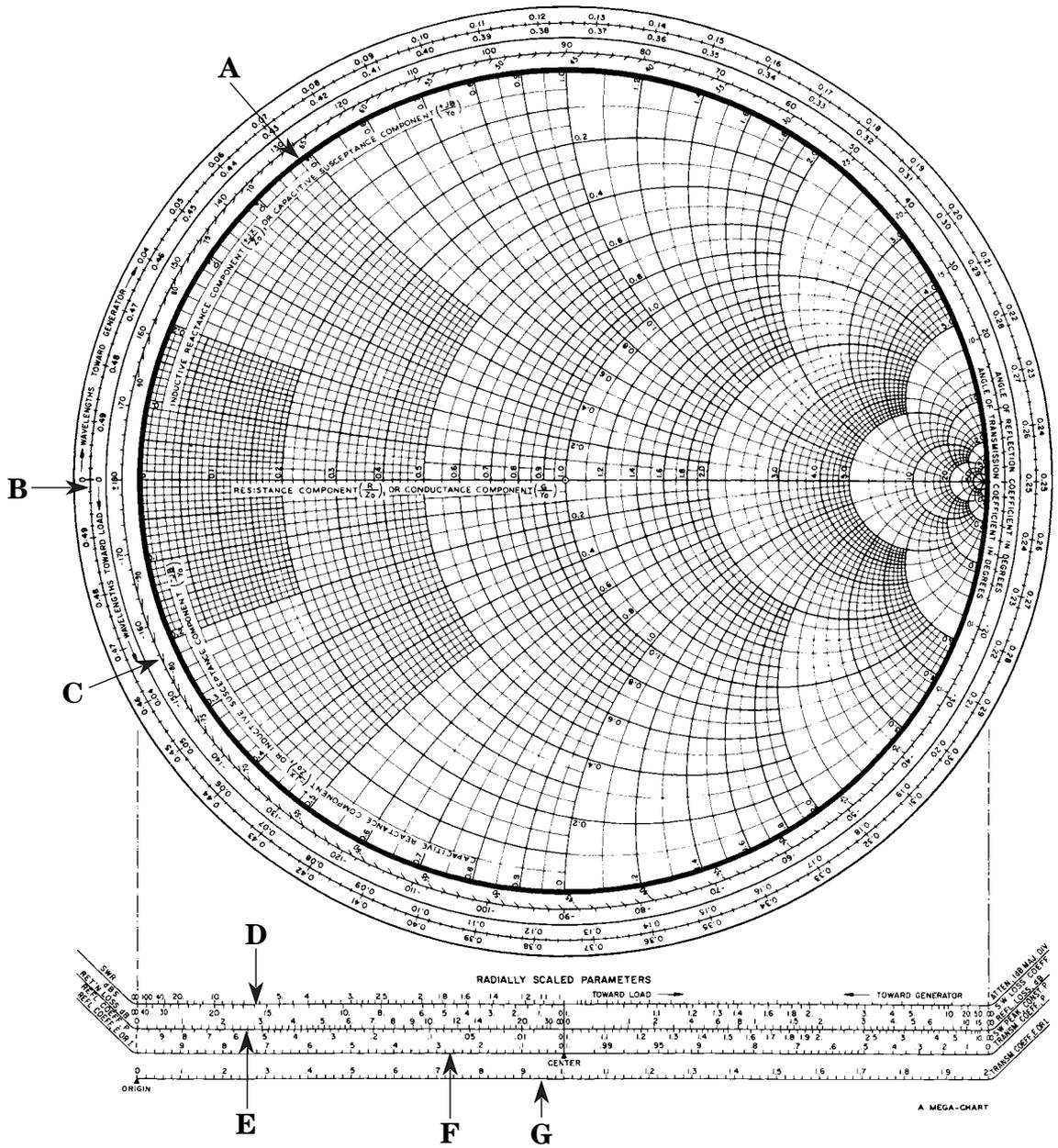
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4-4B Constant capacitive reactance lines. (Courtesy of Kay Elemtetrics)

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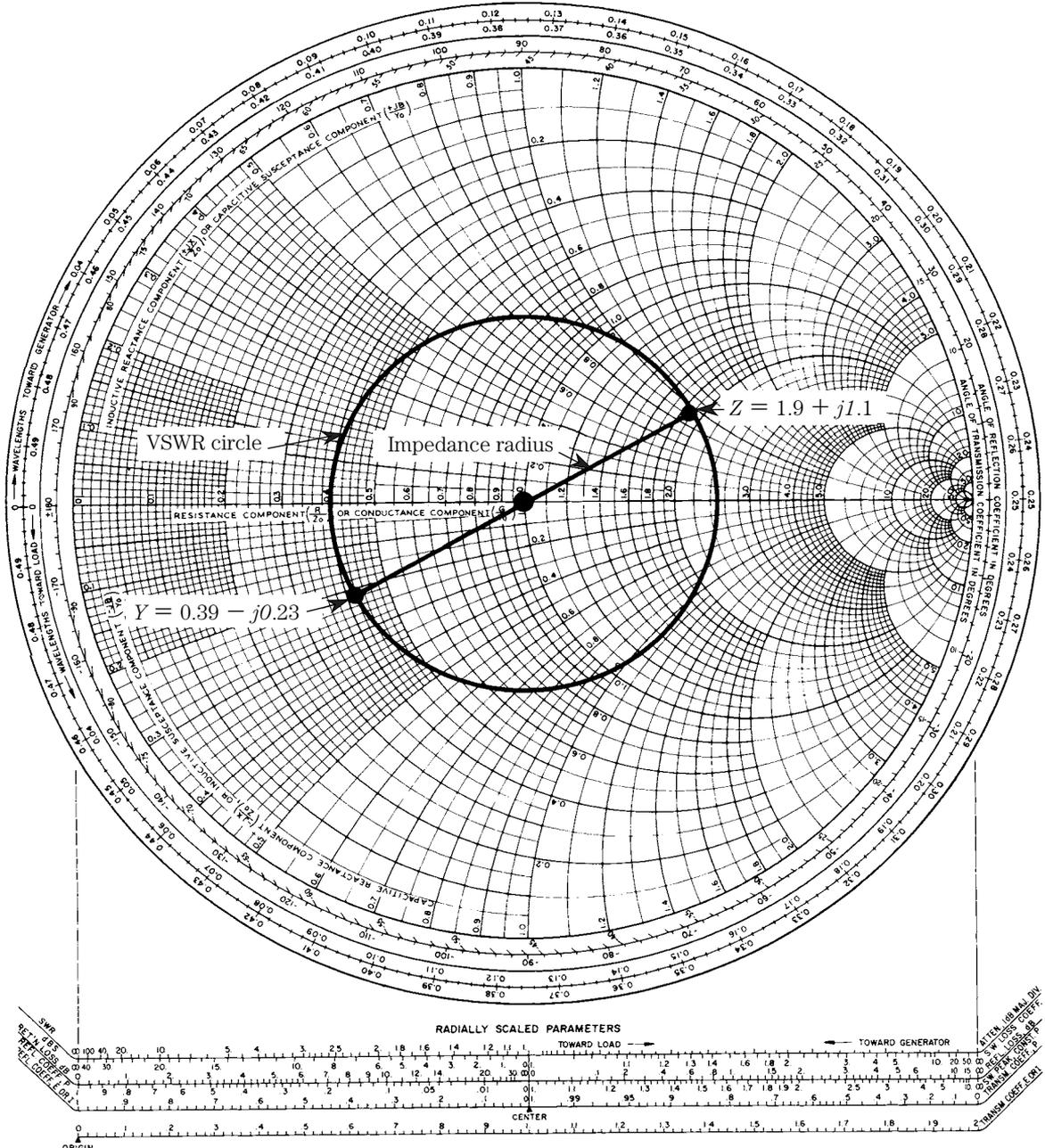
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4-4C Angle of transmission coefficient circle. (Courtesy of Kay Elemtetrics)

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4-4D VSWR circles. (Courtesy of Kay Eletronics)

where

$Z$  is the complex impedance

$R$  is the resistive component

$X$  is the reactive component

In order to find the *complex admittance*, take the reciprocal of the complex impedance by multiplying the simple reciprocal by the complex conjugate of the impedance. For example, when the normalized impedance is  $1.9 + j1.1$ , the normalized admittance will be:

$$Y = \frac{1}{Z} \quad [4.11]$$

$$= \frac{1}{1.9 + j1.1} \times \frac{1.9 - j1.1}{1.9 - j1.1} \quad [4.12]$$

$$= \frac{1.9 - j1.1}{3.6 + 1.2} \quad [4.13]$$

$$= \frac{1.9 - j1.1}{4.8} = 0.39 - j0.23 \quad [4.14]$$

One of the delights of the Smith chart is that this calculation is reduced to a quick graphical interpretation! Simply extend the impedance radius through the 1.0 center point until it intersects the VSWR circle again. This point of intersection represents the normalized admittance of the load.

### Outer circle parameters

The standard Smith chart shown in Fig. 4-4C contains three concentric calibrated circles on the outer perimeter of the chart. Circle  $A$  has already been covered and it is the pure reactance circle. The other two circles define the wavelength distance ( $B$ ) relative to either the load or generator end of the transmission line, and either the transmission or reflection coefficient angle in degrees ( $C$ ).

There are two scales on the *wavelength* circle ( $B$  in Fig 4-4C), and both have their zero origin on the left-hand extreme of the pure resistance line. Both scales represent *one-half wavelength for one entire revolution*, and are calibrated from 0 through 0.50 such that these two points are identical with each other on the circle. In other words, starting at the zero point and travelling  $360^\circ$  around the circle brings one back to zero, which represents one-half wavelength, or  $0.5 \lambda$ .

Although both wavelength scales are of the same magnitude ( $0 - 0.50$ ), they are opposite in direction. The outer scale is calibrated clockwise and it represents *wavelengths toward the generator*; the inner scale is calibrated counterclockwise and represents *wavelengths toward the load*. These two scales are complementary at all points. Thus, 0.12 on the outer scale corresponds to  $(0.50 - 0.12)$  or 0.38 on the inner scale.

The *angle of transmission coefficient* and *angle of reflection coefficient* scales are shown in circle *C* in Fig. 4-4C. These scales are the relative phase angle between reflected and incident waves. Recall from transmission line theory (see Chap. 3), that a short circuit at the load end of the line reflects the signal back toward the generator  $180^\circ$  out of phase with the incident signal; an open line (i.e., infinite impedance) reflects the signal back to the generator in phase (i.e.,  $0^\circ$ ) with the incident signal. These facts are shown on the Smith chart by the fact that both scales start at  $0^\circ$  on the right-hand end of the pure resistance line, which corresponds to an infinite resistance, and it goes halfway around the circle to  $180^\circ$  at the 0 end of the pure resistance line. Note that the upper half-circle is calibrated 0 to  $+180^\circ$ , and the bottom half-circle is calibrated 0 to  $-180^\circ$ , reflecting inductive or capacitive reactance situations, respectively.

### Radially scaled parameters

There are six scales laid out on five lines (*D* through *G* in Fig. 4-4C and in expanded form in Fig. 4-5) at the bottom of the Smith chart. These scales are called the *radially scaled parameters*—and they are both very important, and often overlooked. With these scales, we can determine such factors as VSWR (both as a ratio and in decibels), return loss in decibels, voltage or current reflection coefficient, and the power reflection coefficient.

The *reflection coefficient*  $\Gamma$  is defined as the *ratio of the reflected signal to the incident signal*. For voltage or current:

$$\Gamma = \frac{E_{\text{ref}}}{E_{\text{inc}}} \quad [4.15]$$

and

$$\Gamma = \frac{I_{\text{ref}}}{I_{\text{inc}}} \quad [4.16]$$

Power is proportional to the square of voltage or current, so:

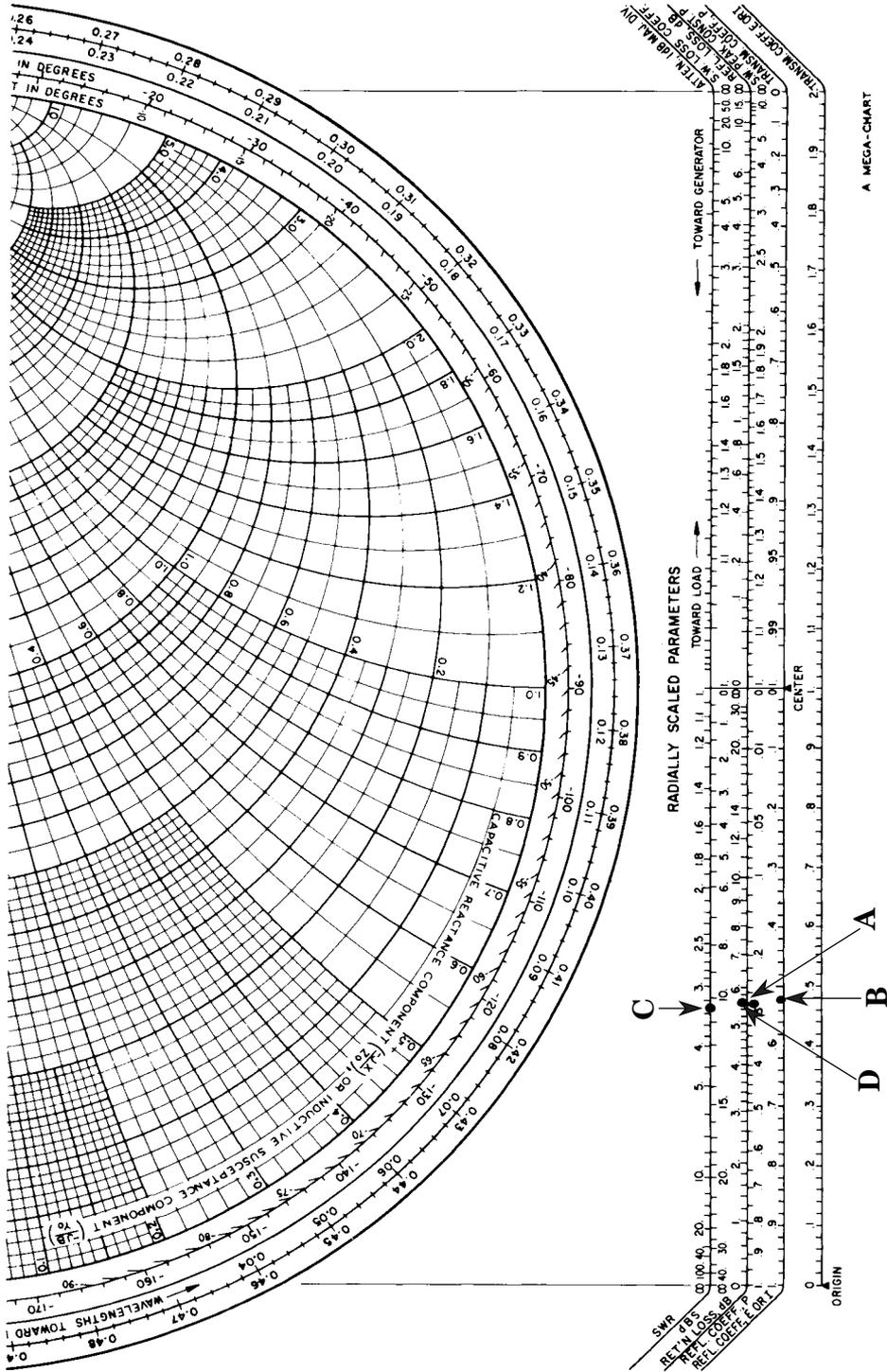
$$P_{\text{pwr}} = \Gamma^2 \quad [4.17]$$

or

$$\Gamma_{\text{pwr}} = \frac{P_{\text{ref}}}{P_{\text{inc}}} \quad [4.18]$$

**Example** 10 W of microwave RF power is applied to a lossless transmission line, of which 2.8 W is reflected from the mismatched load. Calculate the reflection coefficient.

$$\Gamma_{\text{pwr}} = \frac{P_{\text{ref}}}{P_{\text{inc}}} \quad [4.19]$$



4-5 Radially scaled parameters.

$$\Gamma_{\text{pwr}} = \frac{2.8 \text{ W}}{10 \text{ W}} \quad [4.20]$$

$$= 0.28 \quad [4.21]$$

The voltage reflection coefficient  $\Gamma$  is found by taking the square root of the power reflection coefficient, so in this example it is equal to 0.529. These points are plotted at *A* and *B* in Fig. 4-5.

*Standing wave ratio* (SWR) can be defined in terms of reflection coefficient:

$$\text{VSWR} = \frac{1 + \Gamma}{1 - \Gamma} \quad [4.22]$$

or

$$\text{VSWR} = \frac{1 + \sqrt{\Gamma_{\text{pwr}}}}{1 - \sqrt{\Gamma_{\text{pwr}}}} \quad [4.23]$$

or, in our example,

$$\text{VSWR} = \frac{1 + \sqrt{0.28}}{1 - \sqrt{0.28}} \quad [4.24]$$

$$= \frac{1 + 0.529}{1 - 0.529} \quad [4.25]$$

$$= \frac{1.529}{0.471} = 3.25:1 \quad [4.26]$$

or, in decibel form,

$$\text{VSWR}_{\text{dB}} = 20 \log (\text{VSWR}) \quad [4.27]$$

$$= 20 \log (20) \quad [4.28]$$

$$= (20) (0.510) = 10.2 \text{ dB} \quad [4.29]$$

These points are plotted at *C* in Fig. 4-5. Shortly, you will work an example to show how these factors are calculated in a transmission line problem from a known complex load impedance.

*Transmission loss* is a measure of the one-way loss of power in a transmission line because of reflection from the load. *Return loss* represents the two-way loss, so it is exactly twice the transmission loss. Return loss is found from

$$\text{Loss}_{\text{ret}} = 10 \log (\Gamma_{\text{pwr}}) \quad [4.30]$$

and, for our example in which  $\Gamma_{\text{pwr}} = 0.28$ ,

$$\text{Loss}_{\text{ret}} = 10 \log (0.28) \quad [4.31]$$

$$= (10) (-0.553) = -5.53 \text{ dB} \quad [4.32]$$

This point is shown as  $D$  in Fig. 4-5.

The *transmission loss coefficient* can be calculated from

$$\text{TLC} = \frac{1 + \Gamma_{\text{pwr}}}{1 - \Gamma_{\text{pwr}}} \quad [4.33]$$

or, for our example,

$$\text{TLC} = \frac{1 + (0.28)}{1 - (0.28)} \quad [4.34]$$

$$= \frac{1.28}{0.72} = 1.78 \quad [4.35]$$

The TLC is a correction factor that is used to calculate the attenuation caused by mismatched impedance in a lossy, as opposed to the ideal “lossless,” line. The TLC is found from laying out the impedance radius on the *loss coefficient* scale on the radially scaled parameters at the bottom of the chart.

## Smith chart applications

One of the best ways to demonstrate the usefulness of the Smith chart is by practical example. The following sections look at two general cases: *transmission line problems* and *stub matching systems*.

### Transmission line problems

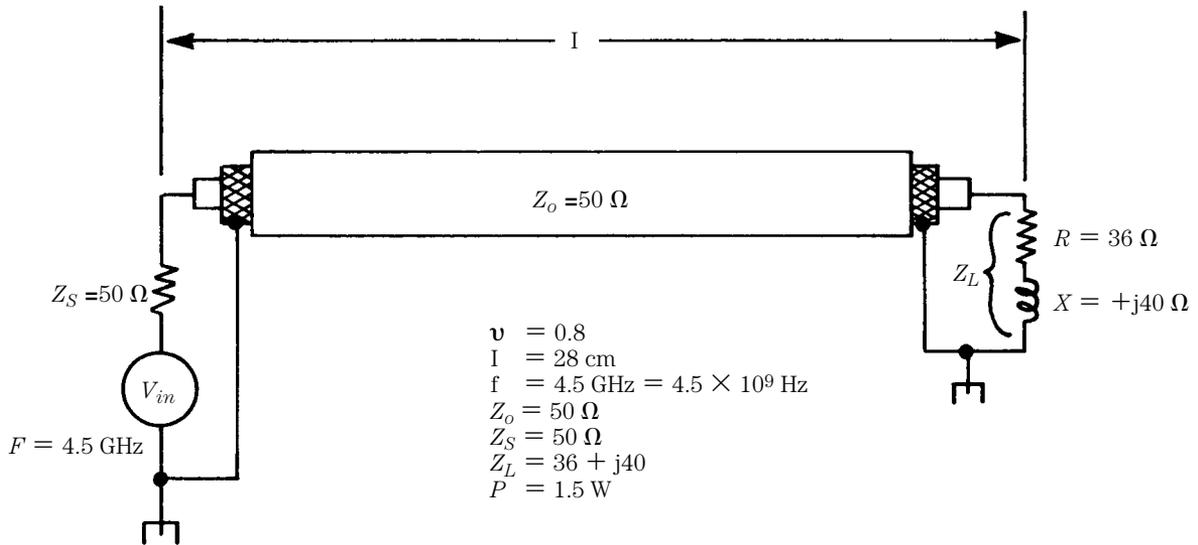
Figure 4-6 shows a 50- $\Omega$  transmission line connected to a complex load impedance  $Z_L$  of  $36 + j40 \Omega$ . The transmission line has a velocity factor  $v$  of 0.80, which means that the wave propagates along the line at 80% the speed of light ( $c = 300,000,000$  m/s). The length of the transmission line is 28 cm. The generator  $V_{\text{in}}$  is operated at a frequency of 4.5 GHz and produces a power output of 1.5 W. See what you can glean from the Smith chart (Fig. 4-7).

First, normalize the load impedance. This is done by dividing the load impedance by the systems impedance (in this case,  $Z_o = 50 \Omega$ ):

$$Z = \frac{36 + j40 \Omega}{50 \Omega} \quad [4.36]$$

$$= 0.72 + j0.8 \quad [4.37]$$

The resistive component of impedance  $Z$  is located along the 0.72 pure resistance circle (see Fig. 4-7). Similarly, the reactive component of impedance  $Z$  is located by traversing the 0.72 constant resistance circle until the  $+j0.8$  constant reactance circle is intersected. This point graphically represents the normalized load impedance  $Z = 0.72 + j0.80$ . A VSWR circle is constructed with an impedance radius equal to the line between 1.0 (in the center of the chart) and the  $0.72 + j0.8$  point.



4-6 Transmission line and load circuit.

At a frequency of 4.5 GHz, the length of a wave propagating in the transmission line, assuming a velocity factor of 0.80, is

$$\lambda_{\text{line}} = \frac{c v}{F_{\text{Hz}}} \quad [4.38]$$

$$= \frac{(3 \times 10^8 \text{ m/s}) (0.80)}{4.5 \times 10^9 \text{ Hz}} \quad [4.39]$$

$$= \frac{2.4 \times 10^8 \text{ m/s}}{4.5 \times 10^9 \text{ Hz}} \quad [4.40]$$

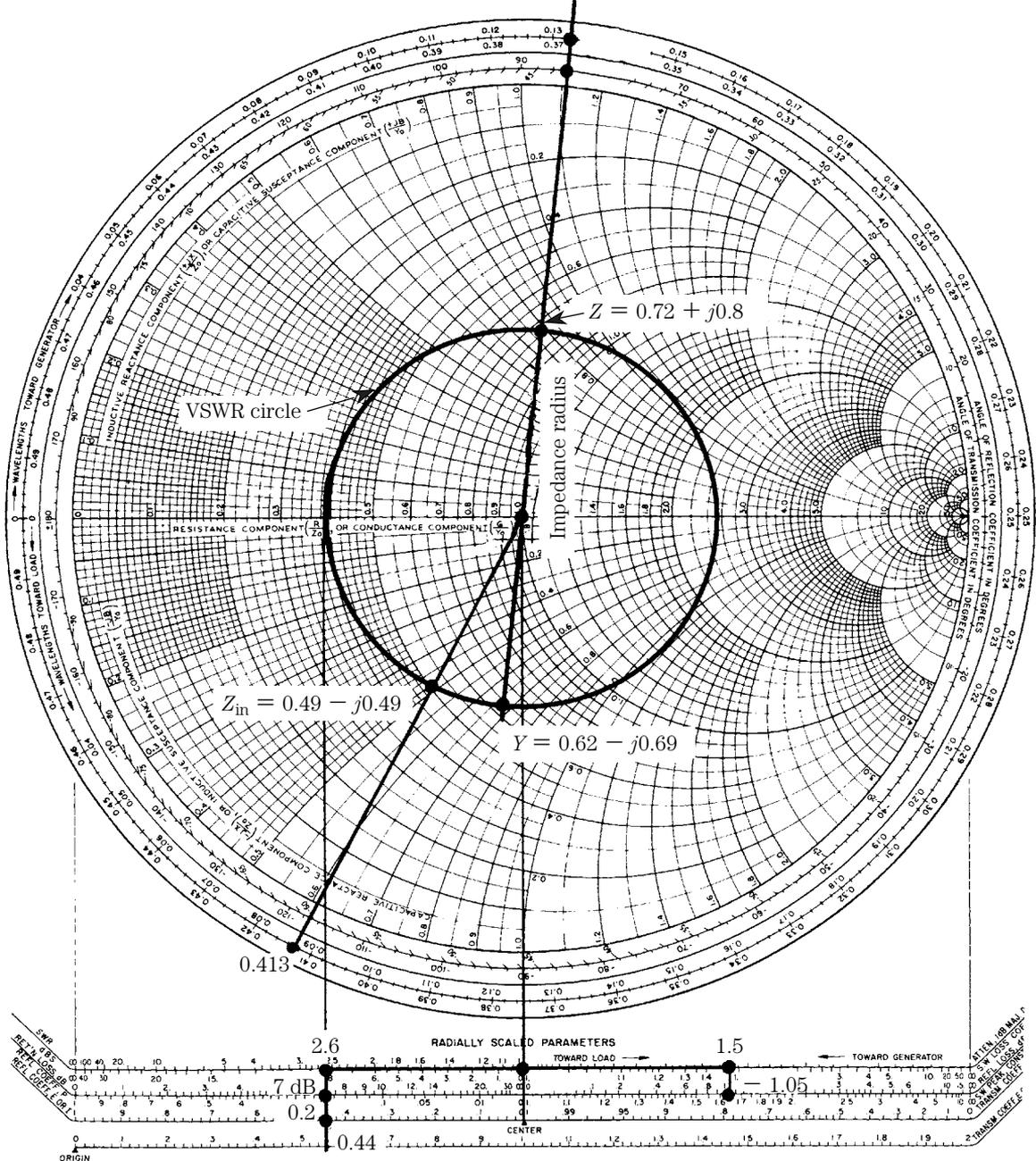
$$= 0.053 \text{ m} \times \frac{100 \text{ cm}}{\text{m}} = 5.3 \text{ cm} \quad [4.41]$$

One wavelength is 5.3 cm, so a half-wavelength is 5.3 cm/2, or 2.65 cm. The 28-cm line is 28 cm/5.3 cm, or 5.28 wavelengths long. A line drawn from the center (1.0) to the load impedance is extended to the outer circle, and it intersects the circle at 0.1325. Because one complete revolution around this circle represents one-half wavelength, 5.28 wavelengths from this point represents 10 revolutions plus 0.28 more. The residual 0.28 wavelengths is added to 0.1325 to form a value of (0.1325 + 0.28), or 0.413. The point 0.413 is located on the circle, and is marked. A line is then drawn from 0.413 to the center of the circle, and it intersects the VSWR circle at  $0.49 - j0.49$ , which represents the input impedance  $Z_{\text{in}}$  looking into the line.

To find the actual impedance represented by the normalized input impedance, you have to “denormalize” the Smith chart impedance by multiplying the result by  $Z_o$ :

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4-7 Solution to example. (Courtesy of Kay Elemtentrics)

$$Z_{\text{in}} = (0.49 - j0.49) (50 \Omega) \quad [4.42]$$

$$= 24.5 - j24.5 \Omega \quad [4.43]$$

It is this impedance that must be matched at the generator by a conjugate matching network.

The admittance represented by the load impedance is the reciprocal of the load impedance, and is found by extending the impedance radius through the center of the VSWR circle until it intersects the circle again. This point is found, and represents the admittance  $Y = 0.62 - j0.69$ . Confirming the solution mathematically:

$$Y = \frac{1}{Z} \quad [4.44]$$

$$= \frac{1}{0.72 + j0.80} \times \frac{0.72 - j0.80}{0.72 - j0.80} \quad [4.45]$$

$$= \frac{0.72 - j0.80}{1.16} = 0.62 - j0.69 \quad [4.46]$$

The VSWR is found by transferring the “impedance radius” of the VSWR circle to the radial scales below. The radius (0.72 – 0.8) is laid out on the VSWR scale (top-most of the radially scaled parameters) with a pair of dividers from the center mark, and we find that the VSWR is approximately 2.6:1. The decibel form of VSWR is 8.3 dB (next scale down from VSWR), and this is confirmed by

$$\text{VSWR}_{\text{dB}} = 20 \log (\text{VSWR}) \quad [4.47]$$

$$= (20) \log (2.7) \quad [4.48]$$

$$= (20) (0.431) = 8.3 \text{ dB} \quad [4.49]$$

The transmission loss coefficient is found in a manner similar to the VSWR, using the radially scaled parameter scales. In practice, once you have found the VSWR you need only drop a perpendicular line from the 2.6:1 VSWR line across the other scales. In this case, the line intersects the voltage reflection coefficient at 0.44. The power reflection coefficient  $\Gamma_{\text{pwr}}$  is found from the scale, and is equal to  $\Gamma^2$ . The perpendicular line intersects the power reflection coefficient line at 0.20.

The angle of reflection coefficient is found from the outer circles of the Smith chart. The line from the center to the load impedance ( $Z = 0.72 + j0.8$ ) is extended to the *angle of reflection coefficient in degrees* circle, and intersects it at approximately 84°. The reflection coefficient is therefore 0.44/84°.

The transmission loss coefficient (TLC) is found from the radially scaled parameter scales also. In this case, the impedance radius is laid out on the *loss coefficient* scale, where it is found to be 1.5. This value is confirmed from:

$$\text{TLC} = \frac{1 + \Gamma_{\text{pwr}}}{1 - \Gamma_{\text{pwr}}} \quad [4.50]$$

$$= \frac{1 + (0.20)}{1 - (0.21)} \quad \mathbf{[4.51]}$$

$$= \frac{1.20}{0.79} = 1.5 \quad \mathbf{[4.52]}$$

The *return loss* is also found by dropping the perpendicular from the VSWR point to the *RET'N LOSS, dB* line, and the value is found to be approximately 7 dB, which is confirmed by

$$\text{Loss}_{\text{ret}} = 10 \log (\Gamma_{\text{pwr}}) \text{ dB} \quad \mathbf{[4.53]}$$

$$= 10 \log (0.21) \text{ dB} \quad \mathbf{[4.54]}$$

$$= (10) (-0.677) \text{ dB} \quad \mathbf{[4.55]}$$

$$= 6.77 \text{ dB} = -6.9897 \text{ dB} \quad \mathbf{[4.56]}$$

The reflection loss is the amount of RF power reflected back down the transmission line from the load. The difference between incident power supplied by the generator (1.5 W in this example),  $P_{\text{inc}} - P_{\text{ref}} = P_{\text{abs}}$  and the reflected power, is the *absorbed power* ( $P_a$ ); or in the case of an antenna, the radiated power. The reflection loss is found graphically by dropping a perpendicular from the TLC point (or by laying out the impedance radius on the *REFL. LOSS, dB* scale), and in this example (Fig. 4-7) is  $-1.05$  dB. You can check the calculations:

The return loss is  $-7$  dB, so

$$-7 \text{ dB} = 10 \log \left( \frac{P_{\text{ref}}}{P_{\text{inc}}} \right) \quad \mathbf{[4.57]}$$

$$-7 = 10 \log \left( \frac{P_{\text{ref}}}{1.5 \text{ watts}} \right) \quad \mathbf{[4.58]}$$

$$\frac{-7}{10} = \log \left( \frac{P_{\text{ref}}}{1.5 \text{ watts}} \right) \quad \mathbf{[4.59]}$$

$$10^{(-7/10)} = \frac{P_{\text{ref}}}{1.5 \text{ W}} \quad \mathbf{[4.60]}$$

$$0.2 = \frac{P_{\text{ref}}}{1.5 \text{ W}} \quad \mathbf{[4.61]}$$

$$(0.2)(1.5 \text{ W}) = P_{\text{ref}} \quad \mathbf{[4.62]}$$

$$0.3 \text{ W} = P_{\text{ref}} \quad \mathbf{[4.63]}$$

The power absorbed by the load ( $P_a$ ) is the difference between incident power  $P_{\text{inc}}$  and reflected power  $P_{\text{ref}}$ . If 0.3 W is reflected, then that means the absorbed power is  $(1.5 - 0.3)$ , or 1.2 W.

The reflection loss is  $-1.05$  dB, and can be checked from:

$$-1.05 \text{ dB} = 10 \log \left( \frac{P_a}{P_{\text{inc}}} \right) \quad [4.64]$$

$$\frac{-1.05}{10} = \log \left( \frac{P_a}{1.5 \text{ W}} \right) \quad [4.65]$$

$$10^{(-1.05/10)} = \frac{P_a}{1.5 \text{ W}} \quad [4.66]$$

$$0.785 = \frac{P_a}{1.5 \text{ W}} \quad [4.67]$$

$$(1.5 \text{ W}) \times (0.785) = P_a \quad [4.68]$$

$$1.2 \text{ W} = P_a \quad [4.69]$$

Now check what you have learned from the Smith chart. Recall that 1.5 W of 4.5-GHz microwave RF signal was input to a  $50\text{-}\Omega$  transmission line that was 28 cm long. The load connected to the transmission line has an impedance of  $36 + j40$ . From the Smith chart:

Admittance (load)	$0.62 - j0.69$
VSWR	2.6:1
VSWR (dB):	8.3 dB
Reflection coefficient ( $E$ )	0.44
Reflection coefficient ( $P$ )	0.2
Reflection coefficient angle	84 degrees
Return loss	$-7$ dB
Reflection loss	$-1.05$ dB
Transmission loss coefficient	1.5

Note that in all cases the mathematical interpretation corresponds to the graphical interpretation of the problem, within the limits of accuracy of the graphical method.

### Stub matching systems

A properly designed matching system will provide a conjugate match to a complex impedance. Some sort of matching system or network is needed any time the load impedance  $Z_L$  is not equal to the characteristic impedance  $Z_o$  of the transmission line. In a transmission line system, it is possible to use a *shorted stub* connected in

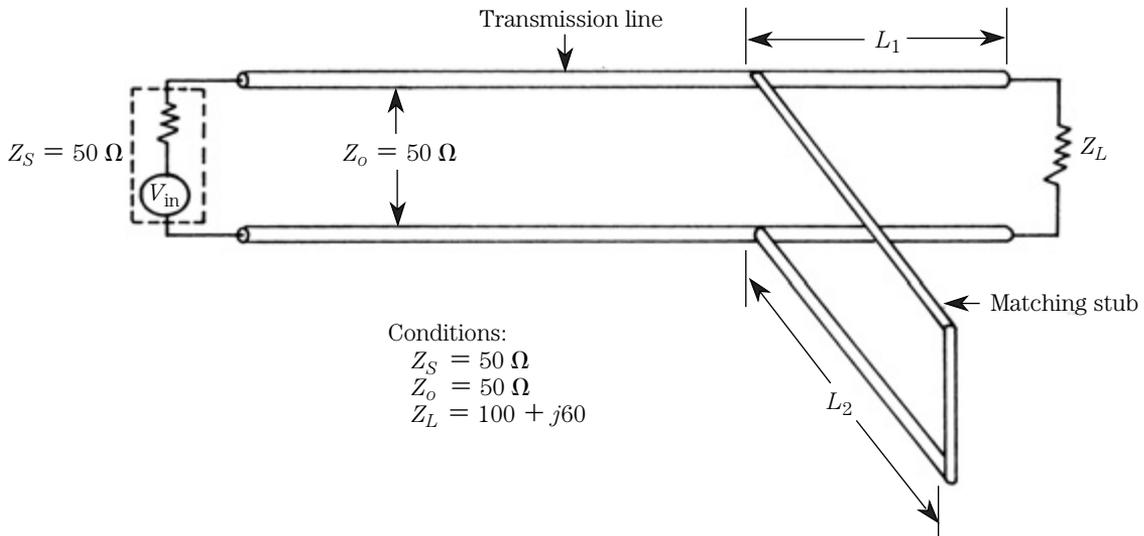
parallel with the line, at a critical distance back from the mismatched load, in order to effect a match. The stub is merely a section of transmission line that is shorted at the end not connected to the main transmission line. The reactance (hence also susceptance) of a shorted line can vary from  $-\lambda$  to  $+\lambda$ , depending upon length, so you can use a line of critical length  $L_2$  to cancel the reactive component of the load impedance. Because the stub is connected in parallel with the line it is a bit easier to work with admittance parameters, rather than impedance.

Consider the example of Fig. 4-8, in which the load impedance is  $Z = 100 + j60$ , which is normalized to  $2.0 + j1.2$ . This impedance is plotted on the Smith chart in Fig. 4-9, and a VSWR circle is constructed. The admittance is found on the chart at point  $Y = 0.37 - j0.22$ .

In order to provide a properly designed matching stub, you need to find two lengths.  $L_1$  is the length (relative to wavelength) from the load toward the generator (see  $L_1$  in Fig. 4-8);  $L_2$  is the length of the stub itself.

The first step in finding a solution to the problem is to find the points where the unit conductance line (1.0 at the chart center) intersects the VSWR circle; there are two such points shown in Fig. 4-9:  $1.0 + j1.1$  and  $1.0 - j1.1$ . Select one of these (choose  $1.0 + j1.1$ ) and extend a line from the center 1.0 point through the  $1.0 + j1.1$  point to the outer circle (*WAVELENGTHS TOWARD GENERATOR*). Similarly, a line is drawn from the center through the admittance point  $0.37 - 0.22$  to the outer circle. These two lines intersect the outer circle at the points 0.165 and 0.461. The distance of the stub back toward the generator is found from:

$$L_1 = 0.165 + (0.500 - 0.461) \lambda \quad [4.70]$$



4-8 Matching stub length and position.



$$= 0.165 + 0.039 \lambda \quad [4.71]$$

$$= 0.204 \lambda \quad [4.72]$$

The next step is to find the length of the stub required. This is done by finding two points on the Smith chart. First, locate the point where admittance is infinite (far right side of the pure conductance line); second, locate the point where the admittance is  $0 - j1.1$ . (Note that the susceptance portion is the same as that found where the unit conductance circle crossed the VSWR circle.) Because the conductance component of this new point is 0, the point will lay on the  $-j1.1$  circle at the intersection with the outer circle. Now draw lines from the center of the chart through each of these points to the outer circle. These lines intersect the outer circle at 0.368 and 0.250. The length of the stub is found from

$$L_2 = (0.368 - 0.250)\lambda \quad [4.73]$$

$$= 0.118\lambda \quad [4.74]$$

From this analysis you can see that the impedance,  $Z = 100 + j60$ , can be matched by placing a stub of a length  $0.118\lambda$  at a distance  $0.204\lambda$  back from the load.

### The Smith chart in lossy circuits

Thus far, you have dealt with situations in which loss is either zero (i.e., ideal transmission lines), or so small as to be negligible. In situations where there is appreciable loss in the circuit or line, however, you see a slightly modified situation. The VSWR circle, in that case, is actually a spiral, rather than a circle.

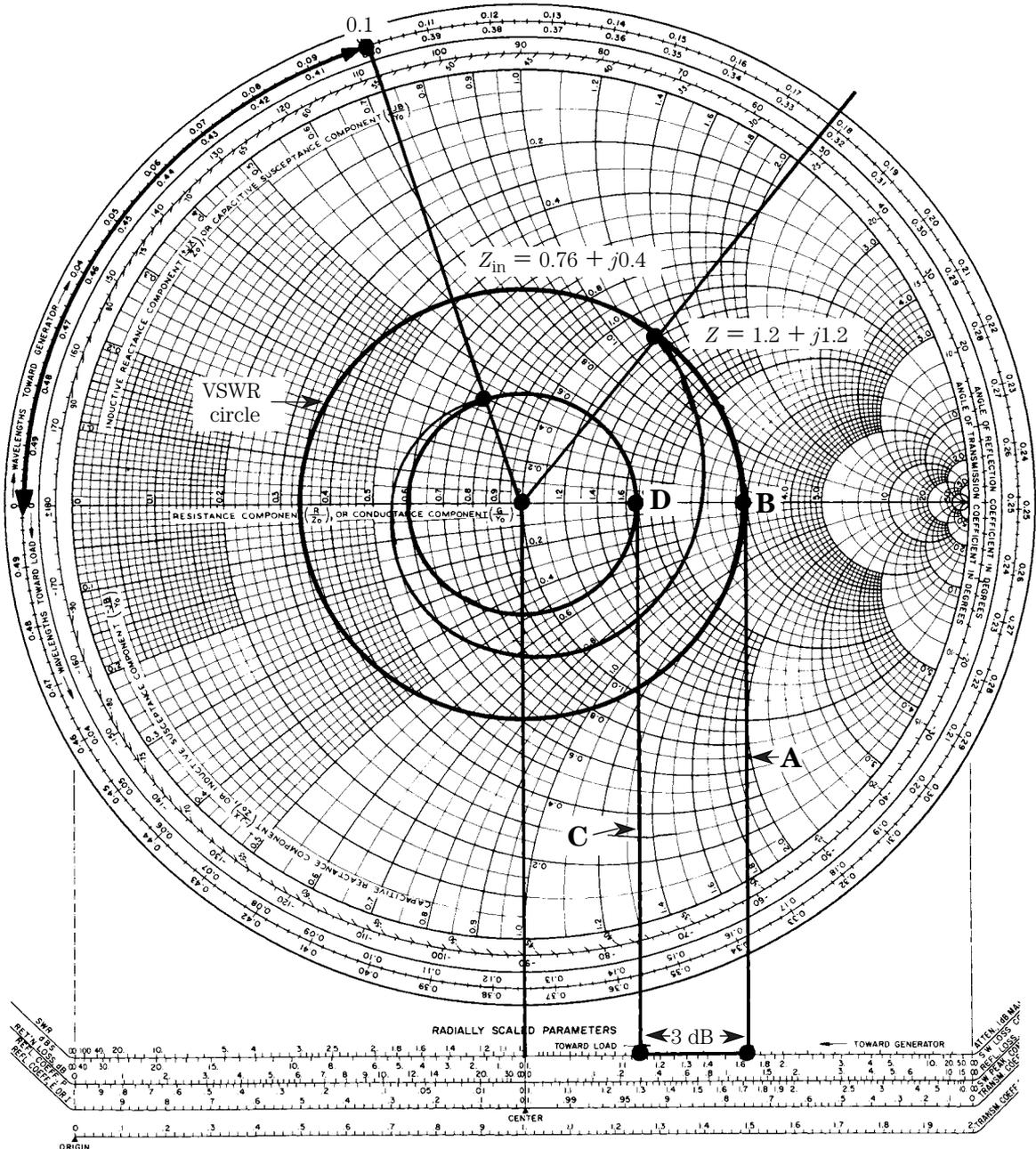
Figure 4-10 shows a typical situation. Assume that the transmission line is  $0.60\lambda$  long, and is connected to a normalized load impedance of  $Z = 1.2 + j1.2$ . An "ideal" VSWR circle is constructed on the impedance radius represented by  $1.2 + j1.2$ . A line (A) is drawn, from the point where this circle intersects the pure resistance baseline (B), perpendicularly to the *ATTEN, 1 dB/MAJ. DIV.* line on the radially scaled parameters. A distance representing the loss (3 dB) is stepped off on this scale. A second perpendicular line is drawn, from the  $-3$  dB point, back to the pure resistance line (C). The point where line C intersects the pure resistance line becomes the radius for a new circle that contains the actual input impedance of the line. The length of the line is  $0.60\lambda$ , so you must step back  $(0.60 - 0.50)\lambda$  or  $0.1\lambda$ . This point is located on the *WAVELENGTHS TOWARD GENERATOR* outer circle. A line is drawn from this point to the 1.0 center point. The point where this new line intersects the new circle is the actual input impedance ( $Z_{in}$ ). The intersection occurs at  $0.76 + j0.4$ , which (when denormalized) represents an input impedance of  $38 + j20 \Omega$ .

### Frequency on the Smith chart

A complex network may contain resistance, inductive reactance, and capacitive reactance components. Because the reactance component of such impedances is a function of frequency, the network or component tends to also be frequency-sensitive.

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4-10 Solution. (Courtesy of Kay Eletronics)

You can use the Smith chart to plot the performance of such a network with respect to various frequencies. Consider the load impedance connected to a 50-Ω transmission line in Fig. 4-11. In this case, the resistance is in series with a 2.2-pF capacitor, which will exhibit a different reactance at each frequency. The impedance of this network is

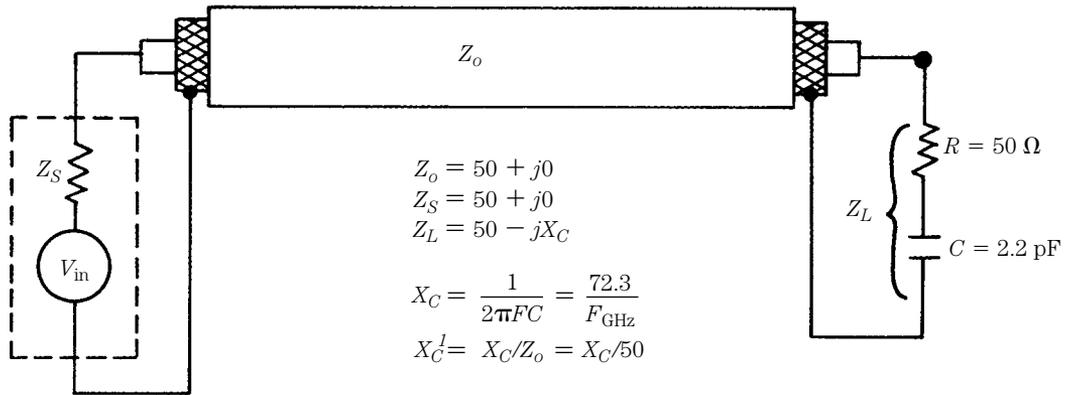
$$Z = R - j\left(\frac{1}{\omega C}\right) \quad [4.75]$$

or

$$Z = 50 - j\left(\frac{1}{2\pi FC}\right) \quad [4.76]$$

And, in normalized form,

$$Z' = 1.0 - \frac{j}{(2\pi FC) \times 50} \quad [4.77]$$

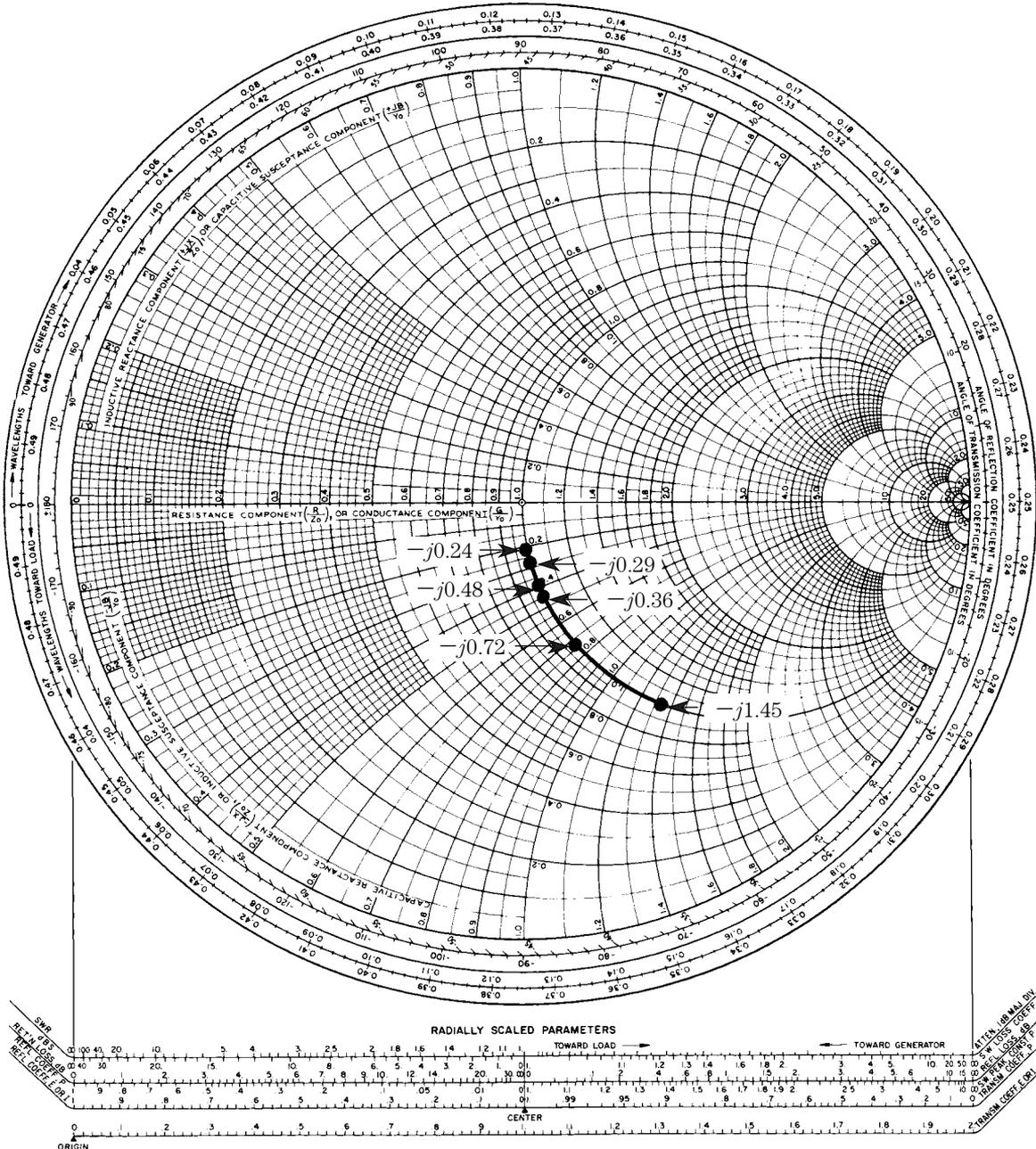


Freq. (GHz)	$X_C$	
1	-j72.3	-j1.45
2	-j36.2	-j0.72
3	-j24.1	-j0.48
4	-j18	-j0.36
5	-j14.5	-j0.29
6	-j12	-j0.24

4-11 Load and source impedance transmission line circuit.

NAME	TITLE	DWG. NO.
SMITH CHART FORM 82-BSPR(9-66)	KAY ELECTRIC COMPANY, PINE BROOK, N. J. © 1966. PRINTED IN U.S.A.	DATE

IMPEDANCE OR ADMITTANCE COORDINATES



4-12 Solution. (Courtesy of Kay Eletronics)

$$= 1.0 - \frac{j}{6.9 \times 10^{-10} F} \quad [4.78]$$

$$= 1.0 - \frac{j \times 7.23 \times 10^{10}}{F} \quad [4.79]$$

Or, converted to gigahertz,

$$Z' = 1.0 - \frac{j72.3}{F_{\text{GHz}}} \quad [4.80]$$

The normalized impedances for the sweep of frequencies from 1 to 6 GHz are therefore

$$Z = 1.0 - j1.45 \quad [4.81]$$

$$Z = 1.0 - j0.72 \quad [4.82]$$

$$Z = 1.0 - j0.48 \quad [4.83]$$

$$Z = 1.0 - j0.36 \quad [4.84]$$

$$Z = 1.0 - j0.29 \quad [4.85]$$

$$Z = 1.0 - j0.24 \quad [4.86]$$

These points are plotted on the Smith chart in Fig. 4-12. For complex networks, in which both inductive and capacitive reactance exist, take the difference between the two reactances (i.e.,  $X = X_L - X_C$ ).

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## 5 CHAPTER

# Fundamentals of radio antennas

AN UNFORTUNATE OVERSIGHT IN MANY BOOKS ON RADIO ANTENNAS IS A LACK OF coverage on the most basic fundamentals of antenna theory. Most books, including the first draft of this one, start with a discussion of dipoles, but overlook that certain physical mechanisms are at work. An antenna is basically a *transducer* that converts electrical alternating current oscillations at a radio frequency to an electromagnetic wave of the same frequency. This chapter looks at the physics of how that job is accomplished.

The material in this chapter was adapted from a U.S. Army training manual on antennas and radio propagation. Although unfortunately no longer in print, the manual contained the best coverage of basics the author of this book could find. Given that U.S. government publications are not protected by copyright, this information can be brought to you in full.

### Antenna fundamentals

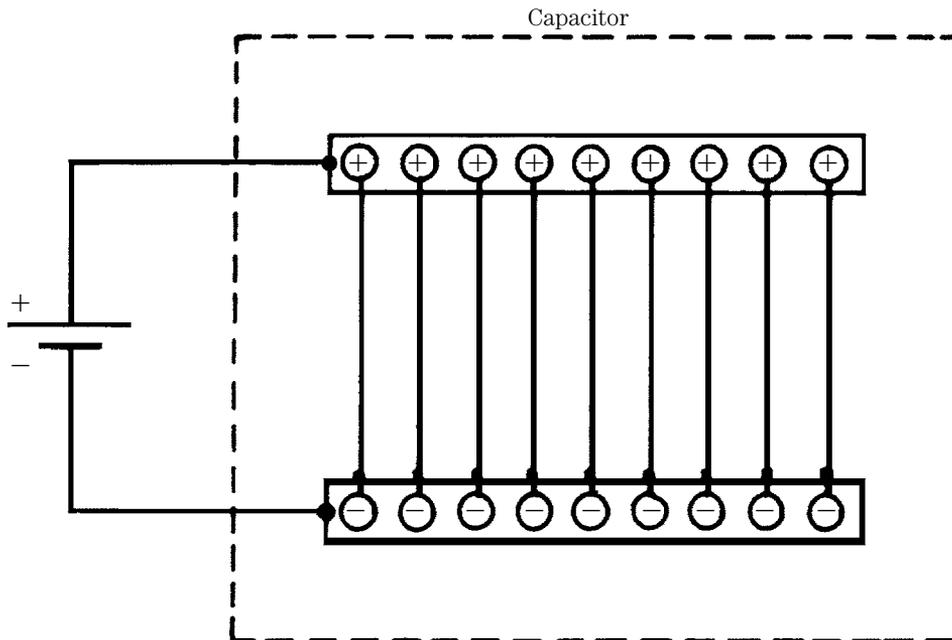
The electric and magnetic fields radiated from an antenna form the electromagnetic field, and this field is responsible for the transmission and reception of electromagnetic energy through free space. An antenna, however, is also part of the electrical circuit of a transmitter (or a receiver); because of its distributed constants, it acts as a circuit containing inductance, capacitance, and resistance. Therefore, it can be expected to display definite voltage and current relationships in respect to a given input. A current through it produces a magnetic field, and a charge on it produces an electrostatic field. These two fields taken together form the induction field. To gain a better understanding of antenna theory, a review of the basic electrical concepts of voltage and electric field and of current and magnetic field is necessary.

### Voltage and electric field

When a capacitor is connected across a source of voltage, such as a battery (Fig. 5-1), it is charged some amount, depending on the voltage and the value of capacitance. Because of the emf (electromotive force) of the battery, negative charges flow to the lower plate, leaving the upper plate positively charged. Accompanying the accumulation of charge is the building up of the electric field. The flux lines are directed from the positive to the negative charges and at right angles to the plates.

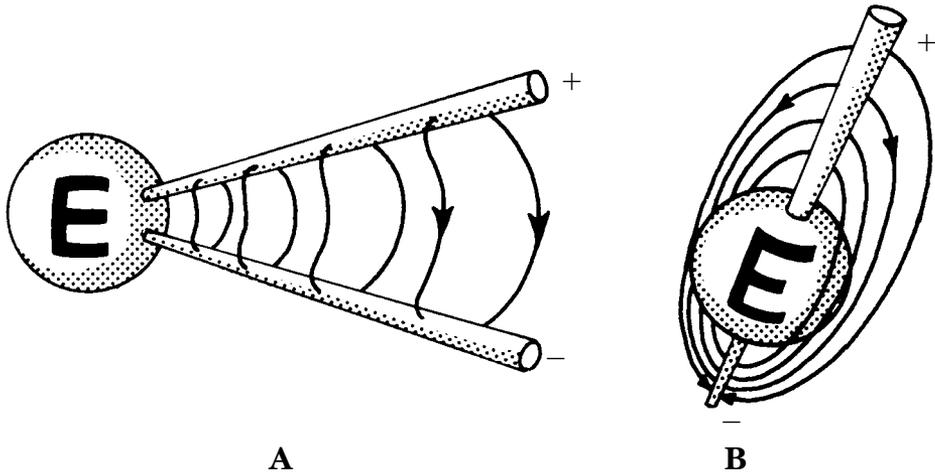
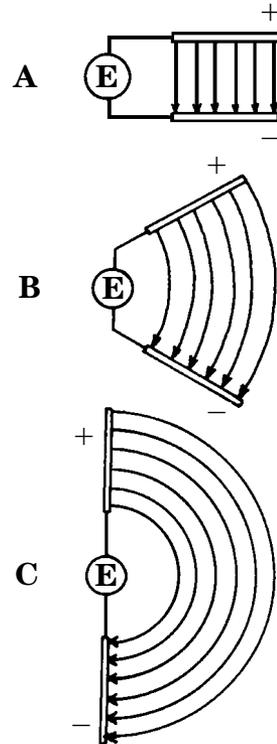
If the two plates of the capacitor are spread farther apart, the electric field must curve to meet the plates at right angles (Fig. 5-2). The straight lines in A become arcs in B, and approximately semicircles in C, where the plates are in a straight line. Instead of flat metal plates, as in the capacitor, the two elements can take the form of metal rods or wires. The three-dimensional view in Fig. 5-3 depicts the electric field more accurately. In A of Fig. 5-3, the wires are approximately  $30^\circ$  apart, and the flux lines are projected radially from the positively charged wire to the negatively charged wire. In B of Fig. 5-3, the two wires lie in a straight line, and the flux lines form a pattern similar to the lines of longitude around the earth. To bring out the picture more clearly, only the lines in one plane are given.

Assume that the sphere marked *E* in Fig. 5-3B is a transmitter supplying RF energy. The two wires then can serve as the antenna for the transmitter. RF energy is radiated from the antenna and charges move back and forth along the wires, alternately compressing and expanding the flux lines of the electric field. The reversals in polarity of the transmitter signal also reverse the direction of the electric field.



5-1 Charges on plates of a capacitor.

5-2 Electric field between plates at various angles.



5-3 Electric field between wires at various angles.

When a charge is put on the plates of a capacitor by means of a battery, an electric field is set up between its plates. The flow of charge from source to capacitor ceases when the capacitor is fully charged, and the capacitor is said to be charged to a *voltage* equal, and of opposite polarity, to that of the source. The charged capacitor

can be used as a source of emf since it stores energy in the form of an electric field. This is the same as saying that *an electric field indicates voltage*. The presence of an electric field about an antenna also indicates voltage. Since the polarity, and the amount of charge, depend on the nature of the transmitter output, the antenna voltage also depends on the energy source. For example, if a battery constitutes the source, the antenna charges to a voltage equal and opposite to that of the battery. If RF energy is supplied to a half-wave antenna, the voltage across the antenna lags the current by  $90^\circ$ . The half-wave antenna acts as if it was a capacitor, and it can be described as being *capacitive*.

### **Current and magnetic field**

A *moving* charge along a conductor constitutes a current and produces a magnetic field around the conductor. Therefore, the flow of charge along an antenna also will be accompanied by a magnetic field. The intensity of this field is directly proportional to the flow of charge. When the antenna is uncharged, the current flow is maximum, since there is no opposing electric field. Because of this current flow, a charge accumulates on the antenna, and an electric field builds up in increasing opposition to the emf of the source. The current flow decreases and when the antenna is fully charged, the current no longer flows.

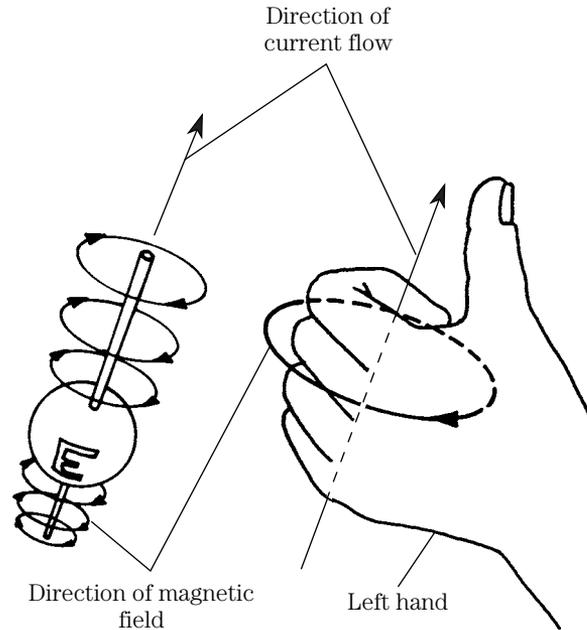
The magnetic field in the space around a current-carrying device has a specific configuration, with the magnetic flux lines drawn according to a definite rule (Fig. 5-4). Whereas, in the electric field, the electric lines are drawn from a positive charge to a negative charge, in the magnetic field, the flux lines are drawn according to the *left-hand rule*. The direction of current flow is upward along both halves of the antenna. The lines of magnetic flux form concentric loops that are perpendicular to the direction of current flow. The arrowheads on the loops indicate the direction of the field. If the thumb of the left hand is extended in the direction of current flow and the fingers are clenched, then the rough circles formed by the fingers indicate the direction of the magnetic field. This is the left-hand rule, or convention, which is used to determine the direction of the magnetic field.

### **Combined electric and magnetic fields**

When RF energy from a transmitter is supplied to an antenna, the effects of charge, voltage, current, and the electric and magnetic fields are taking place simultaneously. These effects (Fig. 5-5) have definite time and space relationships to each other. If a half-wave antenna is used, the relations between charge and current flow can be predicted, because of the capacitive nature of this antenna. The voltage will lag the current by  $90^\circ$ , and the electric and magnetic fields will be  $90^\circ$  out of phase. With no electric field present (no charge), the current flow is unimpeded, and the magnetic field is maximum. As charge accumulates on the antenna, the electric field builds up in opposition to current flow and the magnetic field decreases in intensity. When the electric field reaches its maximum strength, the magnetic field has decayed to zero.

A reversal in polarity of the source reverses the direction of current flow as well as the polarity of the magnetic field, and the electric field aids the flow of current by

5-4 Magnetic field about a half-wave antenna (left-hand rule).

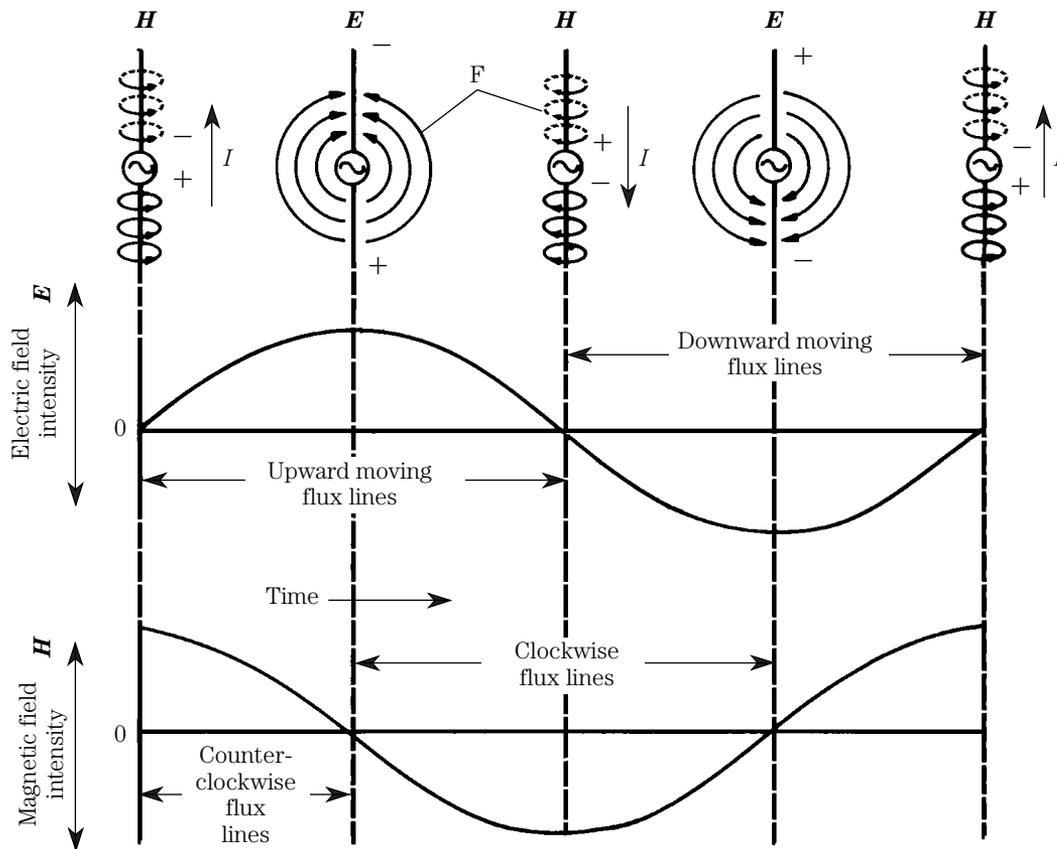


discharging. The magnetic field builds up to a maximum, and the electric field disappears as the charge is dissipated. The following half-cycle is a repetition of the first half-cycle, but in the reverse direction. This process continues as long as energy is supplied to the antenna. The fluctuating electric and magnetic fields combine to form the induction field, in which the electric and magnetic flux maximum intensities occur  $90^\circ$  apart in time, or in *time quadrature*. Physically, they occur at right angles to each other, or in *space quadrature*. To sum up, the electric and magnetic fields about the antenna are in space and time quadrature.

### Standing waves

Assume that it is possible to have a wire conductor with one end extending infinitely, with an RF transmitter connected to this wire. When the transmitter is turned on, an RF current in the form of sine waves of RF energy moves down the wire. These waves of energy are called *traveling waves*. The resistance of the conductor gradually diminishes the amplitude of the waves, but they continue to travel so long as the line does not come to an end.

The antenna, however, has some finite length. Therefore, the traveling waves are halted when they reach the end of the conductor. Assume that the RF transmitter is turned on just long enough to allow one sine wave of energy to get on the line (Fig. 5-6A). This traveling wave is moving down the antenna toward the end. When the wave reaches the end of the conductor, the current path is broken abruptly. With the stoppage of current flow, the magnetic field collapses. A voltage is induced at the end of the conductor that causes current to flow *back toward the source*, as in Fig. 5-6B. The wave is reflected back to the source, and, if a continual succession of

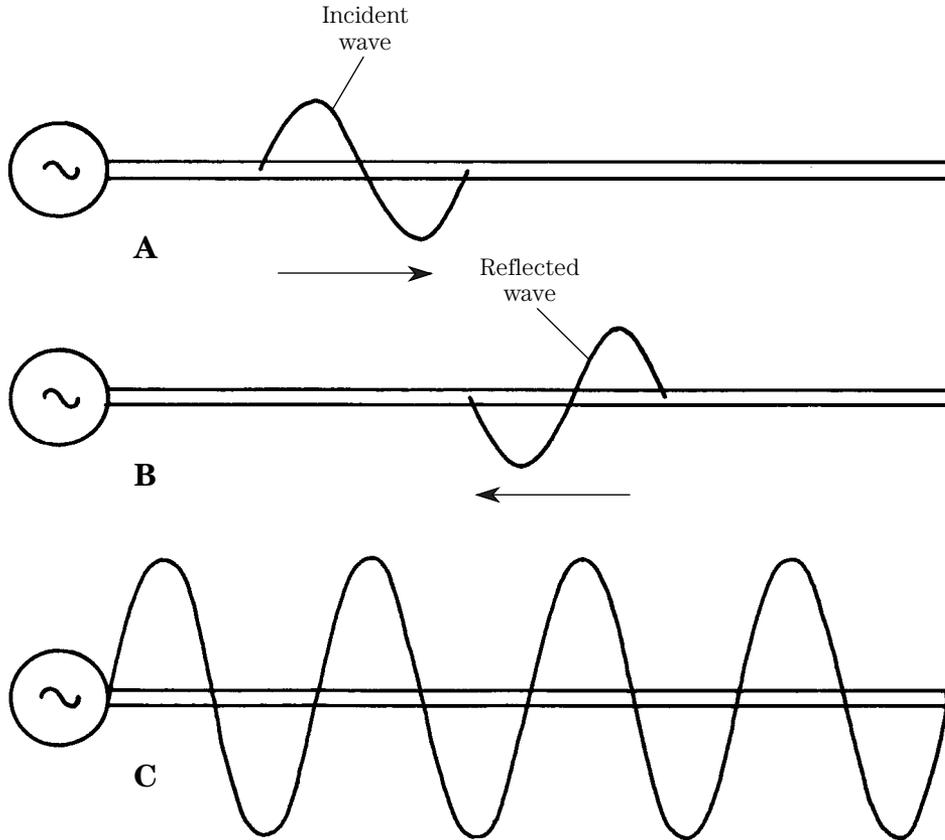
5-5 Electric and magnetic fields  $90^\circ$  out of phase.

waves is sent down the line, waves will be reflected in the same continual pattern. The wave moving from the transmitter toward the end is called the *incident* wave, and its reflection is called the *reflected* wave.

A continuous flow of incident waves results in a continuous flow of reflected waves. Because there is only one conductor, the two waves must pass each other. Electrically, the only current that actually flows is the *resultant* of both of these waves. The waves can reinforce or cancel each other as they move.

When they reinforce, the resultant wave is maximum; when they cancel, the resultant wave is minimum. In a conductor that has a finite length, such as an antenna, the points at which the maxima and minima of the resultant wave occur (Fig. 5-6C) are *stationary*. In other words, the maximum and minimum points stand still, although both the incident and reflected waves are moving. The resultant wave stands still on the line, only its amplitude being subject to change. Because of this effect, the resultant is referred to as a *standing wave*.

The development of the standing wave on an antenna by actual addition of the traveling waves is illustrated in Fig. 5-7. At the instant pictured in A, the incident and reflected waves just coincide. The result is a standing wave having twice the amplitude of either traveling wave. In B, the waves move apart in opposite directions, and



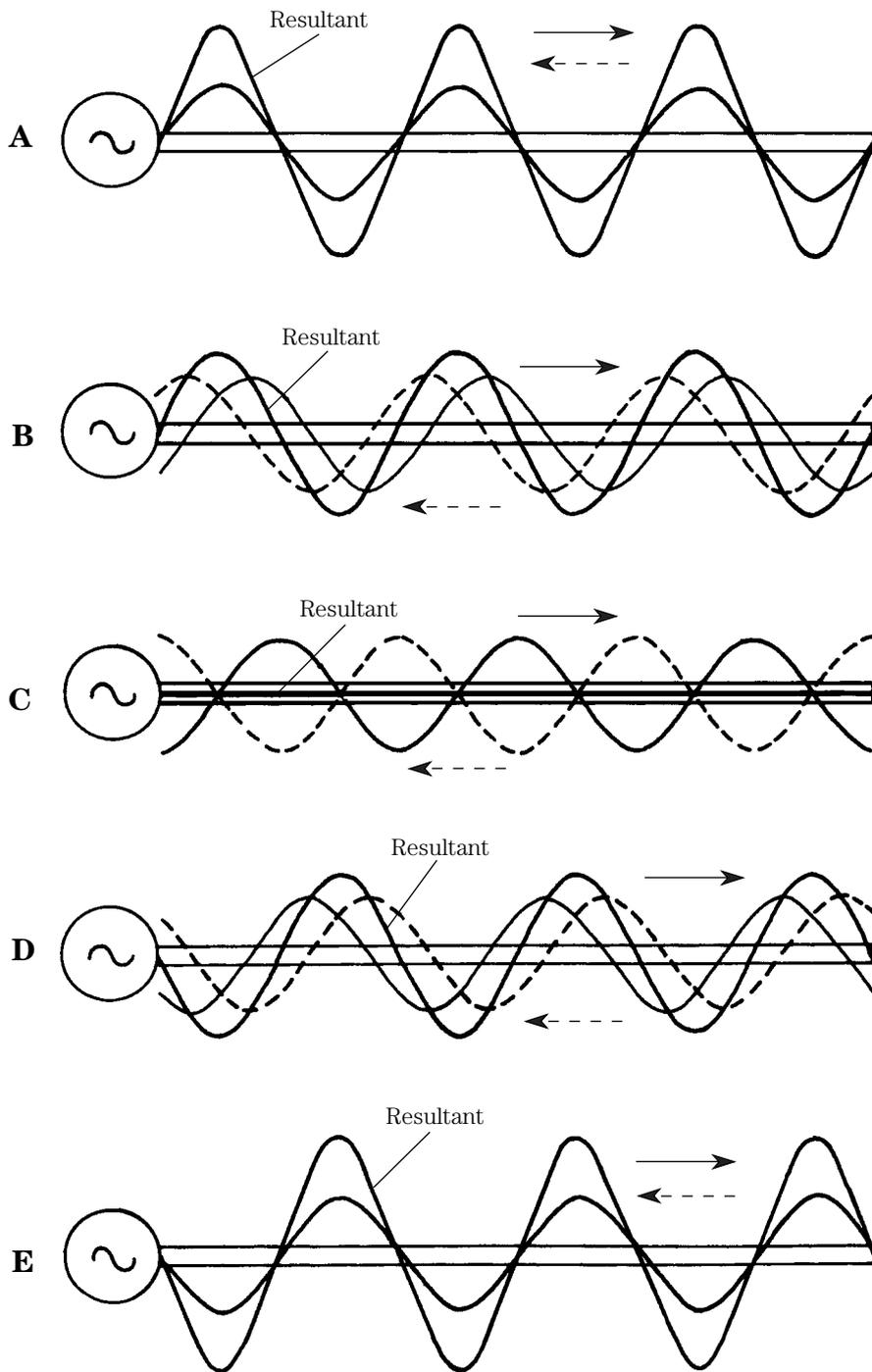
5-6 Traveling waves on an antenna and a typical resultant wave.

the amplitude of the resultant decreases, but the points of maximum and minimum do not move.

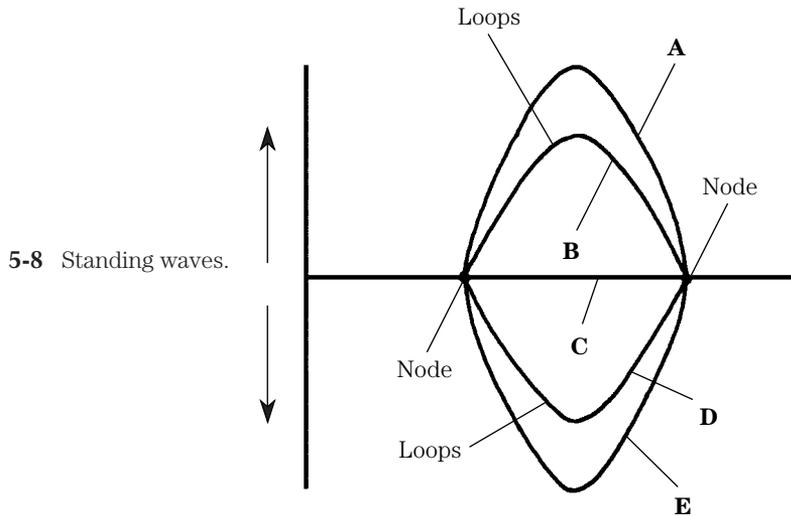
When the traveling waves have moved to a position of  $180^\circ$  phase difference, the resultant is zero along the entire length of the antenna, as shown in C. At this instant, there can be no current flow in the antenna. The continuing movement of the traveling waves, shown in D, builds up a resultant in a direction opposite to that in A. The in-phase condition of the traveling waves results in a standing wave, in E, equal in amplitude, but  $180^\circ$  out of phase with the standing wave in A.

If the progressive pictures of the standing wave are assembled on one set of axes, the result is as in Fig. 5-8. The net effect of the incident and reflected waves is apparent. The curves are lettered with reference to Fig. 5-7. As the traveling waves move past each other, the standing wave changes only its amplitude. The fixed minimum points are called *nodes*, and the curves representing the amplitude are called *loops*.

The concept of the standing wave can be applied to the half-wave antenna with reference to either current or voltage distribution at any instant. This application is possible because there are traveling waves of both voltage and current. Because voltage and current are out of phase on the half-wave antenna, the standing waves also are found to be out of phase.



5-7 Development of standing wave from traveling wave.



### Voltage and current distribution on half-wave antenna

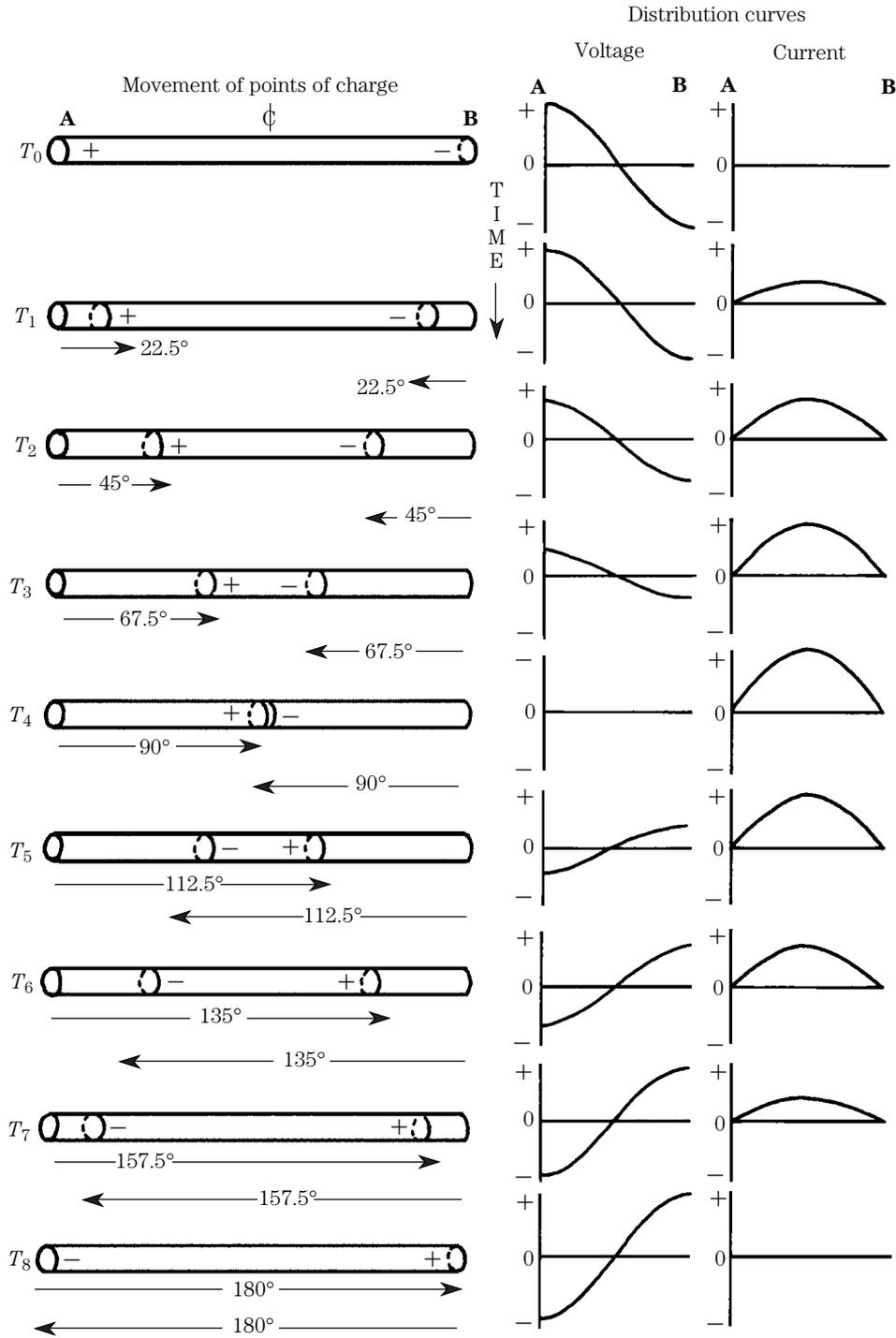
When an RF transmitter is feeding a half-wave antenna, positive and negative charges move back and forth along the antenna (Figs. 5-9 and 5-10). The first picture shows the position of the charges at some arbitrary time,  $T_0$ . The RF charges being observed are at the ends of the antenna, and there is a maximum difference in potential between the ends,  $A$  and  $B$ . The remaining illustrations show the instantaneous positions of the charges at regular intervals of  $22.5^\circ$  throughout a complete cycle.

To the right of each instantaneous position of the charges are curves representing the current and voltage *at that particular time* for any point on the antenna. For example, at time  $T_0$ , the positive and negative charges are at points  $A$  and  $B$  on the antenna. The voltage between these points represents a maximum difference of potential. The current, being  $90^\circ$  out of phase in respect to the voltage, is everywhere zero. These distribution curves are *standing waves* derived in the same manner as those covered in the previous paragraph.

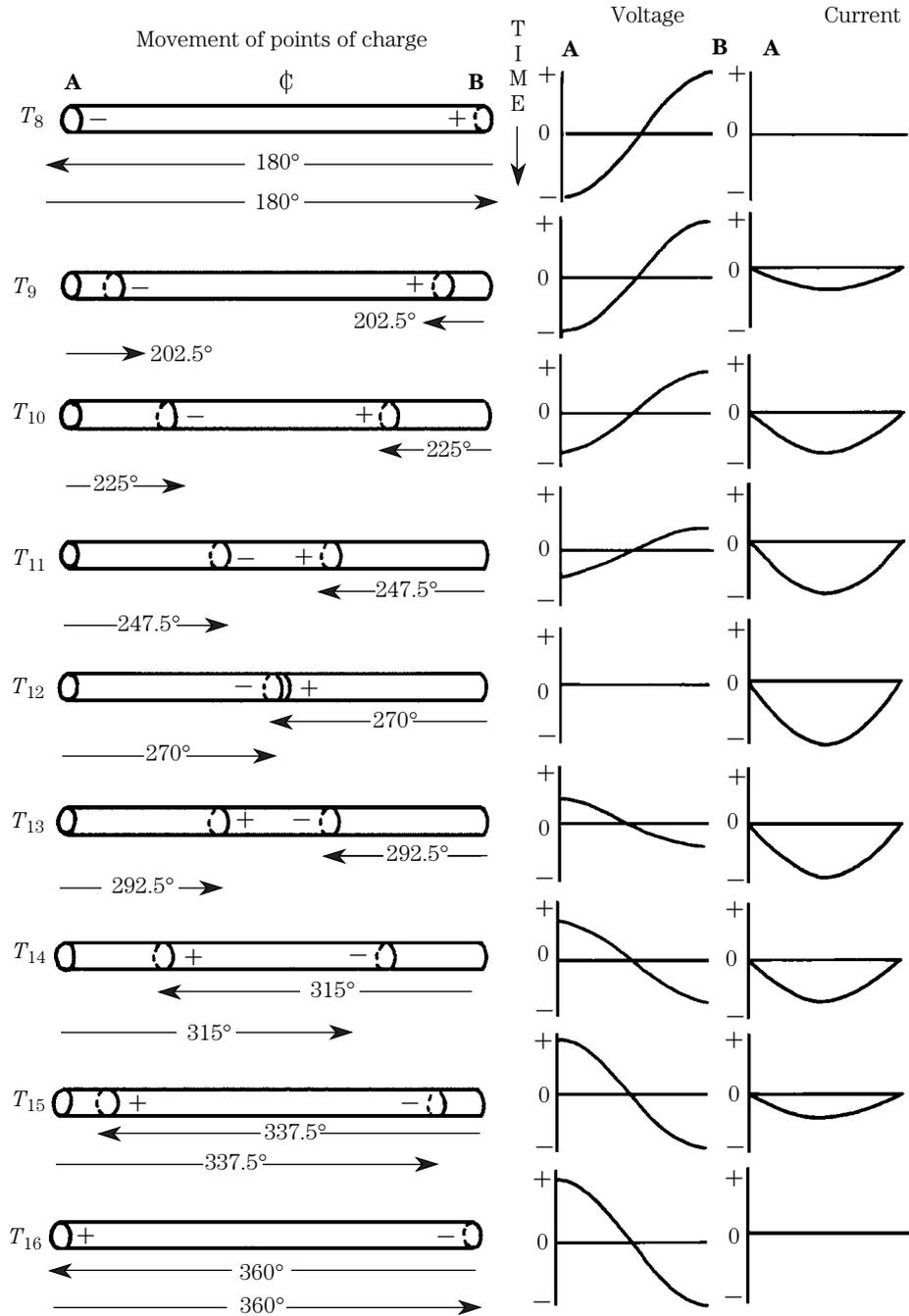
The next illustration shows the position of the charges at time  $T_1$ . The standing wave of current is a relative maximum at the center of the antenna. This current loop has nodes that remain at the ends of the antenna, and it is, therefore,  $90^\circ$  out of phase with the standing wave of voltage.

At  $T_2$  and  $T_3$ , the charges move closer together, and the standing wave of voltage slowly decreases in amplitude. Conversely, the current loop increases in magnitude. When the charges meet after  $90^\circ$  of the RF cycle ( $T_4$ ), the effect is that of having the positive and negative charges cancel. The voltage loop accordingly is zero everywhere on the antenna, and the current loop rises to its maximum value, unimpeded by any charge on the antenna.

At time  $T_5$ , the charges have passed each other, each charge having moved past the center point of the antenna. The polarity of the voltage loops is reversed, and they build up in the opposite direction, keeping the node always at the center point of the antenna. The reversal of polarity is shown in the charge positions at  $T_3$ ,  $T_4$ , and



5-9 Voltage and current distribution in terms of positive and negative charges.



5-10 Voltage and current distribution in terms of positive and negative charges.

$T_5$ . The separation of the charges also is accompanied by a decrease in the amplitude of the current loop.

From  $T_5$  to  $T_8$ , the charges move out to the ends of the antenna. During this time, the voltage loops increase and the current loops decrease in amplitude. At time  $T_8$ , which occurs  $180^\circ$  after  $T_0$  in the RF cycle, the charges have moved to opposite ends of the antenna. Compare the picture in  $T_0$  to the picture in  $T_8$ . It is seen that the negative charge is now at point  $A$  and the positive charge at point  $B$ . Because the positions of the charges have been reversed from  $T_0$  to  $T_8$ , the voltage loops in  $T_8$  are  $180^\circ$  out of phase, compared with the loops in  $T_0$ .

From  $T_8$  to  $T_{16}$  in Fig. 5-10, the movement of the charges is shown in the opposite direction, the current loop reaching a maximum at  $T_{12}$ . When the entire RF cycle is completed at time  $T_{16}$ , the charges have returned to the positions that they occupied at  $T_0$ . The distribution curves of voltage and current also are in their original conditions. The entire process then is repeated for each RF cycle.

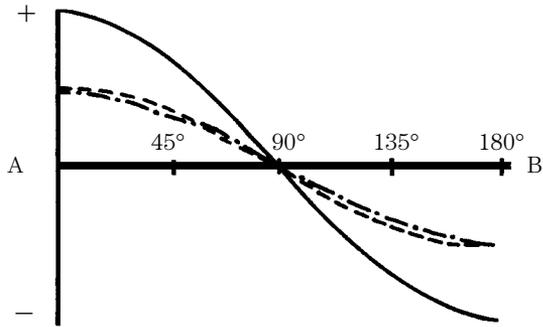
**Standing waves of voltage and current** The distribution curves of the current and voltage are standing waves. This means that they are the resultants obtained by adding two traveling waves. The two traveling waves are associated with the positive and negative charges. The wave caused by the negative charge can be called the *incident wave* and the wave caused by the positive charge the *reflected wave*. This, however, is clearer when the concept of negative and positive charges is used.

The positive charge, taken at time  $T_0$  in Fig. 5-9, produces a traveling wave of voltage, shown by the dashed line in Fig. 5-11A. The negative charge at the opposite end of the antenna produces an identical traveling wave (dash-dot curve). These two add together to produce the  $T_0$  voltage distribution curve, which is the resultant wave of Fig. 5-9A. Both of these waveforms are identical, being the *standing wave of voltage* at time  $T_0$ . All of the following distribution curves of Fig. 5-11 are produced in the same manner. They are the standing wave resultants caused by the traveling waves accompanying the charges.

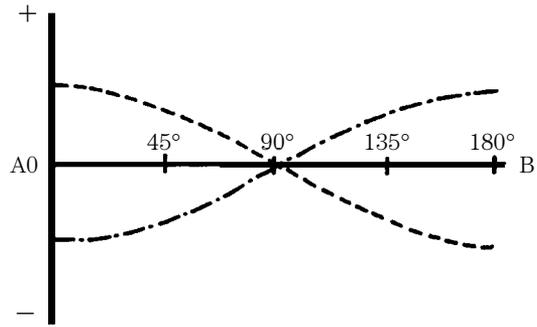
In Fig. 5-11B, each of the traveling waves has moved  $45^\circ$ , the positive traveling wave moving to the right and the negative traveling wave moving to the left. This time corresponds to  $T_2$  in Fig. 5-9. The standing wave produced corresponds to the voltage distribution curve at  $T_2$ . The standing waves of current are produced in the same manner. The current curves at D, E, and F of Fig. 5-11 correspond to times  $T_0$ ,  $T_2$ , and  $T_4$  of Fig. 5-9.

**Standing waves of voltage** In Fig. 5-12A, voltage standing waves occurring at different times are brought together on one axis,  $AB$ , representing a half-wave antenna. Essentially, these are the same curves shown progressively in Figs. 5-9 and 5-10 as voltage distribution curves. They can be used to determine the voltage at any point on the antenna, at any instant of time. For example, if it is desired to know the variations of voltage occurring at point  $Y$  on the antenna over the RF cycle, the variations are graphed in respect to time, as shown in Fig. 5-12B. At  $T_0$ , the voltage at  $Y$  is maximum. From  $T_0$  through  $T_3$ , the voltage decreases, passing through zero at  $T_4$ . The voltage builds up to a maximum in the opposite direction at  $T_8$ , returning through zero to its original position from  $T_8$  to  $T_{16}$ .

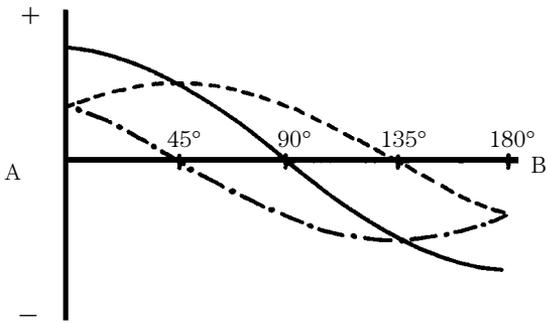
Between  $T_0$  and  $T_{16}$ , therefore, an entire sine-wave cycle,  $Y$ , is reproduced. This is also true of any other point on the antenna, with the exception of the node at  $X$ .



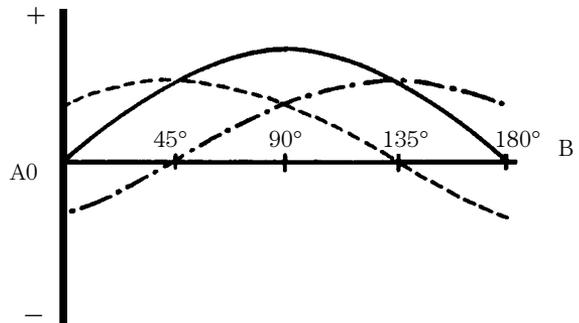
**A**



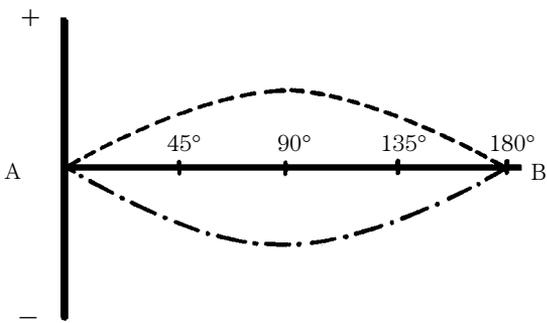
**D**



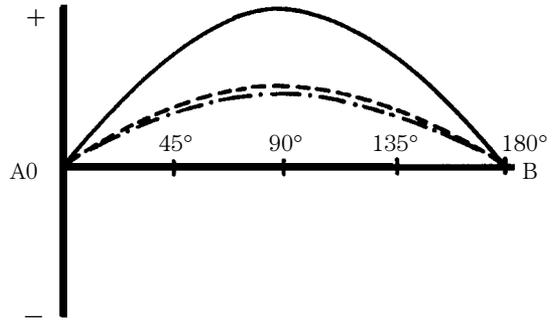
**B**



**E**



**C**



**F**

Voltage waves

Current waves

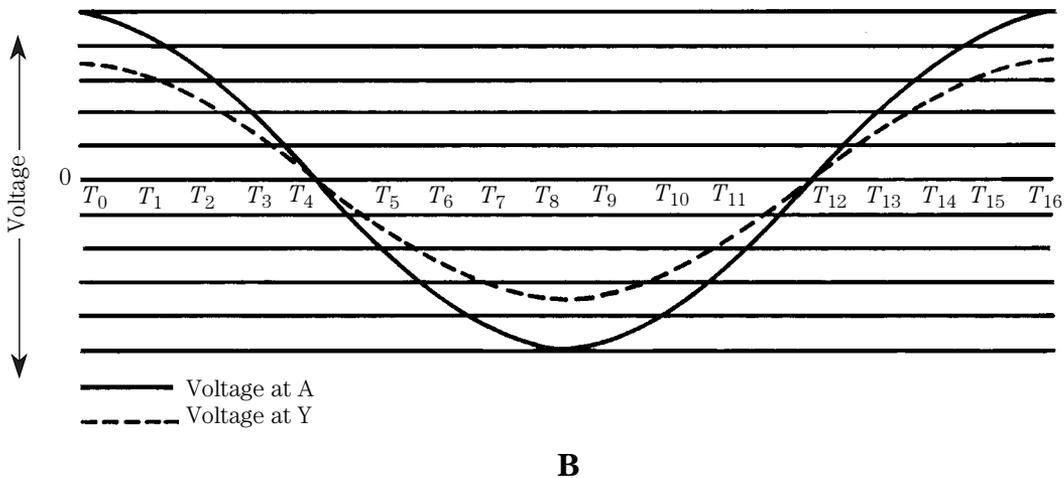
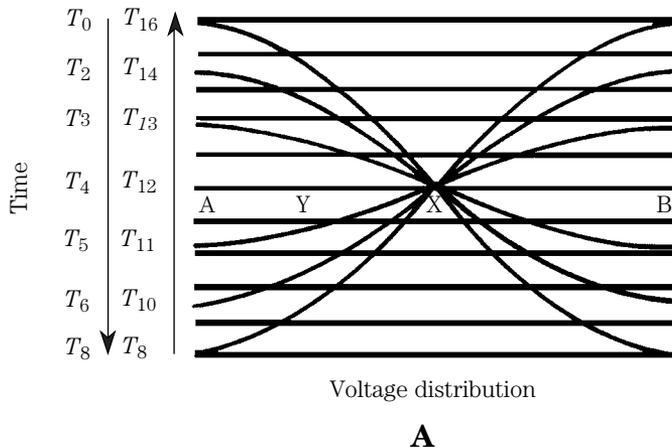
In A, B, and C

In D, E, and F

- Voltage wave due to point of + charge
- - - - Voltage wave due to point of - charge
- Resultant voltage wave

- Current wave due to point of + charge
- - - - Current wave due to point of - charge
- Resultant current wave

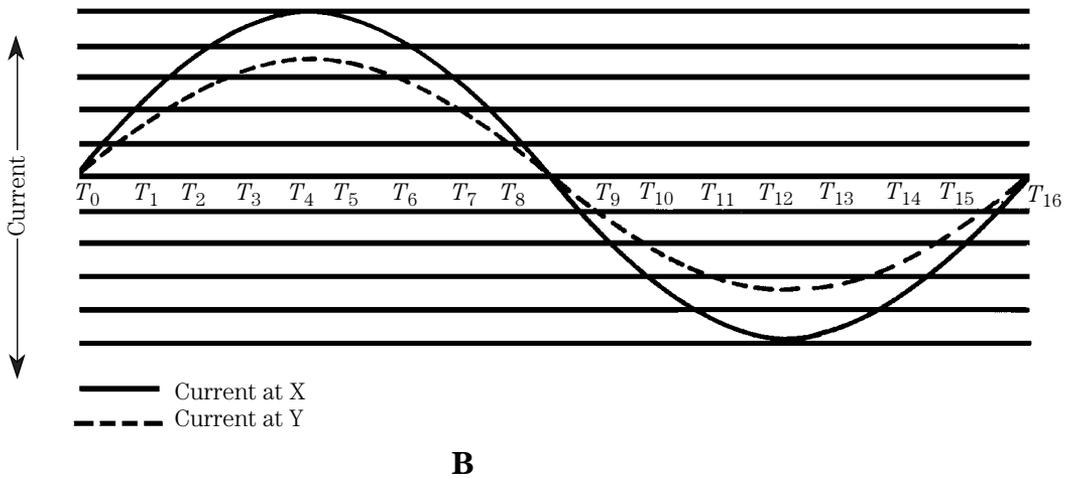
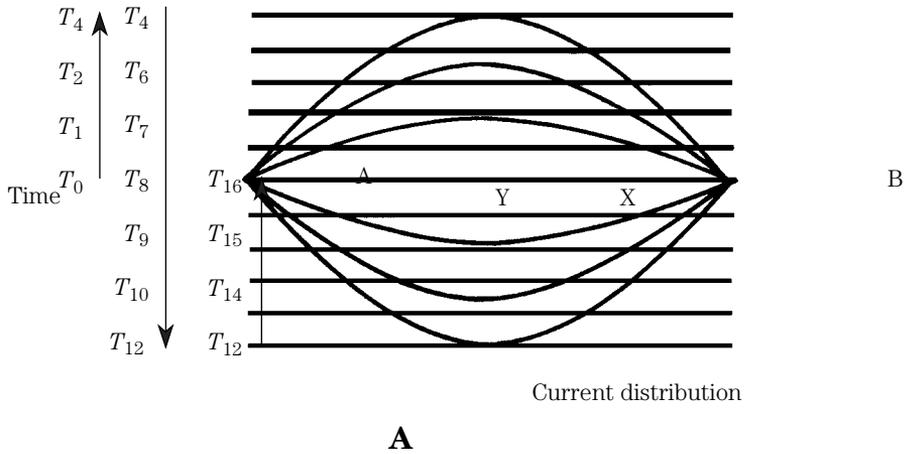
**5-11** Standing waves of voltage and current.



5-12 Standing waves of voltage at a point on the antenna.

The peak amplitude of the sine wave produced at any point depends on its position on the antenna. The nearer the point is to either end, the greater its peak amplitude.

**Standing waves of current** The standing waves of current occurring at various times through the RF cycle are assembled on a single axis in Fig. 5-13. This axis, AB, represents the half-wave antenna. If the current variations at point Y from  $T_0$  to  $T_{16}$  are graphed in respect to time, the result is the sine wave in Fig. 5-13B. This is true for any point along the antenna with the exception of the nodes at the ends. The current has its greatest swing at X, the center of the antenna. Comparison of the voltage variation curve (Fig. 5-12A) with the current variation curve (Fig. 5-13A) shows the voltage curve leading the current curve by  $90^\circ$  at Y. This relation can be expected on any half-wave device.

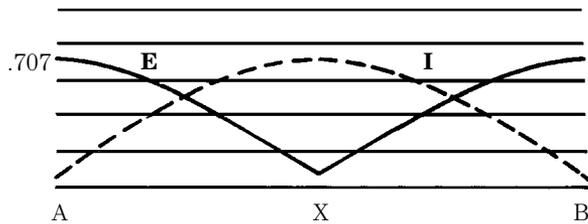


5-13 Standing waves of current at a point on the antenna.

**Measurement of standing waves** In Fig. 5-14, the standing waves of voltage  $E$  and current  $I$  are indicated along the antenna. There are current nodes at  $A$  and  $B$  and a voltage node at  $X$ . These standing waves are found on any half-wave antenna. A meter that indicates the effective value (0.707 of peak) of the ac signal can be used to measure the standing waves present on the half-wave antenna.

### Velocity of propagation and antenna length

In free space, electromagnetic waves travel at a constant velocity of 300,000 km (or approximately 186,000 mi) per second. The RF energy on an antenna, however, moves at a velocity considerably less than that of the radiated energy in free space



5-14 Standing waves measured with a meter.

because the antenna has a *dielectric constant* greater than that of free space. Because the dielectric constant of free space (air or vacuum) is approximately 1, a dielectric constant greater than 1 retards electromagnetic-wave travel.

Because of the difference in velocity between the wave in free space and the wave on the antenna, the *physical* length of the antenna no longer corresponds to its *electrical* length. The antenna is a half-wavelength electrically, but somewhat shorter than this physically. This is shown in the formula for the velocity of electromagnetic waves,

$$V = f\lambda \quad [5.1]$$

where  $V$  is the velocity,  $f$  is the frequency, and  $\lambda$  is the wavelength. Since the frequency of the wave remains constant, a decrease in the velocity results in a decrease in the wavelength. Therefore, the wave traveling in an antenna has a shorter wavelength than the same wave traveling in free space, and the physical length of the antenna can be shorter.

The actual difference between the physical length and the electrical length of the antenna depends on several factors. A thin wire antenna, for example, has less effect on wave velocity than an antenna with a large cross section. As the circumference of the antenna increases, the wave velocity is lowered, as compared with its free-space velocity. The effect of antenna circumference on wave velocity is illustrated in the graph of Fig. 5-15.

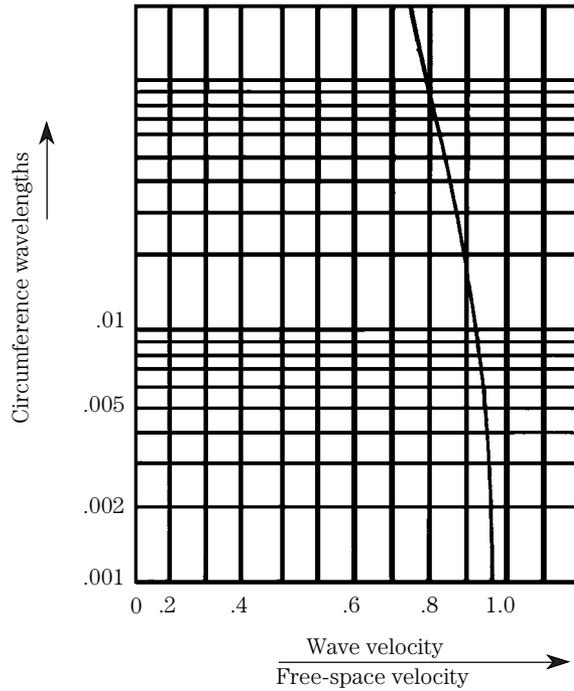
Other factors are involved that lower wave velocity on the antenna. Stray capacitance, for example, increases the dielectric constant and lowers wave velocity. This capacitance can be caused by the line connecting the antenna to the transmitter, the insulators used to give physical support to the antenna, or nearby objects made of metallic or dielectric materials. The change in velocity resulting from stray capacitance is called *end effect* because the ends of the antenna are made farther apart electrically than they are physically. End effect is counteracted by making the physical length about 5 percent shorter than the electrical length, as expressed in the formula

$$\begin{aligned} L &= 0.95 \left( \frac{492}{f} \right) \\ &= \frac{468}{f} \end{aligned} \quad [5.2]$$

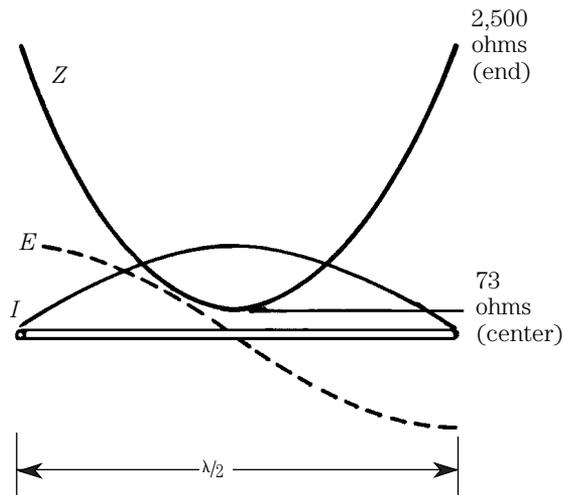
where  $L$  is the physical length in feet and  $f$  is the frequency in megahertz. This formula is accurate for all practical purposes in determining the physical length of a half-wavelength antenna at the operating frequency.

The capacitive end effect also slightly changes the standing waves of voltage and current. When the standing waves are measured, it is found that the nodes have some value and do not reach zero, because some current is necessary to charge the stray capacitance. The standing waves measured in Fig. 5-16 show the results of end effect.

5-15 Effect of antenna circumference on wave velocity.



5-16 Impedance along half-wave antenna.



## Resonance, resistance, and impedance

The antenna is a circuit element having distributed constants of inductance, capacitance, and resistance, which can be made to form a resonant circuit. The half-wave antenna is the shortest resonant length of antenna. However, antennas which are two or more half-wavelengths can also be resonant. Such antennas are said to operate on *harmonics*. If an antenna is four half-wavelengths at the transmitter frequency, it is being operated at the fourth harmonic of its lowest resonant frequency. In other words, this antenna is a half-wavelength at one-quarter of the frequency of operation. An antenna operating on the third harmonic is shown in Fig. 5-7.

**Resistance** A current flowing in the antenna must contend with three kinds of resistance. With the antenna considered as a radiator of energy, the power expended in the form of radiation can be thought of as an  $I^2R_r$  loss.  $R_r$  is called the *radiation resistance*. With the antenna considered as a conductor, a certain amount of energy is dissipated in the form of heat. In this  $I^2R_o$  loss,  $R_o$  is the *ohmic resistance*. There is also an  $I^2R$  loss because of the *leakage resistance* of dielectric elements, such as insulators. This  $R$  usually is included in the ohmic resistance.

The purpose of the antenna is to dissipate as much energy as possible in the form of radiation. The energy dissipated by the radiation resistance, therefore, is the useful part of the total power dissipated. Because the actual power loss depends on the ohmic resistance, this resistance should be kept as low as possible. In the half-wave antenna, the radiation resistance is large compared to the ohmic resistance, and most of the available energy is radiated. The half-wave antenna is, therefore, a very efficient radiator for most purposes.

For a half-wave antenna fed at the center point, the radiation resistance is equal to  $73 \Omega$ . The reference point is the center of the antenna at the time of peak current flow. Ohmic resistance is referred to this point. The total resistance is of importance in matching the antenna to a transmission line.

**Impedance** Because the half-wave antenna has different conditions of voltage and current at different points, and because impedance is equal to the voltage across a circuit divided by the current through it, the impedance will vary along the length of the antenna. If  $E$  is divided by  $I$  at each point of the voltage and current curves in Fig. 5-16, the result is the impedance curve,  $Z$ . The impedance is about  $73 \Omega$  at the center point and rises to a value of about  $2500 \Omega$  at the ends.

The impedance of the half-wave antenna usually is considered to be the impedance as *seen* by the transmitter at the input terminals. This impedance consists of both resistance and reactance. If the antenna is cut to a length of exact resonance, the reactance is zero and the impedance is purely resistive. However, if the antenna is longer or shorter than resonance, reactance is present. When the antenna is made shorter, capacitive reactance is present; when the antenna is made longer, inductive reactance is present.

The impedance at the antenna input terminals is important in terms of power efficiency. If the transmitter is feeding a nonresonant antenna, a power loss is caused by the reactive component of the antenna impedance. Conversely, if the frequency of the transmitter is changed, the electrical length of the antenna also changes. If the frequency is made somewhat higher, the electrical length is made greater, and inductive reactance is added to the impedance. If the frequency is lowered, the electrical length is shortened, and capacitive reactance is added to the impedance.

## 6 CHAPTER

# High-frequency dipole and other doublet antennas

A MYTH AROSE IN RADIO COMMUNICATIONS CIRCLES SOME TIME AGO. PEOPLE CAME TO believe (especially in the ham and CB communities) that large antenna arrays are absolutely necessary for effective communications, especially over long distances. Overlooked, almost to the point of disdain, were effective (but simple) antennas that can be erected by inexperienced people and made to work well. The simple *dipole*, or *doublet*, is a case in point. This antenna is also sometimes called the *Hertz*, or *hertzian*, antenna because radio pioneer Heinrich Hertz reportedly used this form in his experiments.

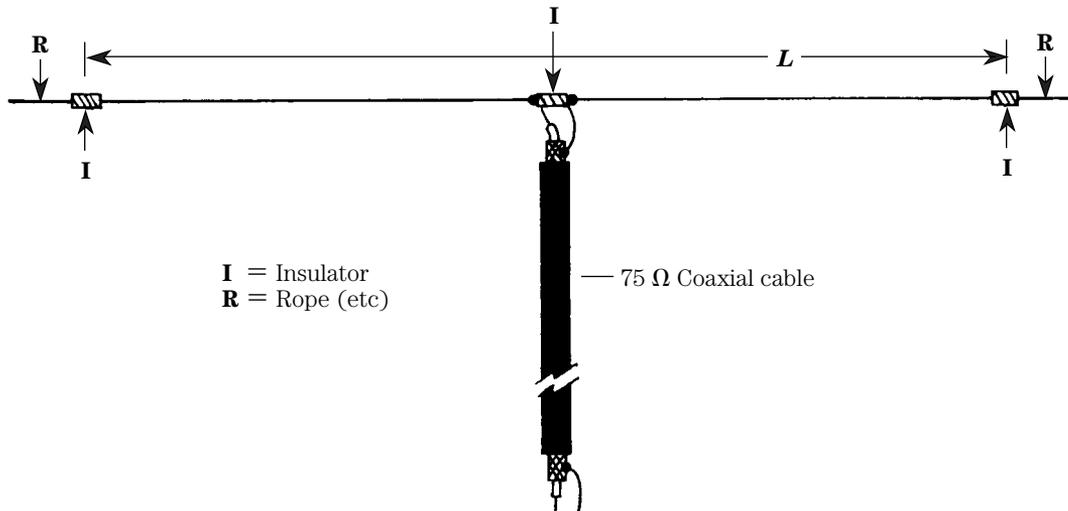
The half-wavelength dipole is a balanced antenna consisting of two radiators (Fig. 6-1) that are each a quarter-wavelength, making a total of a half-wavelength. The antenna is usually installed horizontally with respect to the earth's surface, so it produces a horizontally polarized signal.

In its most common configuration (Fig. 6-1), the dipole is supported at each end by rope and end insulators. The rope supports are tied to trees, buildings, masts, or some combination of such structures.

The length of the antenna is a half-wavelength. Keep in mind that the physical length of the antenna, and the theoretical electrical length, are often different by about 5 percent. A free-space half-wavelength is found from

$$L = \frac{492}{F_{\text{MHz}}} \text{ ft} \quad [6.1]$$

In a perfect antenna, that is self-supported many wavelengths away from any object, Eq. 6.1 will yield the physical length. But in real antennas, the length calculated



6-1 Simple half-wave dipole antenna.

above is too long. The average physical length is shortened by up to about 5 percent because of the velocity factor of the wire and capacitive effects of the end insulators. A more nearly correct *approximation* (remember that word, it's important) of a half-wavelength antenna is

$$L = \frac{468}{F_{\text{MHz}}} \text{ ft} \quad [6.2]$$

where

$L$  is the length of a half-wavelength radiator, in feet  
 $F_{\text{MHz}}$  is the operating frequency, in megahertz

**Example** Calculate the approximate physical length for a half-wavelength dipole operating on a frequency of 7.25 MHz.

**Solution:**

$$\begin{aligned}
 L &= \frac{468}{F_{\text{MHz}}} \text{ ft} \\
 &= \frac{468}{7.25} \text{ ft} = 64.55 \text{ ft}
 \end{aligned}$$

or, restated another way:

$$L = 64 \text{ ft } 6.6 \text{ in}$$

It is unfortunate that a lot of people accept Eq. 6.2 as a universal truth, a kind of immutable law of The Universe. Perhaps abetted by books and articles on antennas that fail to reveal the full story, too many people install dipoles without regard for

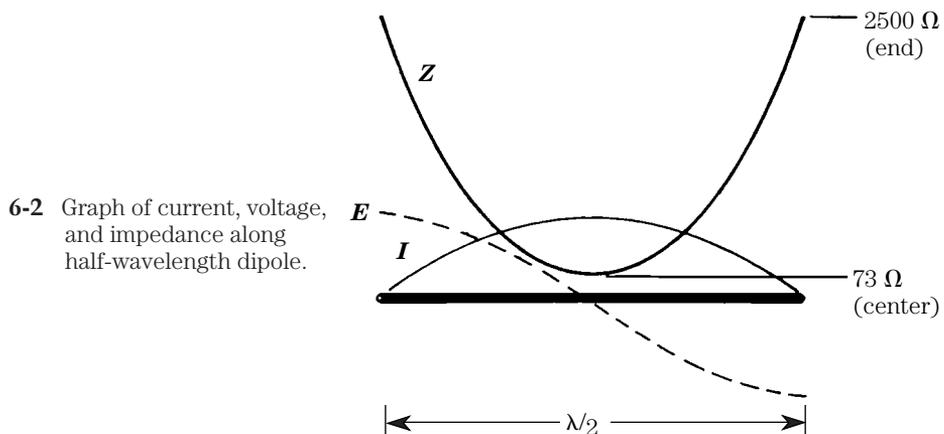
reality. The issue is *resonance*. An antenna is a complex *RLC* network. At some frequency, it will appear like an inductive reactance ( $X = +jX_L$ ), and at others it will appear like a capacitive reactance ( $X = -jX_C$ ). At a specific frequency, the reactances are equal in magnitude, but opposite in sense, so they cancel each other out:  $X_L - X_C = 0$ . At this frequency, the impedance is purely resistive, and the antenna is said to be resonant.

The goal in erecting a dipole is to make the antenna resonant at a frequency that is inside the band of interest, and preferably in the portion of the band most often used by the particular station. Some of the implications of this goal are covered later on, but for the present, assume that the builder will have to custom-tailor the length of the antenna. Depending on several local factors (among them, nearby objects, the shape of the antenna conductor, and the length/diameter ratio of the conductor) it might prove necessary to add, or trim, the length a small amount to reach resonance.

## The dipole feedpoint

The dipole is a half-wavelength antenna fed in the center. Figure 6-2 shows the voltage ( $V$ ) and current ( $I$ ) distributions along the length of the half-wavelength radiator element. The feedpoint is at a voltage minimum and a current maximum, so you can assume that the feedpoint is a current antinode.

At resonance, the impedance of the feedpoint is  $R_o = V/I$ . There are two resistances that make up  $R_o$ . The first is the ohmic losses that generate nothing but heat when the transmitter is turned on. These ohmic losses come from the fact that conductors have electrical resistance and electrical connections are not perfect (even when properly soldered). Fortunately, in a well-made dipole these losses are almost negligible. The second contributor is the *radiation resistance*  $R_r$  of the antenna. This resistance is a hypothetical concept that accounts for the fact that RF power is radiated by the antenna. The radiation resistance is the fictional resistance that would dissipate the amount of power that is radiated away from the antenna.



For example, suppose we have a large-diameter conductor used as an antenna, and it has negligible ohmic losses. If 1000 W of RF power is applied to the feedpoint, and a current of 3.7 A is measured, what is the radiation resistance?

$$R_r = \frac{P}{I^2} \quad [6.3]$$

$$= \frac{1000 \text{ W}}{(3.7)^2} \quad [6.4]$$

$$= 73 \Omega \quad [6.5]$$

It is always important to match the feedpoint impedance of an antenna to the transmission-line impedance. Maximum power transfer always occurs (in any system) when the source and load impedances are matched. In addition, if some applied power is not absorbed by the antenna (as happens in a mismatched system), then the unabsorbed portion is reflected back down the transmission line toward the transmitter. This fact gives rise to standing waves, and the so-called standing wave ratio (SWR or VSWR) discussed in Chap. 3. This is a problem to overcome.

Matching antenna feedpoint impedance seems to be simplicity itself because the free-space feedpoint impedance of a simple dipole is about  $73 \Omega$ , seemingly a good match to  $75\text{-}\Omega$  coaxial cable. Unfortunately, the  $73\text{-}\Omega$  feedpoint impedance is almost a myth. Figure 6-3 shows a plot of approximate radiation resistance ( $R_r$ ) versus height above ground (as measured in wavelengths). As before, we deal in approximations in Fig. 6-3; in this case, the ambiguity is introduced by ground losses.

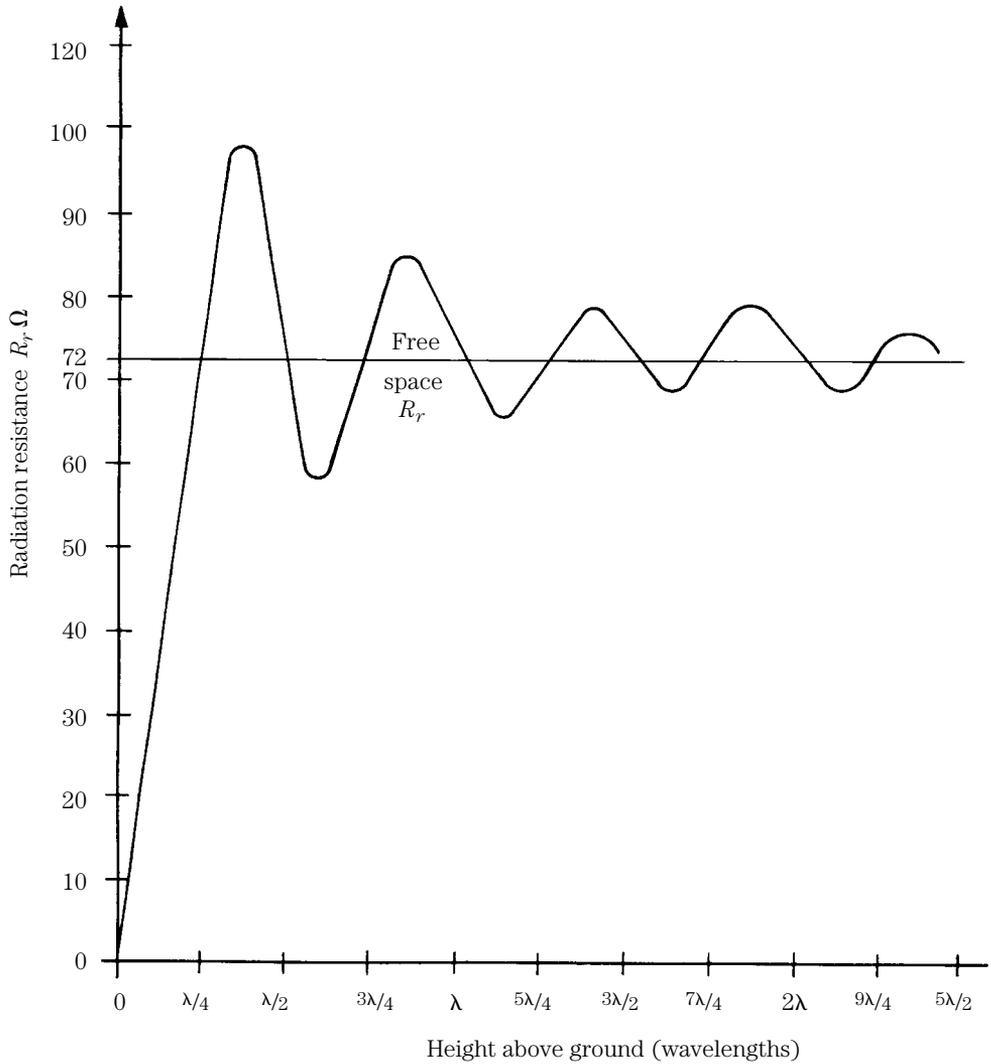
Despite the fact that Fig. 6-3 is based on approximations, you can see that radiation resistance varies from less than  $10 \Omega$ , to around  $100 \Omega$ , as a function of height. At heights of many wavelengths, this oscillation of the curve settles down to the free-space impedance ( $72 \Omega$ ). At the higher frequencies, it might be possible to install a dipole at a height of many wavelengths. In the 2-m amateur radio band (144 to 148 MHz), one wavelength is around 6.5 ft (i.e.,  $2 \text{ m} \times 3.28 \text{ ft/m}$ ), so “many wavelengths” is relatively easy to achieve at reasonably attainable heights. In the 80-m band (3.5 to 4.0 MHz), however, one wavelength is on the order of 262 ft, so “many wavelengths” is a practical impossibility.

There are three tactics that can be followed. First, ignore the problem altogether. In many installations, the height above ground will be such that the radiation resistance will be close enough to present only a slight impedance mismatch to a standard coaxial cable. The VSWR is calculated (among other ways) as the ratio:

$$1. Z_o > R_r:$$

$$\text{VSWR} = \frac{Z_o}{R_r} \quad [6.6]$$

$$2. Z_o < R_r:$$



6-3 Radiation resistance versus height above ground.

$$\text{VSWR} = \frac{R_r}{Z_o} \tag{6.7}$$

where

$Z_o$  is the coaxial-cable characteristic impedance

$R_r$  is the radiation resistance of the antenna

Consider an antenna mounted at a height somewhat less than a quarter-wave-length, such that the radiation resistance is 60  $\Omega$ . Although not recommended as

good engineering practice (there are sometimes practical reasons) it is nonetheless necessary to install a dipole at less than optimum height. So, if that becomes necessary, what are the implications of feeding a 60- $\Omega$  antenna with either 52- or 75- $\Omega$  standard coaxial cable? Some calculations are revealing:

For 75- $\Omega$  coaxial cable:

$$\text{VSWR} = \frac{Z_o}{R_r} \quad [6.8]$$

$$= \frac{75 \Omega}{60 \Omega} = 1.25:1 \quad [6.9]$$

For 52- $\Omega$  coaxial cable:

$$\text{VSWR} = \frac{R_r}{Z_o} \quad [6.10]$$

$$= \frac{60 \Omega}{52 \Omega} = 1.15:1 \quad [6.11]$$

In neither case is the VSWR created by the mismatch too terribly upsetting.

The second approach is to mount the antenna at a convenient height, and use an impedance matching scheme to reduce the VSWR. In Chap. 23, you will find information on various suitable (relatively) broadbanded impedance matching methods including Q-sections, coaxial impedance transformers, and broadband RF transformers. “Homebrew” and commercially available transformers are available to cover most impedance transformation tasks.

The third approach is to mount the antenna at a height (Fig. 6-3) at which the expected radiation resistance crosses a standard coaxial cable characteristic impedance. The best candidate seems to be a height of a half-wavelength because the radiation resistance is close to the free-space value of 72  $\Omega$ , and is thus a good match for 75- $\Omega$  coaxial cable (such as RG-11/U or RG-59/U).

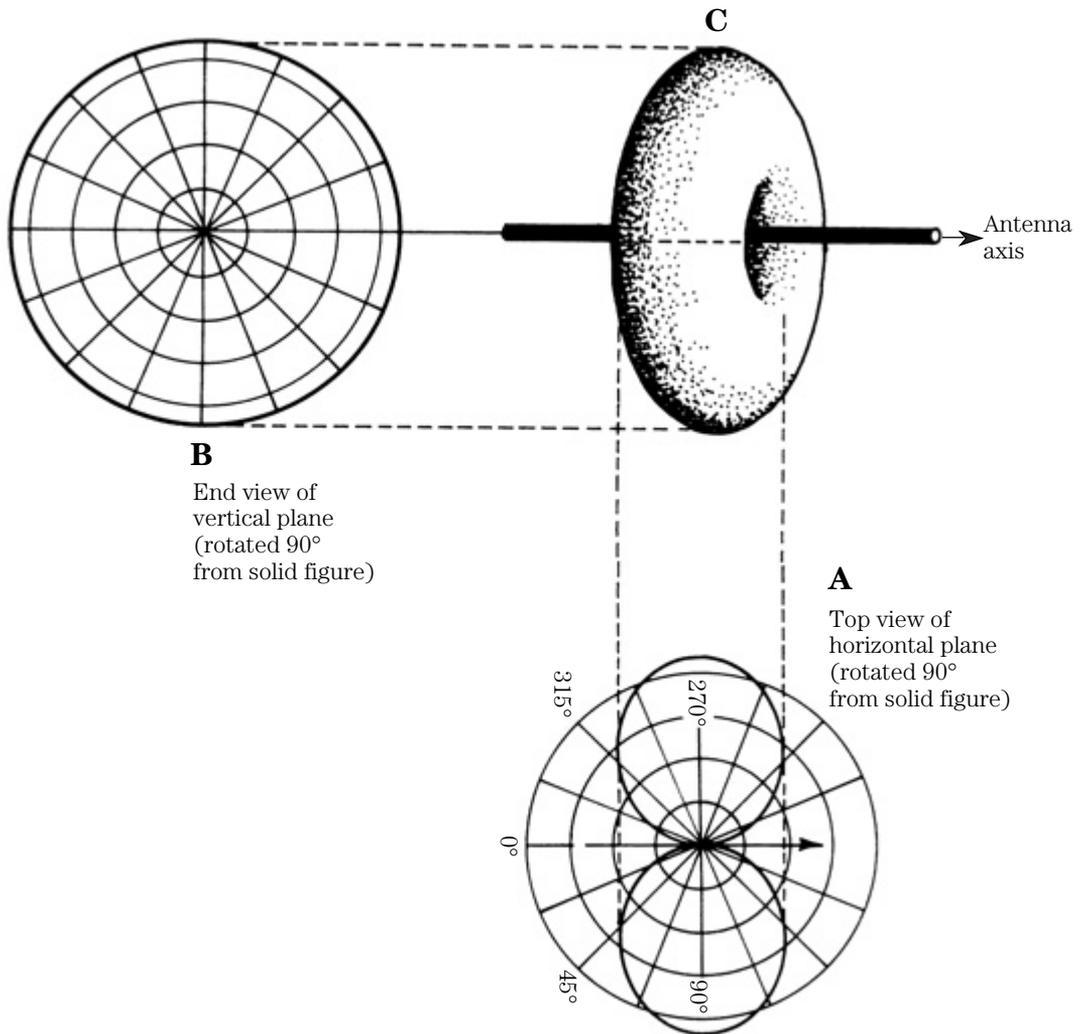
## The dipole radiation pattern

The radiation patterns of various antennas are covered in this book. Some basic theory of patterns is repeated at each section, not to fill space, but rather to drive home a point and refresh reader memories. We keep harking back to the concepts *directivity* and *gain* (which are actually different expressions of the same thing).

Antenna theory recognizes a point of reference called the *isotropic radiator*. This device is a theoretical construct consisting of a spherical point source of RF radiation in all directions (see Chap. 2). It is truly *omnidirectional* (“all directions”) because it creates an ever-expanding sphere as the RF wavefront propagates outward. Antenna gain is a measure of how the antenna focuses available power away from a spherical wavefront to a limited number of directions (two, in the case of the dipole). Thus are related the concepts *directivity* and *gain*.

Always keep in mind that *directivity and gain are specified in three dimensions*. Too many times, people simplify the topic too much by publishing only part of the radiation pattern (azimuth aspect). In other words, the reader is given a pattern viewed from above that shows the directivity in the horizontal plane. But a signal does not propagate away from an antenna in an infinitely thin sheet, as such presentations seem to imply; rather, it has an elevation extent in addition to the azimuth extent. Thus, proper evaluation of an antenna takes into consideration both horizontal and vertical plane patterns.

Figure 6-4 shows the radiation pattern of a dipole antenna in free space “in the round.” In the horizontal plane (6-4A), when viewed from above, the pattern is a “figure



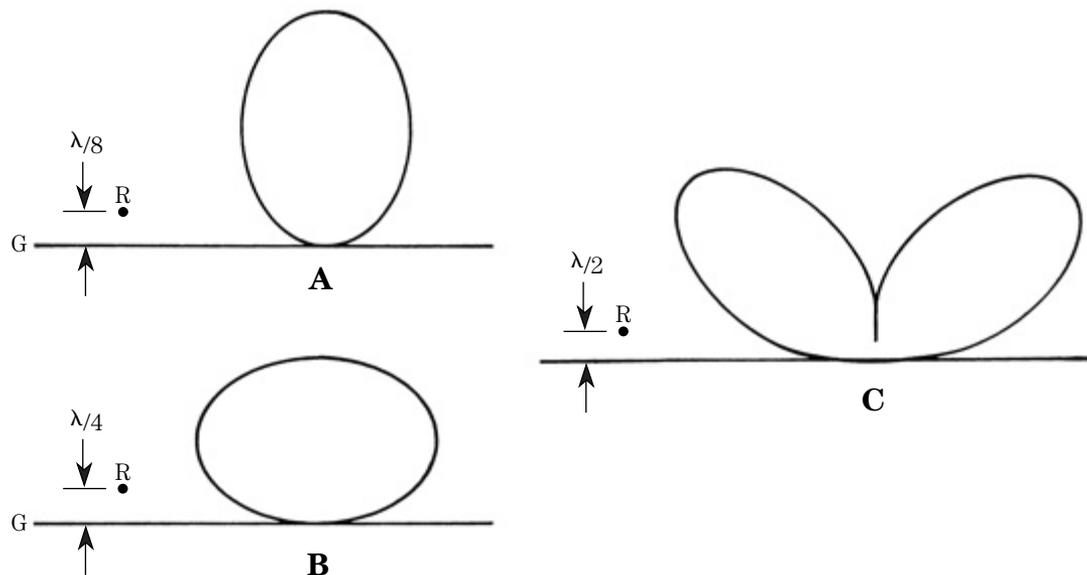
6-4 Idealized dipole radiation pattern.

8” that exhibits bidirectional radiation. Two main “lobes” contain the RF power from the transmitter, with sharp nulls of little or no power off the ends of the antenna axis. This pattern is the classical dipole pattern that is published in most antenna books.

Also shown, however, is the vertical plane pattern for a dipole antenna in free space. Note that when sliced in this aspect the radiation pattern is circular (Fig. 6-4B). When the two patterns are combined in the round, you can see the three-dimensional doughnut-shaped pattern (Fig. 6-4C) that most nearly approximates the true pattern of an unobstructed dipole in free space.

When a dipole antenna is installed close to the earth’s surface, not in free space, as is the case at most stations, the pattern is distorted from that of Fig. 6-4. Two effects must be taken into consideration. First, and most important, is the fact that the signal from the antenna is reflected from the surface and bounces back into space. This signal will be phase-shifted both by the reflection and by the time required for the transit to occur. At points where the reflected wave combines in phase with the radiated signal, the signal is reinforced; in places where it combines out of phase, the signal is attenuated. Thus, the reflection of the signal from the ground alters the pattern from the antenna. The second factor is that the ground is lossy, so not all of the signal is reflected; some of it heats the ground underneath the antenna. Thus, the signal is attenuated at a greater rate than the inverse square law, so it further alters the expected pattern.

Figure 6-5 shows patterns typical of dipole antennas installed close to the earth’s surface. The views in this illustration correspond to Fig. 6-4B, in that they are looking at the vertical plane from a line along the antenna axis. Therefore, the antenna is represented by *R* in each case shown. Figure 6-5A shows the case for a dipole



**6-5** Vertical extent of dipole antenna at (A)  $\frac{1}{8}$  wavelength, (B)  $\frac{1}{4}$  wavelength, and (C)  $\frac{1}{2}$  wavelength above ground.

installed at  $1/8$  wavelength above the surface. For this antenna, most of the RF energy is radiated almost straight up (not very useful). This type of antenna is basically limited to ground wave and very short skip (when available). The second case (Fig. 6-5B) shows the pattern when the antenna is a quarter wavelength above the earth's surface. Here the pattern is flattened, but it still shows considerable energy in the vertical direction (where it is useless). Finally, you can see the pattern obtained when the antenna is installed a half-wavelength above the surface. In this case, the pattern is best for long-distance work because energy is redirected away from straight-up into lobes at relatively shallow angles.

### Tuning the dipole antenna

There are two issues to address when tuning an antenna (any antenna, not just the dipole): *resonance* and *impedance matching*. Although frequently treated in the literature as the same issue, they are not. This section deals mostly with the process of tuning the antenna to resonance. Although not all forms of antenna are resonant, the dipole *is* an example of a resonant antenna.

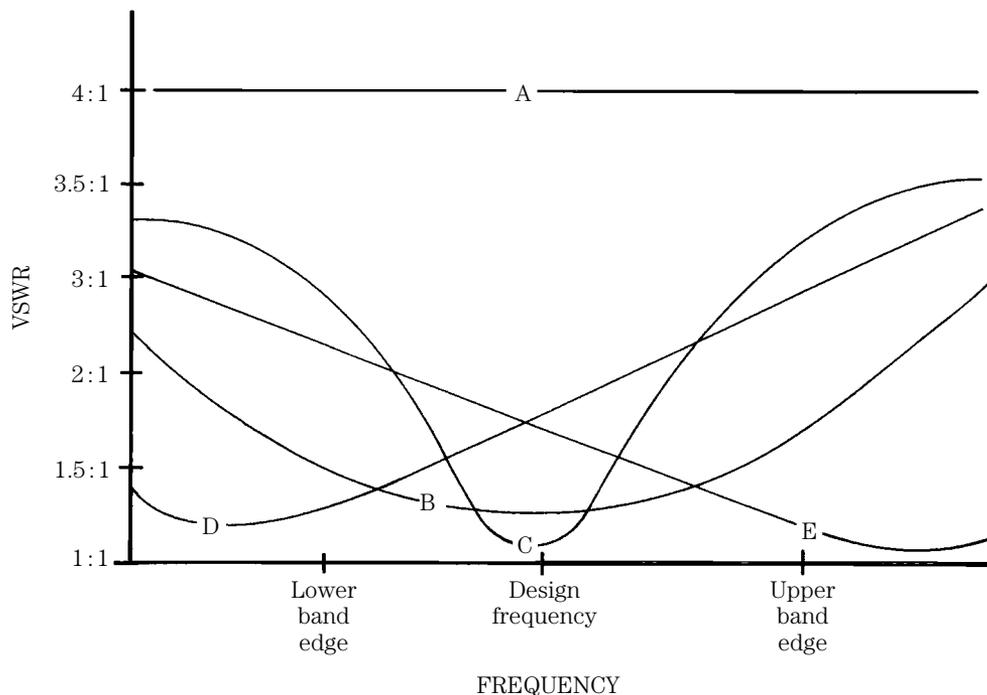
There is a lot of misinformation abroad concerning the tuning of antennas. Perhaps much of what is believed comes from the fact that VSWR is used as the indicator of both impedance matching and resonance. Quite a few people honestly, but erroneously, believe that the VSWR can be “tuned out” by adjusting the length of the feedline. That myth probably derives from the fact that voltage or current sensing instruments are used for VSWR measurement, and these are affected by transmission line length. But that fact is caused by a weakness in the instruments, not by radio physics.

There is only one proper way to tune a dipole antenna: *adjust the length of the antenna elements*, not the transmission line. It was in order to make these adjustments that we purposely did not initially tell you to solder the electrical connections at the center insulator.

**Resonance** The indicator of resonance is the minimum point in the VSWR curve. Figure 6-6 shows a graph of VSWR vs. frequency for several different cases. Curve *A* represents a disaster: a high VSWR all across the band. The actual value of VSWR can be anything from about 3.5:1 to 10:1, or thereabouts, but the cause is nonetheless the same: the antenna is either open or shorted; or it is so far off resonance as to appear to be open or shorted to the VSWR meter.

Curves *B* and *C* represent antennas that are resonant within the band of interest. Curve *B* represents a broadbanded antenna that is relatively flat all across the band and does not exhibit excessive VSWR until the frequency is outside of the band. Curve *C* is also resonant within the band, but this antenna has a much higher  $Q$  than curve *B*. In the naive sense, the broadbanded antenna is best, but that statement is true only if the broadness is not purchased at the expense of efficiency. Losses tend to broaden the antenna, but also reduce its effectiveness. So, if broad bandedness is purchased at the risk of increased loss, then it is less than desirable.

Curves *D* and *E* are resonant outside the band of interest. The curve marked *D* is resonant at a frequency on the low side of the band, so the dipole is too long. In this case, you need to shorten the antenna a bit to raise the resonant point inside



6-6 VSWR versus frequency for several cases.

the band. Curve *E* represents an antenna that is resonant outside the upper limit of the band, so this antenna is too short, and must be lengthened. Because this situation is possible, the antenna elements are typically made longer than needed when they are first cut.

How much to cut? That depends on two factors: how far from the desired frequency the resonant point is found, and which band is being used. The latter requirement comes from the fact that the “frequency per unit length” varies from one band to another. Look at an example of how to calculate this figure. The procedure is simple:

1. Calculate the length required for the upper end of the band.
2. Calculate the length required for the bottom end of the band.
3. Calculate the difference in lengths for the upper and lower ends of the band.
4. Calculate the width of the band in kilohertz by taking the difference between the upper frequency limit and the lower frequency limit.
5. Divide the length difference by the frequency difference; the result is in kilohertz per unit length.

**Example 6-1** Calculate the frequency change per unit of length for 80 and for 15 m.

**Solution** For 80 m (3.5 to 4.0 MHz):

1.  $L_{ft} = \frac{468}{4 \text{ MHz}} = 117 \text{ ft}$
2.  $L_{ft} = \frac{468}{3.5 \text{ MHz}} = 133.7 \text{ ft}$
3. Difference in length:  $133.7 \text{ ft} - 117 \text{ ft} = 16.7 \text{ ft}$
4. Frequency difference:  $4000 \text{ kHz} - 3500 \text{ kHz} = 500 \text{ kHz}$
5. Calculate  $\frac{\text{frequency}}{\text{unit length}} : \frac{500 \text{ kHz}}{16.7 \text{ ft}} = 30 \text{ kHz/ft}$

For 15 m (21.0 – 21.45 MHz):

1.  $L_{ft} = \frac{468}{21.45} = 21.82 \text{ ft}$
2.  $L_{ft} = \frac{468}{21} = 22.29 \text{ ft}$
3. Difference in length:  $22.29 \text{ ft} - 21.82 \text{ ft} = 0.47 \text{ ft}$
4. Convert to inches:  $0.47 \text{ ft} \times 12 \text{ in/ft} = 5.64 \text{ in}$
5. Frequency difference:  $21,450 \text{ kHz} - 21,000 \text{ kHz} = 450 \text{ kHz}$
6. Calculate  $\frac{\text{frequency}}{\text{unit length}} : \frac{450 \text{ kHz}}{5.64 \text{ in}} = 80 \text{ kHz/in}$

At 80 m, the frequency change per foot is small, but at 15 m small changes can result in very large frequency shifts. You can calculate approximately how much to add (or subtract) from an antenna under construction from this kind of calculation. If, for example, you design an antenna for the so-called “net frequency” on 15 m (21,390 kHz), and find the actual resonant point is 21,150 kHz, the frequency shift required is  $21,390 - 21,150$ , or 240 kHz. To determine how much to add or subtract (as a first guess):

1. The factor for 15 m is 80 kHz/in, which is the same as saying 1 in/80 kHz.
2. The required frequency shift is 240 kHz.
3. Therefore:

$$\text{Length change} = 240 \text{ kHz} \times \frac{1 \text{ in}}{80 \text{ kHz}} \quad \text{[6.12]}$$

$$= 3 \text{ in} \quad \text{[6.13]}$$

Each side of the antenna must be changed by half of the length calculated above, or 1.5 in. Because the first resonant frequency is less than the desired frequency, the length should be shortened 1.5 in. Once the length is correct, as proven by the VSWR curve, the connections at the center insulator are soldered and made permanent, and the antenna rehoisted to the operating level.

**Impedance matching** The difference between resonance and impedance matching is seen in the value of the VSWR minimum. While the minimum indicates the resonant point, its value is a measure of the relationship between the feedpoint

impedance of the antenna and the characteristic impedance of the transmission line. Earlier in this chapter you learned that

$$1. Z_o > R_r:$$

$$\text{VSWR} = \frac{Z_o}{R_r} \quad [6.14]$$

$$2. Z_o < R_r:$$

$$\text{VSWR} = \frac{R_r}{Z_o} \quad [6.15]$$

where

$Z_o$  is the coaxial-cable characteristic impedance

$R_r$  is the radiation resistance of the antenna

Although knowledge of the VSWR will not show which situation is true, you can know that there is a high probability that *one* of them is true, and you can experiment to find which is the case. Of course, if the VSWR is less than about 1.5:1 or 2:1, then forget about it—the improvement is not generally worth the expense and cost. When coupled to a transmitter that is equipped with the tunable output network (most tube-type transmitters or final amplifiers), then it can accommodate a relatively wide range of reflected antenna impedances. But modern solid-state final amplifiers tend to be a little more picky about the load impedance. For these transmitters, a coax-to-coax *antenna tuning unit* (ATU) is needed.

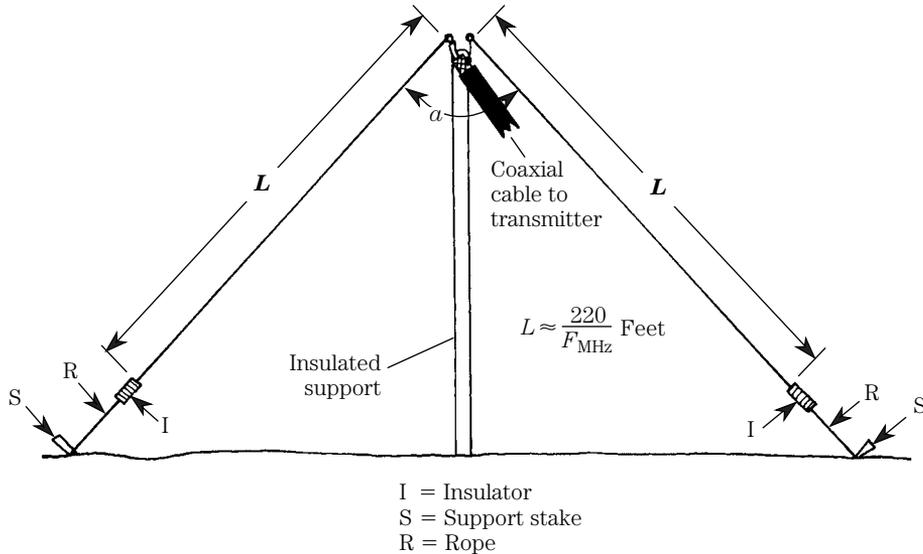
## Other dipoles

Thus far, the dipoles covered in this chapter have been the classic form, in which a half-wavelength single-conductor radiator element is connected to a coaxial transmission line. This antenna is typically installed horizontally at a half-wavelength above the earth's surface (or wherever convenient if that is impossible). This section looks at other forms of the dipole. Some of these dipoles are in every way the equal of the horizontal dipole, and others are basically compensation antennas in that they are used when a proper dipole is not practical.

### Inverted-vee dipole

The *inverted-vee dipole* is a half-wavelength antenna fed in the center like a dipole. By the rigorous definition, the inverted-vee is merely a variation on the dipole theme. But in this form of antenna (Fig. 6-7), the center is elevated as high as possible from the earth's surface, but the ends droop to very close to the surface. Angle  $a$  can be almost anything convenient, provided that  $a > 90$  degrees; typically, most inverted-vee antennas use an angle of about 120 degrees. Although essentially a compensation antenna for use when the dipole is not practical, many operators believe that it is essentially a better performer on 40 and 80 m in cases where the dipole cannot be mounted at a half-wavelength (64 ft or so).

By sloping the antenna elements down from the horizontal to an angle (as shown in Fig. 6-7), the resonant frequency is effectively lowered. Thus, the antenna will



6-7 Inverted-vee dipole.

need to be shorter for any given frequency than a dipole. There is no absolutely rigorous equation for calculation of the overall length of the antenna elements. Although the concept of “absolute” length does not hold for regular dipoles, it is even less viable for the inverted-vee. There is, however, a rule of thumb that can be followed for a starting point: Make the antenna about 6 percent shorter than a dipole for the same frequency. The initial cut of the antenna element lengths (each quarter wavelength) is

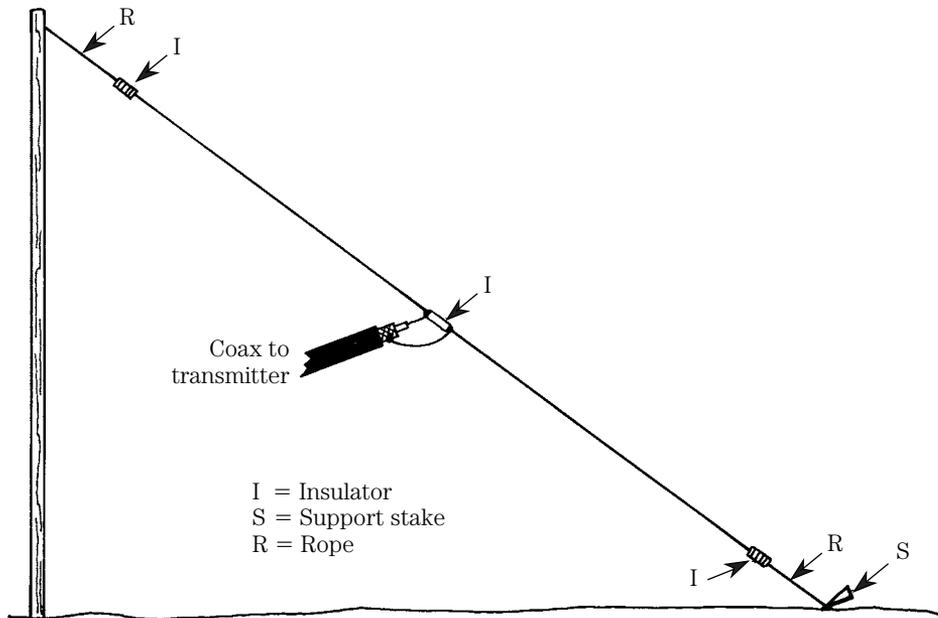
$$L = \frac{220}{F_{\text{MHz}}} \text{ ft} \quad [6.16]$$

After this length is determined, the actual length is found from the same cut-and-try method used to tune the dipole in the previous section.

Bending the elements downward also changes the feedpoint impedance of the antenna and narrows its bandwidth. Thus, some adjustment in these departments is in order. You might want to use an impedance matching scheme at the feedpoint, or an antenna tuner at the transmitter.

### Sloping dipole (“sloper” or “slipole”)

The *sloping dipole* (Fig. 6-8) is popular with those operators who need a low angle of radiation, and are not overburdened with a large amount of land to install the antenna. This antenna is also called the *sloper* and the *slipole* in various texts. The author prefers the term “slipole,” in order to distinguish this antenna from a sloping vertical of the same name. Whatever it is called, however, it is a half-wavelength dipole that is built with one end at the top of a support, and the other end close to the



6-8 "Slipole" or "sloper."

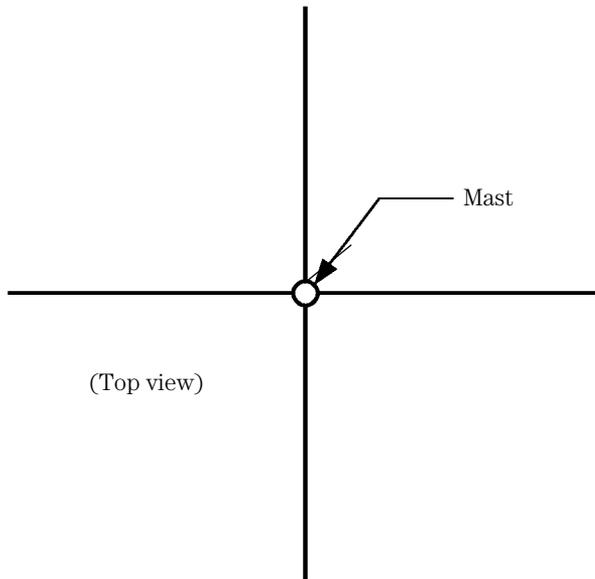
ground, and being fed in the center by coaxial cable. Some of the same comments as obtained for the inverted-vee antenna also apply to the sloping dipole, so please see that section also.

Some operators like to arrange four sloping dipoles from the same mast such that they point in different directions around the compass (Fig. 6-9). A single four-position coaxial cable switch will allow switching a directional beam around the compass to favor various places in the world.

### Broadbanded dipoles

One of the rarely discussed aspects of antenna construction is that the length/diameter ratio of the conductor used for the antenna element is a factor in determining the bandwidth of the antenna. In general, the rule of thumb states that large cross-sectional area makes the antenna more broadbanded. In some cases, this rule suggests the use of aluminum tubing instead of copper wire for the antenna radiator. On the higher-frequency bands that is a viable solution. Aluminum tubing can be purchased for relatively small amounts of money, and is both lightweight and easily worked with ordinary tools. But, as the frequency decreases, the weight becomes greater because the tubing is both longer and (for structural strength) must be of greater diameter. On 80 m, aluminum tubing is impractical, and at 40 m it is nearly so. Yet, 80 m is a significant problem, especially for older transmitters, because the band is 500 kHz wide, and the transmitters often lack the tuning range for the entire band.

Some other solution is needed. Here are three basic solutions to the problem of wide-bandwidth dipole antennas: *folded dipole*, *boutie dipole*, and *cage dipole*.



6-9 Directional antenna made of four slopers (top view).

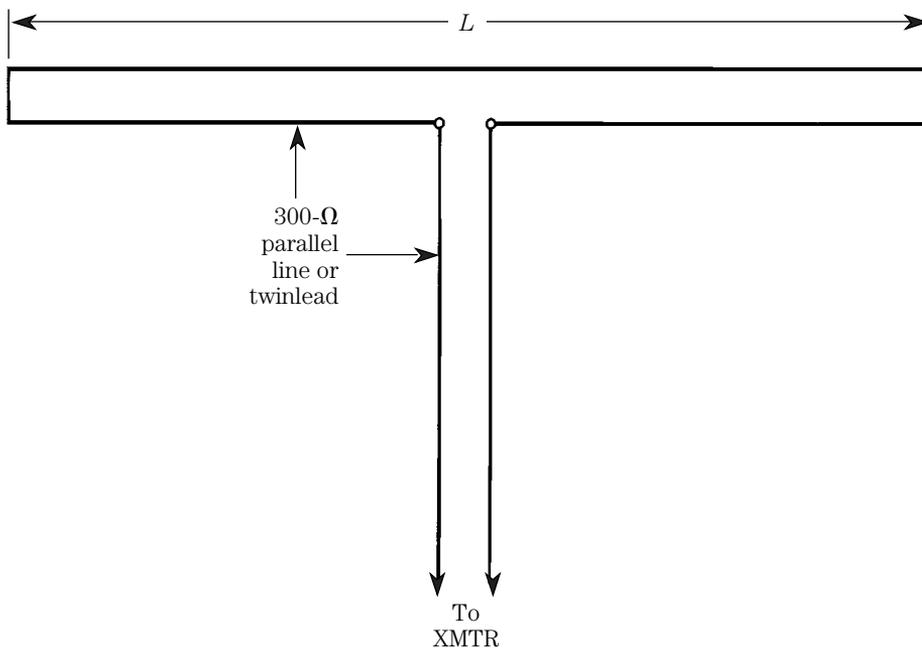
Figure 6-10A shows the *folded dipole* antenna. This antenna basically consists of two half-wavelength conductors shorted together at the ends and fed in the middle of one of them. The folded dipole is most often built from 300- $\Omega$  television antenna twin-lead transmission line. Because the feedpoint impedance is nearly 300  $\Omega$ , the same type of twin lead can also be used for the transmission line. The folded dipole will exhibit excellent wide-bandwidth properties, especially on the lower bands.

A disadvantage of this form of antenna is that the transmitter has to match the 300- $\Omega$  balanced transmission line. Unfortunately, most modern radio transmitters are designed to feed coaxial-cable transmission line. Although an antenna tuner can be placed at the transmitter end of the feedline, it is also possible to use a 4:1 balun transformer at the feedpoint (Fig. 6-10B). This arrangement makes the folded dipole a reasonable match to 52- or 75- $\Omega$  coaxial-cable transmission line.

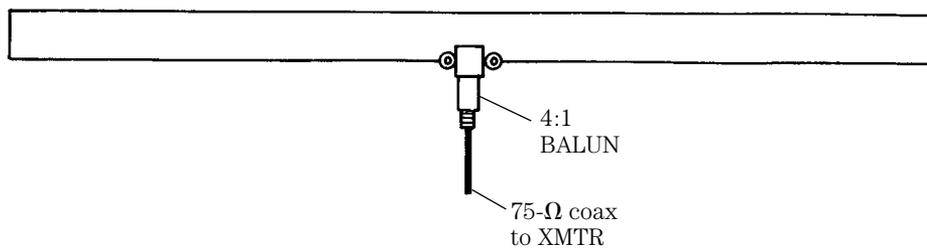
Another method for broadbanding the dipole is to use two identical dipoles fed from the same transmission line, and arranged to form a “bowtie” as shown in Fig. 6-11. The use of two identical dipole elements on each side of the transmission line has the effect of increasing the conductor cross sectional area so that the antenna has a slightly improved length/diameter ratio.

The *bowtie dipole* was popular in the 1930s and 1940s, and became the basis for the earliest television receiver antennas (TV signals are 3 to 5 MHz wide, so they require a broadbanded antenna). It was also popular during the 1950s as the so-called Wonder Bar antenna for 10 m. It still finds use, but it has faded somewhat in popularity.

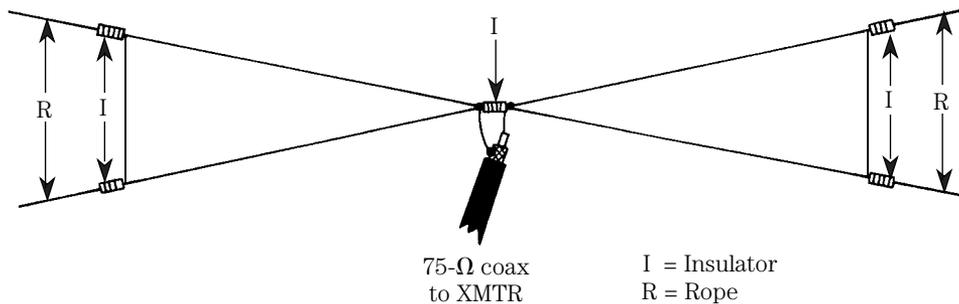
The *cage diople* (Fig. 6-12) is similar in concept, if not construction, to the bowtie. Again, the idea is to connect several parallel dipoles together from the same transmission line in an effort to increase the apparent cross-sectional area. In the case of the cage dipole, however, spreader disk insulators are constructed to keep the wires separated. The insulators can be built from plexiglass, lucite, or ceramic.



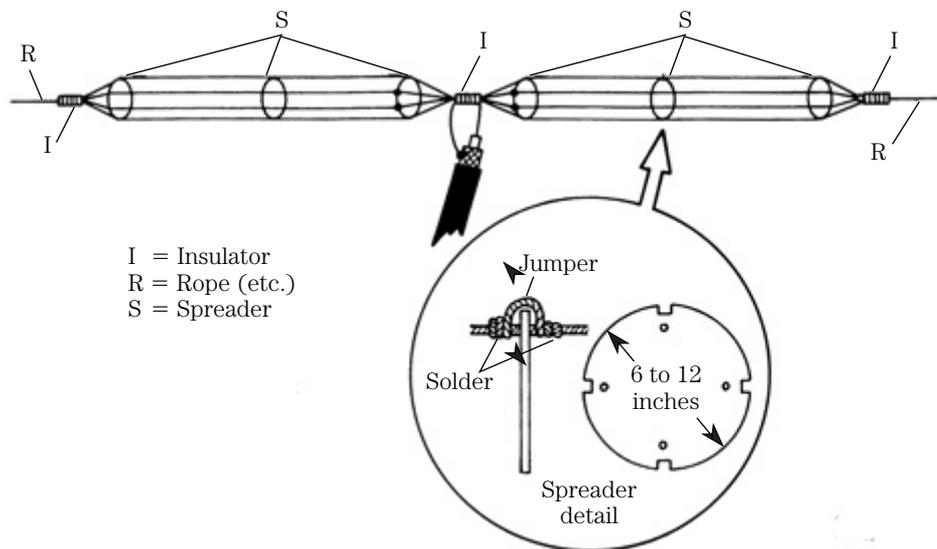
**6-10A** Folded dipole fed with 300- $\Omega$  line.



**6-10B** Folded dipole fed with coaxial cable.



**6-11** Bowtie dipole.



6-12 Cage dipole.

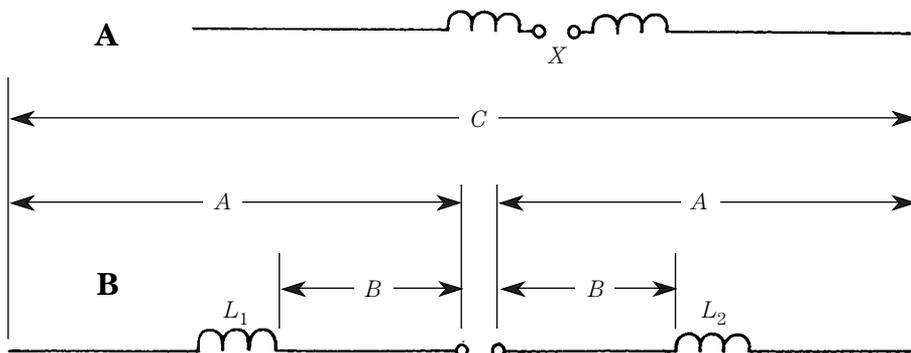
They can also be constructed of materials such as wood, if the wood is properly treated with varnish, polyurethane, or some other material that prevents it from becoming waterlogged. The spreader disks are held in place with wire jumpers (see inset to Fig. 6-12) that are soldered to the main element wires.

A tactic used by some builders of both bowtie and cage dipoles is to make the elements *slightly* different lengths. This “stagger tuning” method forces one dipole to favor the upper end of the band, and the other to favor the lower end of the band. The overall result is a slightly flatter frequency response characteristic across the entire band. On the cage dipole, with four half-wavelength elements, it should be possible to overlap even narrower sections of the band in order to create an even flatter characteristic.

### Shortened coil-loaded dipoles

The half-wavelength dipole is too long for some applications where real estate is at a premium. The solution for many operators is to use a coil-loaded shortened dipole such as shown in Fig. 6-13. A shortened dipole (i.e., one which is less than a half-wavelength) is capacitive reactance. There is no reason why the loading coil cannot be any point along the radiator, but in Figs. 6-13A and 6-13B they are placed at 0 percent and 50 percent of the element length, respectively. The reason for this procedure is that it makes the calculation of coil inductances easier, and it also represents the most common practice.

Figure 6-13C shows a table of inductive reactances as a function of the percentage of a half-wavelength, represented by the shortened radiator. It is likely that the percentage figure will be imposed on you by the situation, but the general rule is to pick the largest figure consistent with the available space. For example, sup-



**A**

$$L_1 = L_2 = L$$

$$C = 2A$$

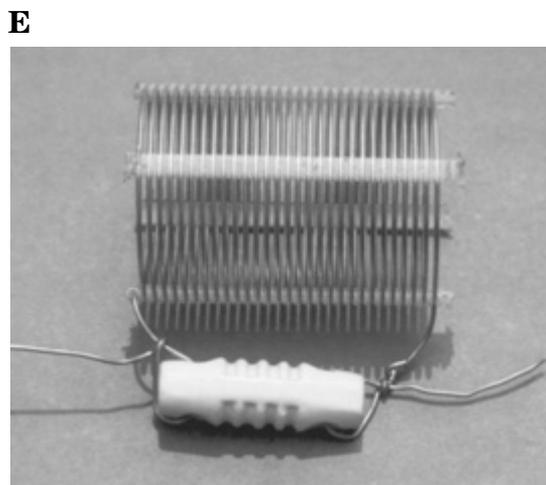
$$A = 1/2C$$

**B**

$$C = \frac{468M}{F_{\text{MHz}}}$$

$$0 \leq M \leq 1$$

Percent of half-wavelength	Coils at feedpoint ( $\Omega$ )	Coils at middle of radiators ( $\Omega$ )
20	1800	2800
30	950	1800
40	700	1800
50	500	1300
60	360	950
70	260	700
80	160	500
90	75	160
95	38	80
98	15	30



**6-13** Shortened or loaded dipole. (A) inductors at feedpoint; (B) inductors midway along elements; (C) chart of reactances for coils; (D) commercially available coils; (E) homemade coil based on commercial coil stock.

pose you have about 40 ft available for a 40-m antenna that normally needs about 65 ft for a half-wavelength. Because 39 ft is 60 percent of 65 ft, you could use this value as the design point for this antenna. Looking on the chart, a 60 percent antenna with the loading coils at the midpoint of each radiator element wants to see an inductive reactance of  $700 \Omega$ . You can rearrange the standard inductive reactance equation ( $X_L = 6.28 FL$ ) to the form

$$L_{\mu\text{H}} = \frac{X_L \times 10^6}{6.28F} \quad [6.17]$$

where

$L_{\mu\text{H}}$  is the required inductance, in microhenrys

$F$  is the frequency, in hertz (Hz)

$X_L$  is the inductive reactance calculated from the table in Fig. 6-13C.

**Example 6-2** Calculate the inductance required for a 60 percent antenna operating on 7.25 MHz. The table requires a reactance of  $700 \Omega$  for a loaded dipole with the coils in the center of each element (Fig. 6-15B).

**Solution:**

$$L_{\mu\text{H}} = \frac{X_L \times 10^6}{6.28F} \quad [6.18]$$

$$= \frac{(700)(10^6)}{(6.28)(7,250,000)} \quad [6.19]$$

$$= \frac{7 \times 10^8}{4.6 \times 10^7} \quad [6.20]$$

$$= 15.4 \mu\text{H} \quad [6.21]$$

The inductance calculated above is approximate, and it might have to be altered by cut-and-try methods.

The loaded dipole antenna is a very sharply tuned antenna. Because of this fact, you must either confine operation to one segment of the band, or provide an antenna tuner to compensate for the sharpness of the bandwidth characteristic. However, efficiency drops, markedly, far from resonance even with a transmission line tuner. The function of the tuner is to overcome the bad effects on the transmitter, but it does not alter the basic problem. Only a variable inductor in the antenna will do that trick (at least one commercial loaded dipole once used a motor-driven inductor at the center feedpoint).

Figures 6-13D and E show two methods for making a coil-loaded dipole antenna. Figure 6-13D shows a pair of commercially available loading coils especially designed for this purpose. The ones shown here are for 40 m, but other models are also available. The inductor shown in Fig. 6-13E is a section of commercial coil stock connected to a standard end or center insulator. No structural stress is assumed by the coil—all forces are applied to the insulator, which is designed to take it.

Inductance values for other length antennas can be approximated from the graph in Fig. 6-14. This graph contains three curves for coil-loaded, shortened dipoles that are 10, 50, and 90 percent of the normal half-wavelength size. Find the proposed location of the coil, as a percentage of the wire element length, along the horizontal axis. Where the vertical line from that point intersects with one of the three curves, that intersection yields the inductive reactance required (see along vertical axis). Inductances for other overall lengths can be “rough-guessed” by interpolating between the three available curves, and then validated by cut and try.

### Tunable dipoles

Dipoles are resonant antennas, so they naturally tend to prefer one frequency over the others. The VSWR will be quite low at the resonant point (assuming no feedline mismatch problems), and will rise at frequencies above and below resonance. If the antenna is a *high-Q* model, then the effect is quite profound, and it might render the antenna nearly useless at frequencies on the other end of the same band (especially where modern transmitters equipped with VSWR shutdown circuitry are used). Figure 6-15 shows a method for overcoming this problem. Here is how it works.

An antenna that is too long for the desired resonant frequency will act inductively (i.e., it will show a feedpoint impedance of the form  $Z = R + jX_L$ ). To counteract the inductive reactance component  $+jX_L$ , it is necessary to add a bit of capacitive reactance  $-jX_C$ . This approach is taken in tuning certain antenna forms and it can be used with dipoles to make the antenna tunable over a range of about 15 percent of the frequency.

The tunable dipole of Fig. 5-15 is longer than other dipoles and it has an overall length of

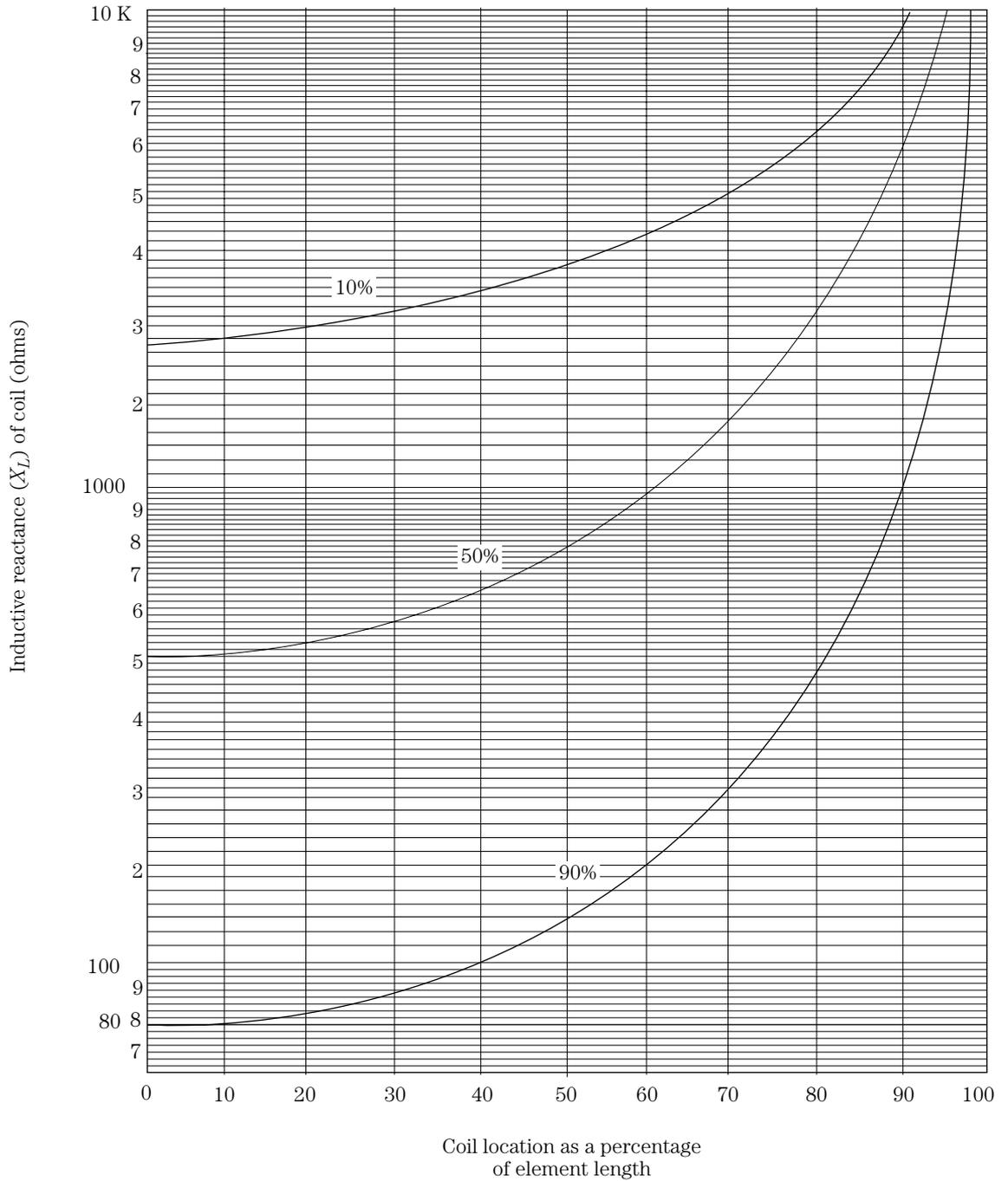
$$L' = \frac{505}{F_{\text{MHz}}} \quad [6.22]$$

Note that the velocity constant is 505, rather than the 468 that is used with ordinary dipoles.

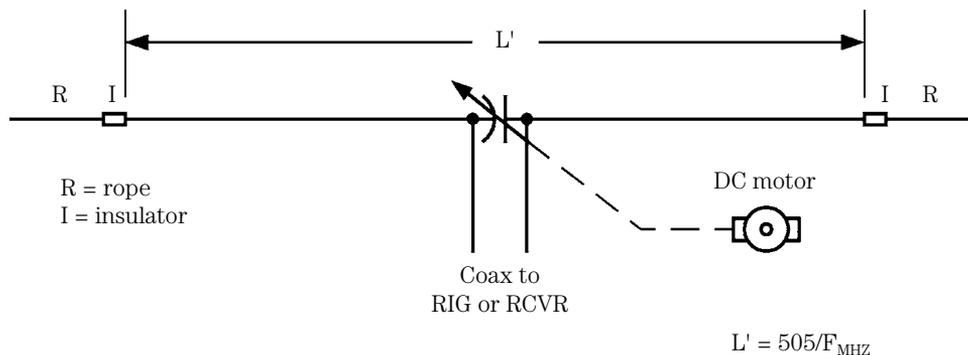
The required capacitive reactance, which is used to electrically shorten the antenna, is provided by  $C_1$ . In one effective design that I've tested, the capacitor was a 500-pF transmitting variable type. When the Heath Company was making kits, they offered an antenna tuner based on this idea. It was a motor-driven 500-pF variable capacitor inside of a weather-resistant metal-shielded case. The Heath tuner can be used on inverted vees and dipoles (if a center support is provided). A low-voltage (which is necessary for safety reasons) dc motor drives the shaft of the capacitor to tune the antenna to resonance.

### Stacked dipoles

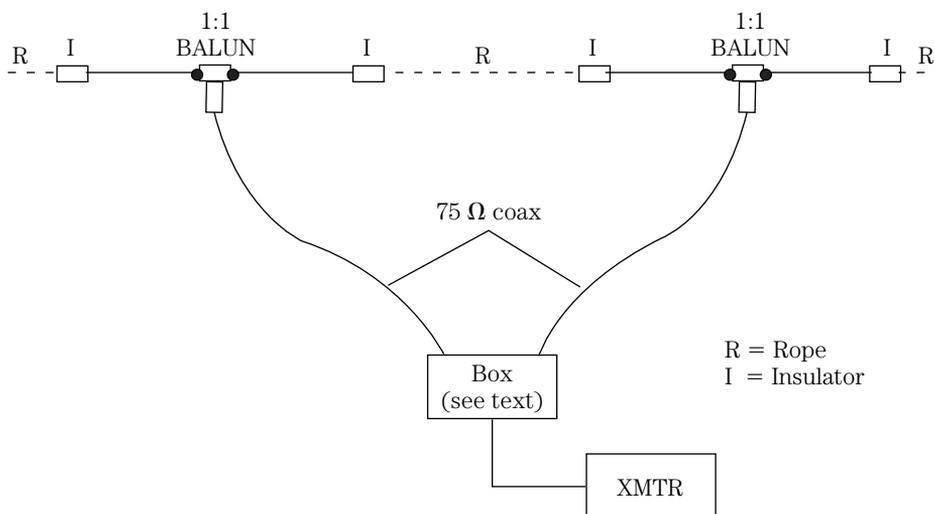
Figure 6-16 shows a *double dipole* (i.e., two half-wavelength dipole antennas spaced a half-wavelength apart). The transmission lines are connected in parallel at the receiver. This antenna provides about 3-dB gain over a single dipole, and it adds a bit of fade protection because two side-by-side antennas provide a bit of space diversity (see the end of Chap. 2 for information on diversity reception).



6-14 Inductive reactance versus coil location for shortened inductance-loaded dipole.



6-15 Electrically reducing the length of a dipole antenna by placing a capacitor at the feedpoint.



6-16 Phased dipoles provide a 3-dB gain over single dipole.

The two dipoles are supported by a common structure consisting of ropes (R) and end insulators (I) to support masts on the ends. In the center, a half-wavelength space is taken up by a rope so that the structure is maintained. The space is determined by

$$\text{Space}_{\text{feet}} = \frac{492}{F_{\text{MHz}}} \quad [6.23]$$

and, for the dipoles,

$$\text{Length}_{\text{feet}} = \frac{468}{F_{\text{MHz}}} \quad [6.24]$$

The stacked dipole gets a bit lengthy on low-frequency bands, but is easily achievable by most people on the upper HF bands.

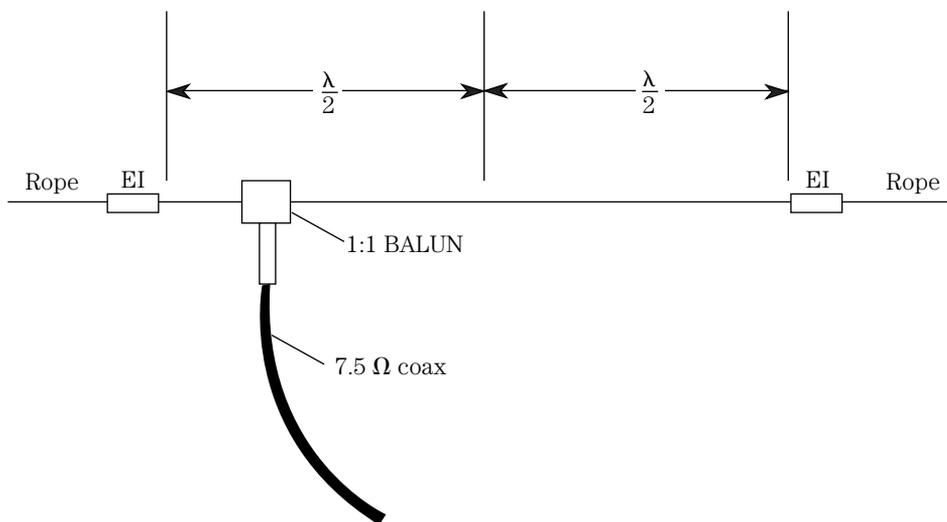
The feedlines for the two dipoles can be connected directly in parallel and fed from the transmitter, provided that the antenna tuning unit will support one-half the normal expected impedance. Alternatively, a box can be provided that includes a matching transformer for 1:2 ratio. These can be built like a balun transformer on a toroidal core. A trifilar winding is used. Alternatively, a phasing box can be built that will allow altering the directionality of the antenna by  $90^\circ$ . This is done by using the switching circuit to reverse the sense of  $L_3$ .

## Off-center-fed full-wave doublet (OCFD) antennas

An antenna that superficially resembles the Windom is the off-center-fed doublet (OCFD) antenna of Fig. 6-17. It is a single-band antenna, although at harmonics it will begin to act as a resonant, standing wave, longwire antenna. The overall length is one wavelength long:

$$L_{\text{feet}} = \frac{936}{F_{\text{MHz}}} \quad [6.25]$$

This antenna works best at heights of at least  $\lambda/2$  above ground, so practical considerations limit it to frequencies above about 10 MHz (i.e., 30-m band). The feedpoint of the antenna is placed at a distance of  $\lambda/4$  from one end, and is a good match for  $75\text{-}\Omega$  coaxial cable. A 1:1 balun transformer at the feedpoint is highly recommended. The pattern of a  $1\lambda$  antenna is a four-lobe “cloverleaf,” with the major lobes being about  $53^\circ$  from the wire. The gain is about 1 dB.



6-17 Off-center-fed 1-wire antenna.

## Off-center-fed nonresonant sloper (OCFS)

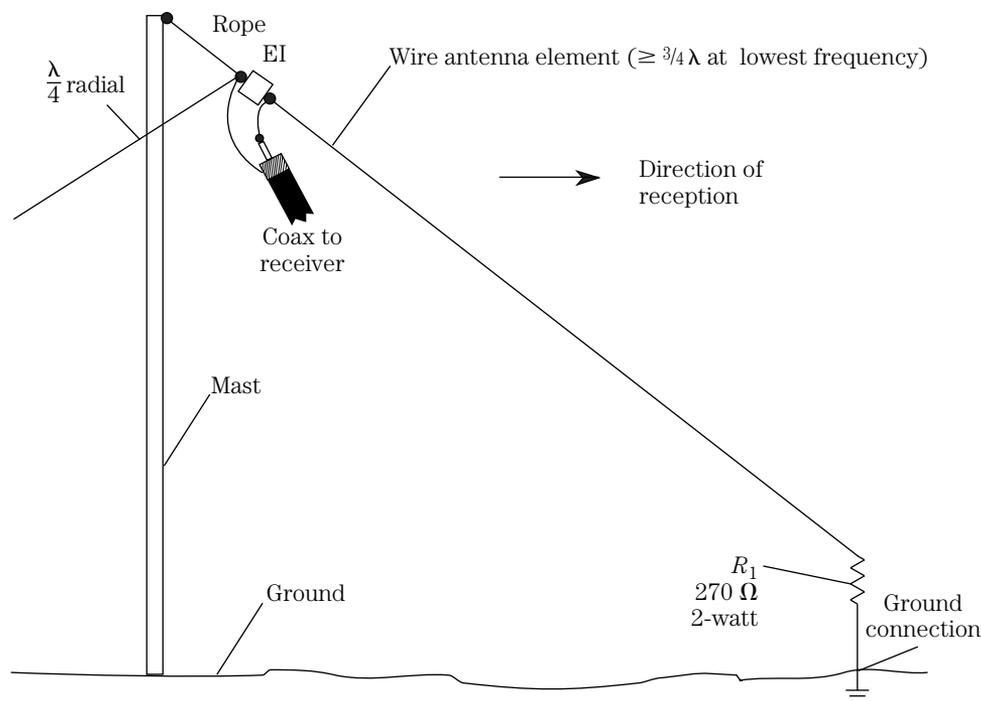
Perhaps more viable for many people is the *nonresonant off-center-fed sloper* (OCFS) antenna of Fig. 6-18. This antenna consists of a wire radiator that must be longer than  $3\lambda/2$  at the lowest frequency of operation. The feedpoint is elevated at least  $\lambda/4$  above ground at the lowest operating frequency. The antenna is fed with 75- $\Omega$  coaxial cable. The shield of the coax is connected to a  $\lambda/4$  resonant radial (counterpoise ground). There should be at least one radial (more is better) per band of operation.

The far end of the radiator element is sloped to ground, where it is terminated in a 270- $\Omega$  noninductive resistor. The resistor should be able to dissipate up to one-third of the power level applied by the transmitter.

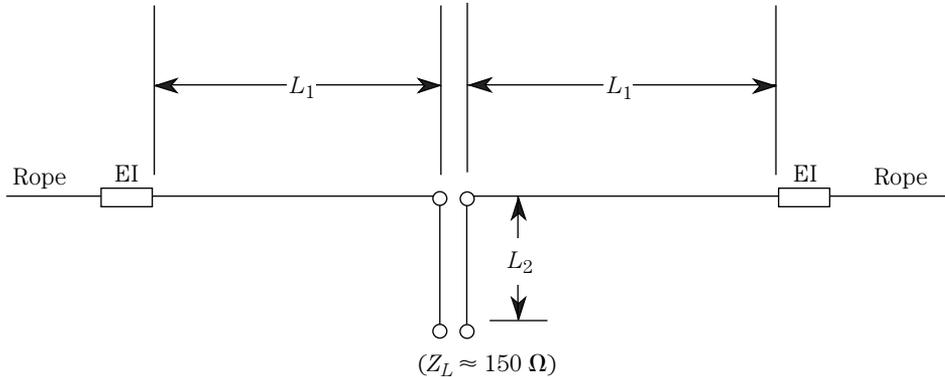
## Double extended Zepp antenna

The *double extended Zepp* antenna (Fig. 6-19) provides a gain of about 2 dB over a dipole at right angles to the antenna wire plane. It consists of two sections of wire, each one of a length

$$L_{1\text{feet}} = \frac{600}{F_{\text{MHz}}} \quad [6.26]$$



6-18 Terminated sloper antenna.



6-19 Double extended Zepp antenna.

Typical lengths are 20.7 ft on the 10-m band, 28 ft on the 15-m band, 42 ft on the 20-m band, and 84 ft on the 40-m band.

The double extended Zepp antenna can be fed directly with 450-Ω twin lead, especially if a balanced antenna tuner is available at the receiver. Alternatively, it can be fed from a quarter-wavelength matching section (made of 450-Ω twin lead, or equivalent open air parallel line), as shown, and a balun if coax is preferred. The length of the matching section should be

$$L_{2\text{feet}} = \frac{103}{F_{\text{MHz}}} \quad [6.27]$$

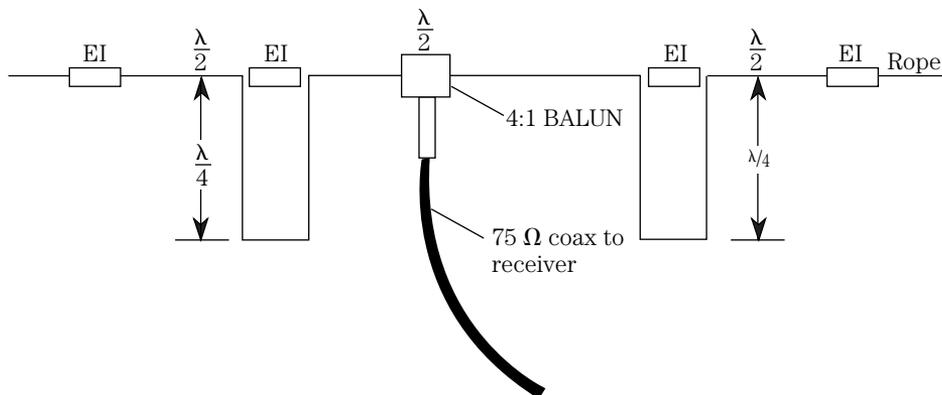
The double extended Zepp will work on several different bands. For example, a 20-m-band double extended Zepp will work as a Zepp on the design band, a dipole on frequencies below the design band, and as a four-lobed cloverleaf antenna on frequencies above the design band.

## Collinear "Franklin" array antenna

Perhaps the cheapest approach to very serious antenna gain is the collinear Franklin array shown in Fig. 6-20. This antenna pushes the dipole and double extended Zepp concepts even farther. It consists of a half-wavelength dipole that is center-fed with a 4:1 balun and 75-Ω coaxial cable. At each end of the dipole, there is a quarter-wavelength *phase reversal stub* that end-feeds another half-wavelength element. Each element is a half-wavelength ( $\lambda/2$ ) long, and its length can be calculated from

$$\frac{\lambda}{2} = \frac{492}{F_{\text{MHz}}} \quad [6.28]$$

The phase reversal stubs are a quarter-wavelength long, or one half the length calculated by Eq. 6.28.



6-20 Wire collinear antenna.

The version of the “Collinear” shown in Fig. 6-20 has a gain of about 3 dB. There is no theoretical reason why you can’t extend the design indefinitely, but there is a practical limit set by how much wire can be held by your supports, and how much real estate you own. A 4.5-dB version can be built by adding another half-wavelength section at each end, with an intervening quarter-wavelength phase reversal stub in between each new section, and the preceding section. Once you get longer than five half-wavelengths, which provides the 4.5-dB gain, the physical size becomes a bit of a bother for most folks.

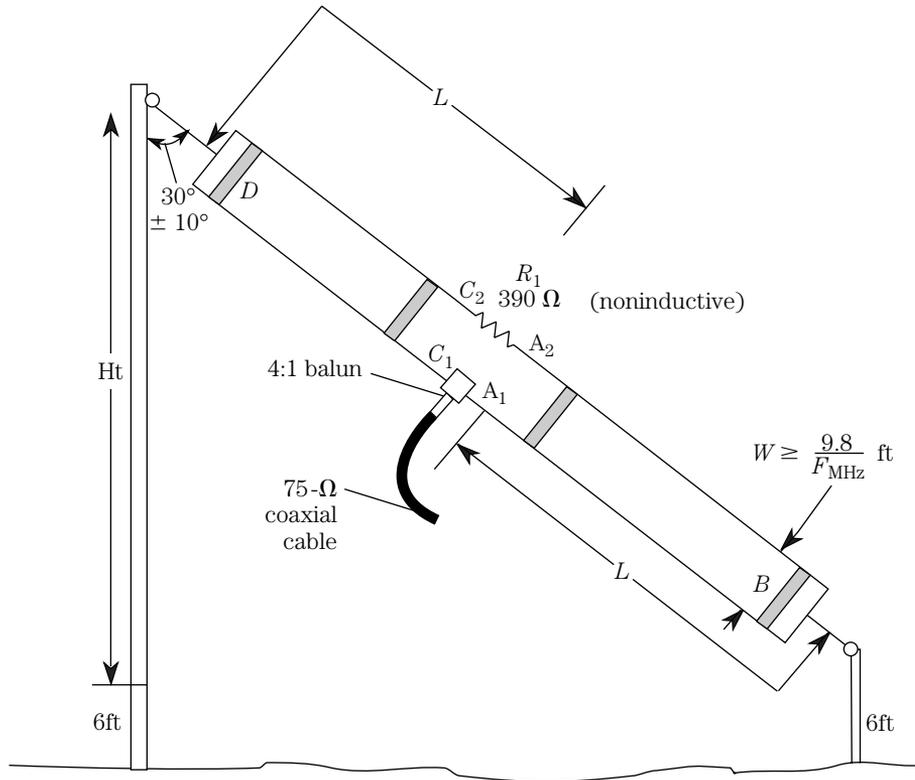
## The TCFTFD dipole

The *tilted, center-fed, terminated, folded dipole* (TCFTFD, also called the T<sup>2</sup>FD or TTFD) is an answer to both the noise pickup and length problems that sometimes affect other antennas. For example, a random-length wire, even with antenna tuner, will pick up considerable amounts of noise. A dipole for 40 m is 66 ft long.

This antenna was first described publicly in 1949 by Navy Captain C. L. Countryman, although the U.S. Navy tested it for a long period in California during World War II. The TCFTFD can offer claimed gains of 4 to 6 dB over a dipole, depending on the frequency and design, although 1 to 3 dB is probably closer to the mark in practice, and less than 1 dB will be obtained at some frequencies within its range (especially where the resistor has to absorb a substantial portion of the RF power). The main attraction of the TCFTFD is not its gain, but rather its *broad bandedness*.

In addition, the TCFTFD can also be used at higher frequencies than its design frequency. Some sources claim that the TCFTFD can be used over a 5 or 6:1 frequency range, although my own observations are that 4:1 is more likely. Nonetheless, a 40-m antenna will work over a range of 7000 to 25,000 kHz, with at least some decent performance up into the 11-m Citizen’s Band (27,000 kHz).

The basic TCFTFD (Fig. 6-21) resembles a folded dipole in that it has two parallel conductors of length  $L$ , spaced a distance  $W$  apart, and shorted together at the



6-21 TTFD antenna.

ends. The feedpoint is the middle of one conductor, where a 4:1 balun coil and 75- $\Omega$  coaxial-cable transmission line to the transceiver are used. A noninductive, 390- $\Omega$  resistor is placed in the center of the other conductor. This resistor can be a carbon-composition (or metal-film) resistor, but it must not be a wirewound resistor or any other form that has appreciable inductance. The resistor must be able to dissipate about one-third of the applied RF power. The TCFTFD can be built from ordinary no.14 stranded antenna wire.

For a TCFTFD antenna covering 40 through 11 m, the spread between the conductors should be 19 $\frac{1}{2}$  in, while the length  $L$  is 27 ft. Note that length  $L$  includes one-half of the 19-in spread because it is measured from the center of the antenna element to the center of the end supports.

The TCFTFD is a sloping antenna, with the lower support being about 6 ft off the ground. The height of the upper support depends on the overall length of the antenna. For a 40-m design, the height is on the order of 50 ft.

The parallel wires are kept apart by spreaders. At least one commercial TCFTFD antenna uses PVC spreaders, while others use ceramic. You can use wooden dowels of between 1-in and  $\frac{5}{8}$ -in diameter; of course, a coating of varnish (or urethane spray) is recommended for weather protection. Drill two holes, of a size sufficient to

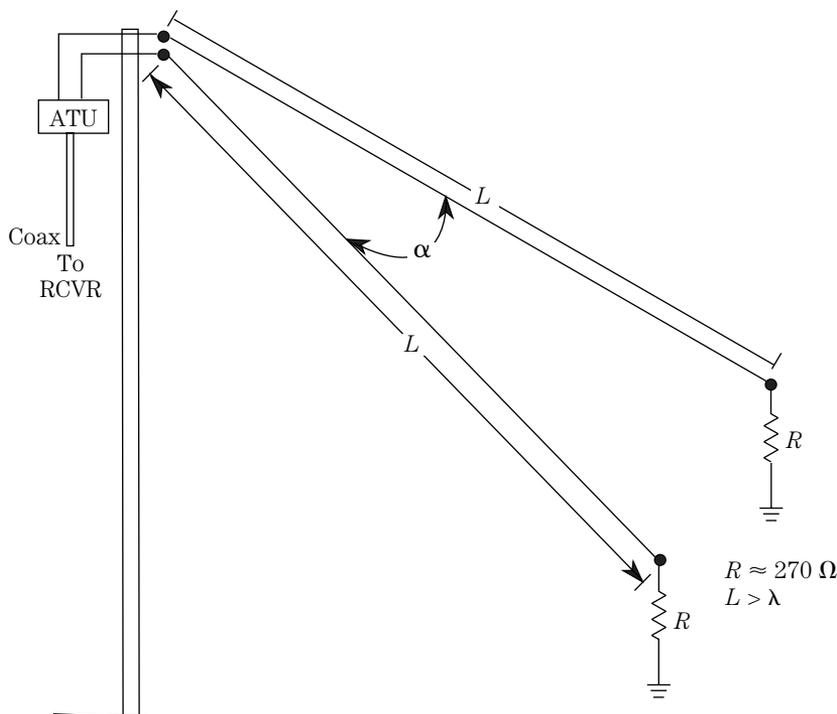
pass the wire, that are the dimension  $W$  apart (19 in for 40 m). Once the spreaders are in place, take about a foot of spare antenna wire and make jumpers to hold the dowels in place. The jumper is wrapped around the antenna wire on either side of the dowel, and then soldered.

The two end supports can be made of  $1 \times 2$  in wood treated with varnish or urethane spray. The wire is passed through screw eyes fastened to the supports. A support rope is passed through two holes on either end of the  $1 \times 2$  and then tied off at an end insulator.

The TCFTFD antenna is noticeably quieter than the random-length wire antenna, and somewhat quieter than the half-wavelength dipole. When the tilt angle is around  $30^\circ$ , the pattern is close to omnidirectional. Although a little harder to build than dipoles, it offers some advantages that ought not to be overlooked. These dimensions will suffice when the “bottom end” frequency is the 40-m band, and it will work well on higher bands.

## Vee-sloper antenna

The *vee-sloper antenna* is shown in Fig. 6-22. It is related to the vee beam (covered in Chap. 9), but it is built like a sloper (i.e., with the feed end of the antenna high above ground). The supporting mast height should be about half (to three-fourths) of the length of either antenna leg. The legs are sloped downward to terminating



6-22 Vee-sloper antenna.

resistors at ground level. Each wire should be longer than  $1\lambda$  at the lowest operating frequency. The terminating resistors should be on the order of  $270\ \Omega$  (about one-half of the characteristic impedance of the antenna), with a power rating capable of dissipating one-third of the transmitter power. Like other terminating resistors, these should be noninductive (carbon composition or metal film).

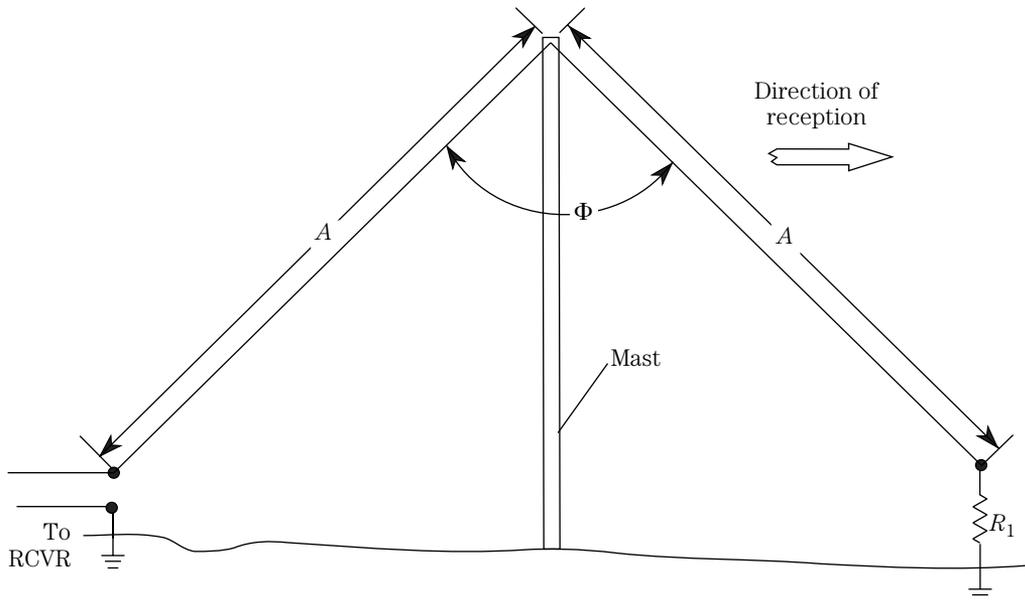
The advantage of this form of antenna over the vee beam is that it is vertically polarized, and the resistors are close to the earth, so they are easily grounded.

## Rhombic inverted-vee antenna

A variation on the theme is the vertically polarized rhombic of Fig. 6-23. Although sometimes called an *inverted vee*—not to be confused with the dipole variant of the same name—this antenna is half a rhombic, with the missing half being “mirrored” in the ground (similar to a vertical). The angle at the top of the mast ( $\Phi$ ) is typically  $\geq 90^\circ$ , and 120 to  $145^\circ$  is more common. Each leg ( $A$ ) should be  $\geq \lambda$ , with the longer lengths being somewhat higher in gain, but harder to install for low frequencies. A requirement for this type of antenna is a very good ground connection. This is often accomplished by routing an underground wire between the terminating resistor ground and the feedpoint ground.

## Multiband fan dipole

The basic half-wavelength dipole antenna is a very good performer, especially when cost is a factor. The dipole yields relatively good performance for practically no in-



6-23 Inverted-vee antenna.

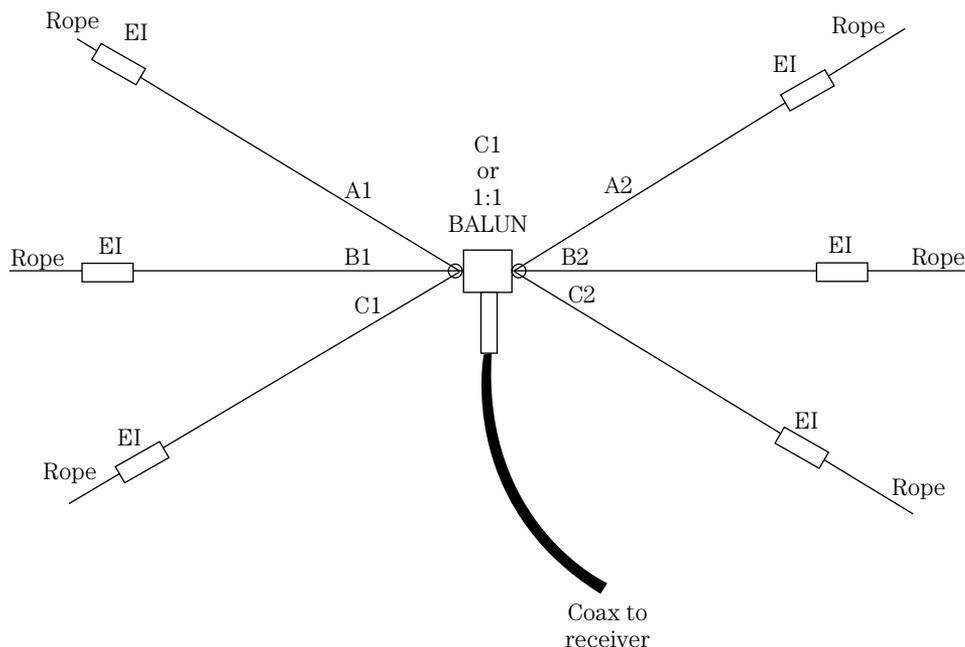
vestment. A standard half-wavelength dipole offers a bidirectional figure-8 pattern on its basic band (i.e., where the length is a half-wavelength), and a four-lobe cloverleaf pattern at frequencies for which the physical length is  $3\lambda/2$ . Thus, a 40-m half-wavelength dipole produces a bidirectional pattern on 40 m, and a four-lobe cloverleaf pattern on 15 m.

The dipole is not easily multibanded without resorting to traps (covered in Chap. 8). One can, however, tie several dipoles to the same center insulator or balun transformer. Figure 6-24 shows three dipoles cut for different bands, operating from a common feedline and balun transformer: A1–A2, B1–B2, and C1–C2. Each of these antennas is a half-wavelength (i.e.,  $L_{\text{feet}} = 468/F_{\text{MHz}}$ ).

There are two points to keep in mind when building this antenna. First, try to keep the ends spread a bit apart, and second, make sure that none of the antennas is cut as a half-wavelength for a band for which another is  $3\lambda/2$ . For example, if you make A1–A2 cut for 40 m, then don't cut any of the other three for 15 m. If you do, the feedpoint impedance and the radiation pattern will be affected.

## The counterpoise longwire

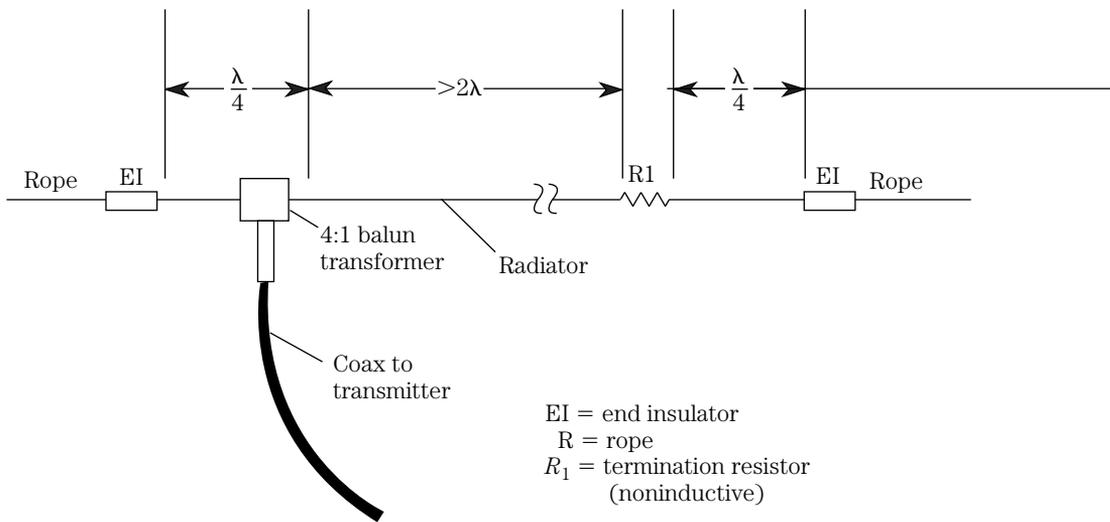
The *longwire antenna* is an end-fed wire more than  $2\lambda$  long. It provides considerable gain over a dipole, especially when a very long length can be accommodated. Although 75- to 80-m, or even 40-m longwires are a bit difficult to erect at most locations, they are well within reason at the upper end of the HF spectrum. Low-VHF



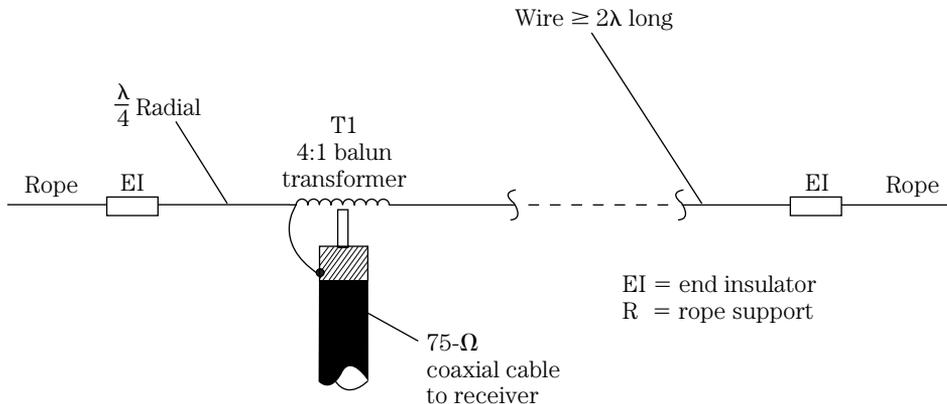
6-24 Multiband dipole antenna.

band operation is also practical. Indeed, I know one fellow who lived in far southwest Virginia as a teenager, and he was able to get his family television reception for very low cost by using a TV longwire (channel 6) on top of his mountain.

There are some problems with longwires that are not often mentioned. Two problems seem to insinuate themselves into the process. First, the Zepp feed is a bit cumbersome (not everyone is enamored of parallel transmission line). Second, how do you go about actually grounding that termination resistor? If it is above ground, then the wire to ground is long, and definitely not at ground potential for RF. If you want to avoid both the straight Zepp feed system employed by most such antennas, as well as the resistor-grounding problem, then you might want to consider the *counterpoise longwire antennas* shown in Fig. 6-25.



6-25A Longwire with  $\lambda/4$  radials used as counterpoise ground at feedpoint and terminating resistor.



6-25B Balun transformer feed for longwire antenna.

A counterpoise ground is a structure that acts like a ground, but is actually electrically floating above real ground (and it is not connected to ground). A groundplane of radials is sometimes used as a counterpoise ground for vertical antennas that are mounted above actual earth ground. In fact, these antennas are often called *ground plane verticals*. In those antennas, the array of four (or more) radials from the shield of the coaxial cable are used as an artificial, or counterpoise, ground system.

In the counterpoise longwire of Fig. 6-25A, there are two counterpoise grounds (although, for one reason or another, you might elect to use either, but not both). One counterpoise is at the feedpoint, where it connects to the “cold” side of the transmission line. The parallel line is then routed to an antenna tuning unit (ATU), and from there to the transmitter. The other counterpoise is from the cold end of the termination resistor to the support insulator. This second counterpoise makes it possible to eliminate the earth ground connection, and all the problems that it might entail, especially in the higher end of the HF spectrum, where the wire to ground is of substantial length compared with  $1\lambda$  of the operating frequency.

A slightly different scheme used to adapt the antenna to coaxial cable is shown in Fig. 6-25B. In this case, the longwire is a resonant type (nonterminated). Normally, one would expect to find this antenna fed with  $450\text{-}\Omega$  parallel transmission line. But with a  $\lambda/4$  radial acting as a counterpoise, a 4:1 balun transformer can be used to effect a reasonable match to  $75\text{-}\Omega$  coaxial cable. The radial is connected to the side of the balun that is also connected to the coaxial cable shield, and the other side of the balun is connected to the radiator element.

## 7 CHAPTER

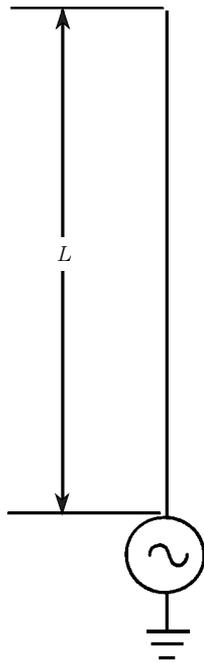
# Vertically polarized HF antennas

IN PREVIOUS CHAPTERS, YOU HAVE FOUND THAT THE POLARITY OF AN ANTENNA IS THE direction of the electrical (E) field. Because the transmitted signal is an orthogonal electromagnetic wave, the magnetic field radiated from the antenna is at right angles to the electric field. The direction of the electric field, which sets the polarity of the antenna, is a function of the geometry of the radiator element. If the element is vertical, then the antenna polarity is also vertical. The signal propagates out from the radiator in all directions of azimuth, making this antenna an “omnidirectional” radiator.

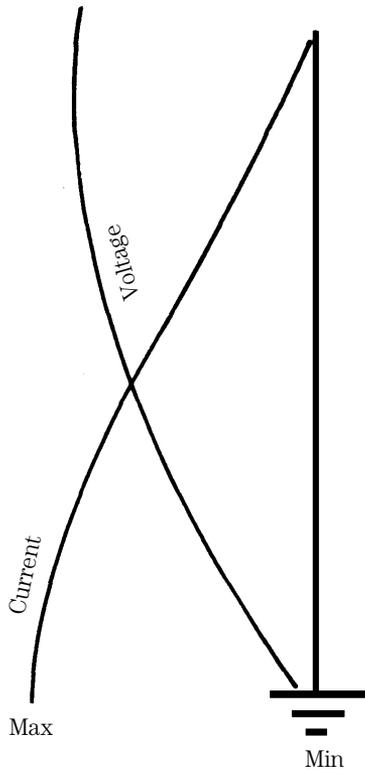
Figure 7-1A shows the basic geometry of the vertical antenna: an RF generator (transmitter or transmission line from a transmitter) at the base of a radiator of length  $L$ . Although most commonly encountered verticals are quarter-wavelength ( $L = \lambda/4$ ), that length is not the only permissible length. In fact, it may not even be the most desirable length. This chapter covers the standard quarter-wavelength vertical antenna (because it is so popular), and other-length verticals (both greater and less than quarter-wavelength).

The quarter-wavelength vertical antenna is basically half of a dipole placed vertically, with the “other half” of the dipole being the ground. Because of this fact, some texts show the vertical with a double-line ghost radiator, or image antenna, in the earth beneath the main antenna element. Figure 7-1B shows the current and voltage distribution for the quarter-wavelength vertical. Like the dipole, the quarter-wavelength vertical is fed at a current loop, so the feedpoint impedance is at a minimum (typically 2 to 37  $\Omega$ , depending upon nearby objects and design). As a result, the current is maximum and the voltage is minimum at the feedpoint. As you will see, however, not all vertical antennas are fed directly at the current loop. As a result, some designs require antenna tuning units to make them match the antenna impedance to the transmitter output impedance.

Figure 7-1C and 7-1D show the two basic configurations for the HF vertical antenna. Figure 7-1C shows the ground-mounted vertical antenna. The radiator

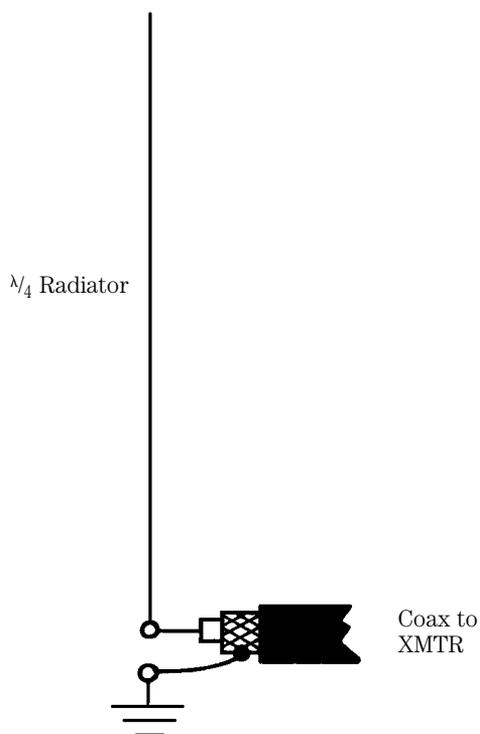


7-1A Basic elements of the vertical antenna.

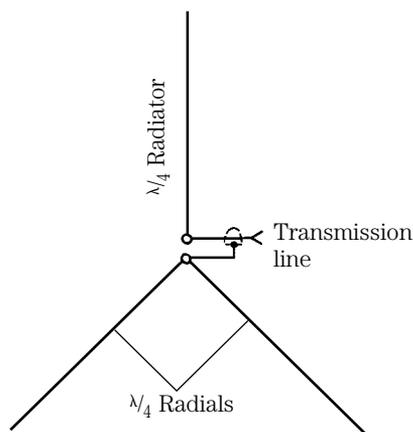


7-1B Current and voltage distribution along vertical.

**7-1C** Simple coaxial-fed vertical antenna. There is a slight mismatch, but it usually is within tolerable limits.



**7-1D** Mast-mounted vertical uses radials as a counterpoise ground.



element is mounted at ground level, but it is insulated from ground. Because the antenna shown is a quarter-wavelength, it is fed at a current loop with  $52\text{-}\Omega$  coaxial cable. The inner conductor of the coaxial cable is connected to the radiator element, and the coaxial cable shield is connected to the ground. As you will see shortly, the ground system for the vertical antenna is critical to its performance. Normally, the feedpoint impedance is not  $52\ \Omega$ , but rather is somewhat lower ( $37\ \Omega$ ). As a

result, without some matching there will be a slight VSWR, but in most cases, the VSWR is a tolerable tradeoff for simplicity. If the antenna has a feedpoint impedance of  $37 \Omega$ , which is the value usually quoted, then the VSWR will be  $52 \Omega/37 \Omega$ , or 1.41:1.

A vertical mounted above the ground level is shown in Fig. 7-1D. This antenna is equally as popular as the ground mounted. Amateurs and CB operators find it easy to construct this form of antenna because the lightweight vertical can be mounted at reasonable heights (15 to 60 ft) using television antenna slip-up telescoping masts that are reasonably low in cost. A problem with the non-ground-level vertical antenna is that there is no easy way to connect it to ground. The solution to the problem is to create a counterpoise (artificial) ground with a system of quarter-wavelength radials.

In general, at least two radials are required for each band, and even that number is marginal. The standard wisdom holds that the greater the number of radials, the better the performance. Although that statement is true, there are both theoretical and practical limits to the number of radials. The theoretical limit is derived from the fact that more than 120 radials returns practically no increase in operational effectiveness, and at more than 16 radials, the returned added effectiveness per new radial is less than is the case for fewer radials. That is, going from 16 to 32 radials (doubling the number) creates less of an increase in received field strength at a distant point than going from 8 to 16 radials (both represent doubling the density of the radial system). The practical limit is 16 radials.

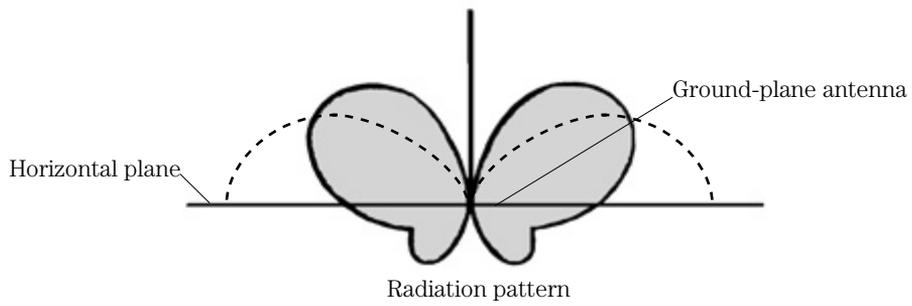
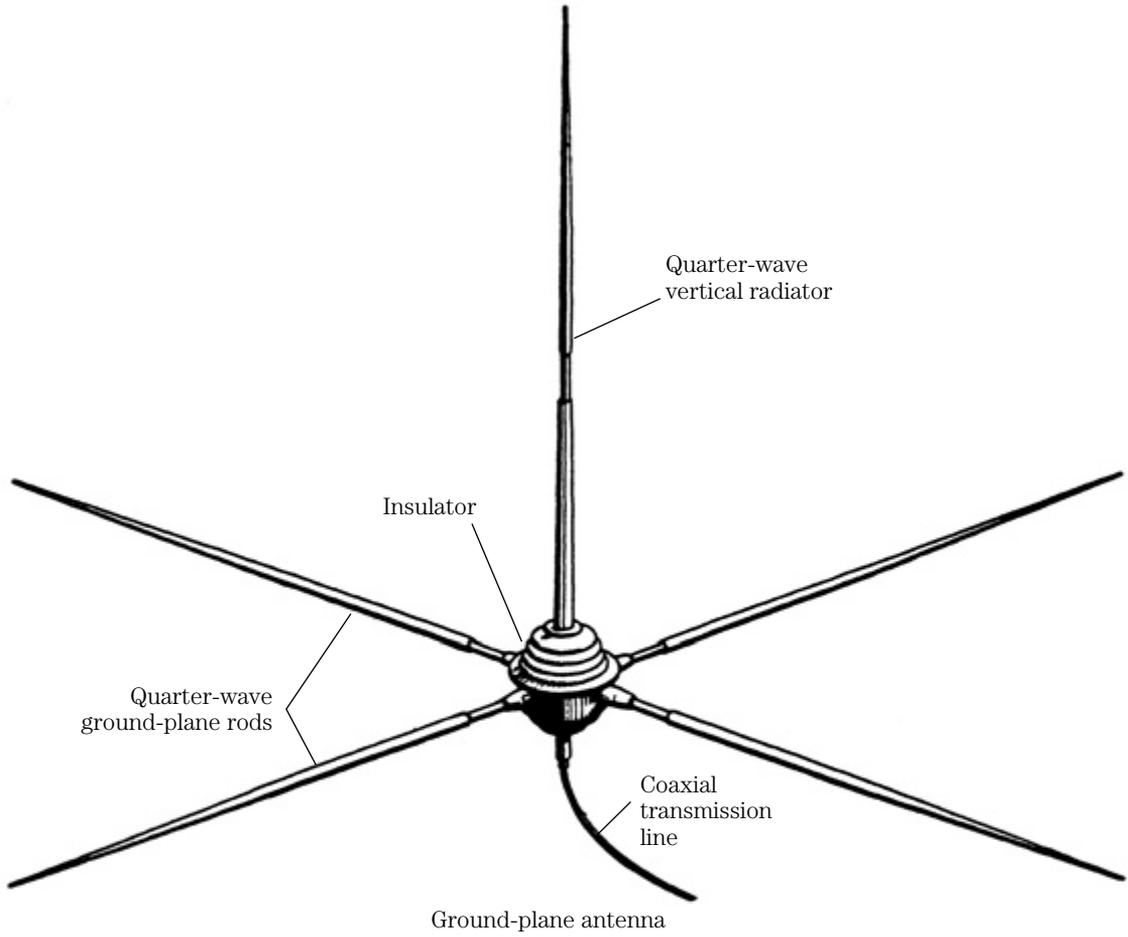
The radials of the off-ground-level vertical antenna can be at any angle. In Fig. 7-1D, they are “drooping radials” (i.e., the angle is greater than  $90^\circ$  relative to the vertical radiator element). Similarly, Fig. 7-1E shows a vertical antenna that is equipped with radials at exactly  $90^\circ$  (no common antenna has radials less than  $90^\circ$ ). Both of these antennas are called *ground plane vertical antennas*.

The angle of the vertical’s radials is said to affect the feedpoint impedance and the angle of radiation of the vertical antenna. Although those statements are undoubtedly true in some sense, there are other factors that also affect those parameters, and they are probably more important in most practical installations. Before digging further into the subject of vertical antennas, take a look at the subjects of angle of radiation and gain in vertical antennas.

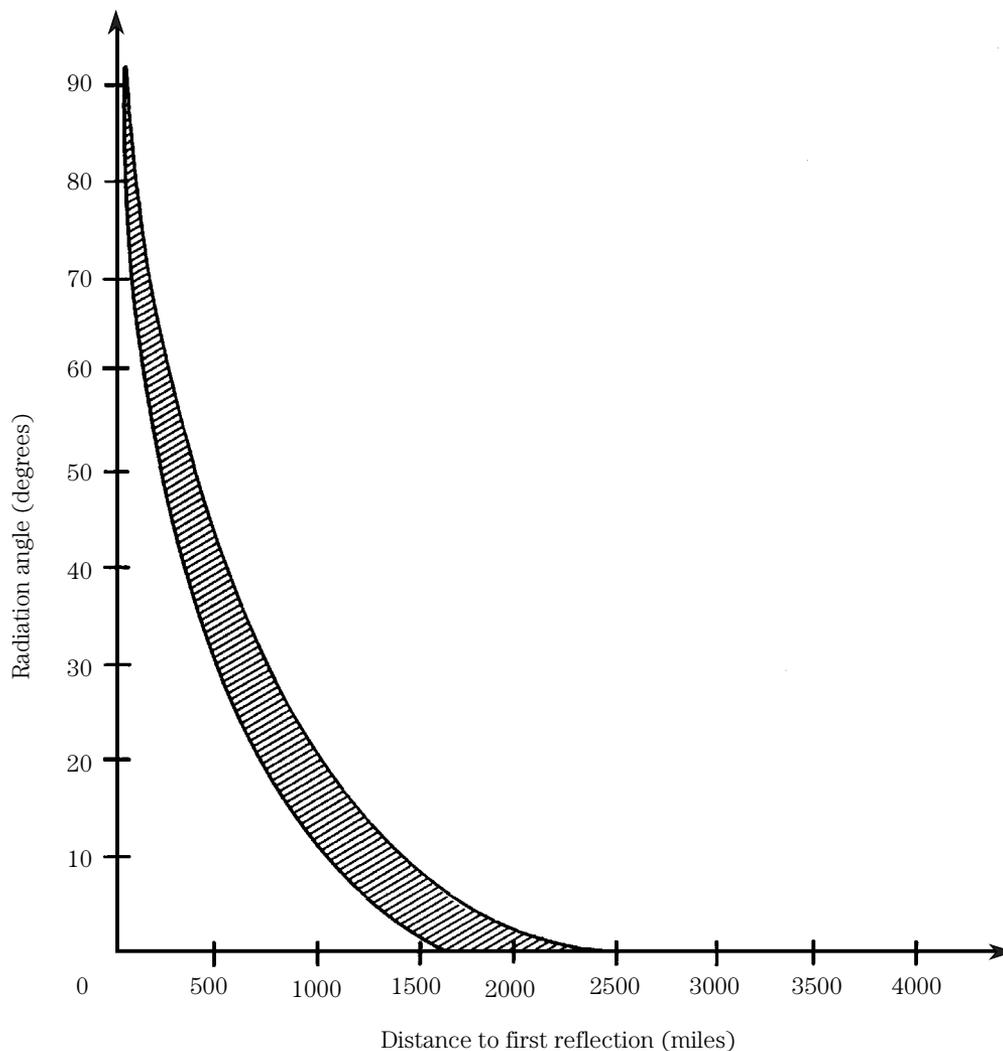
## Angle of radiation

Long-distance propagation in the HF region depends upon the ionospheric phenomena called “skip.” (See Chap. 2 for a more extensive explanation.) In this type of propagation, the signal leaves the transmitting antenna at some angle  $a$ , called the *angle of radiation*, and enters the ionosphere where it is refracted back to earth at a distance  $D$  from the transmitting station. The signal in the zone between the outer edge of the antenna’s ground-wave region and the distant skip point is weak or nonexistent.

The distance covered by the signal on each skip is a function of the angle of radiation. Figure 7-2 shows a plot of the angle of radiation of the antenna, and the distance to the first skip zone. The angle referred to along the vertical axis is the angle of radiation away from the antenna relative to the horizon. For example, an angle of



7-1E Ground-plane vertical antenna.



7-2 Effect of radiation angle on skip communications.

$10^\circ$  is elevated  $10^\circ$  above the horizon. Shorter distances are found when the angle of radiation is increased. At an angle of about  $30^\circ$ , for example, the distance per skip is only a few hundred miles.

Although you might expect on first blush to see a single line on the graph, there is actually a zone shown (shaded). This phenomenon exists because the ionosphere is found at different altitudes at different times of the day and different seasons of the year. Generally, however, in the absence of special event phenomena in the ionosphere, you can expect from 1500 to 2500 mi per bounce in the HF bands for low angles of radiation. Note, for example, that for a signal that is only a degree or two above the horizon the skip distance is maximum.

At distances greater than those shown in Fig. 7-2, the signal will make multiple hops. Given a situation where the skip distance is 2500 mi, covering a distance of 7500 mi requires three hops. Unfortunately, there is a signal strength loss on each hop of 3 to 6 dB, so you can expect the distant signal to be attenuated from making multiple hops between the earth's surface and the ionosphere. For maximizing distance, therefore, the angle of radiation needs to be minimized.

So what is the ideal angle of radiation? It is standard—but actually erroneous—wisdom among amateur radio operators (and even commercial operators, it turns out) that the lower the angle of radiation, the better the antenna. That statement is only true if long distance is wanted, so it reflects a strong bias toward the DX community. The correct answer to the question is: “It depends on where you want the signal to go.” For example, I live in Virginia. If I want to communicate with stations in the Carolinas or New England, then it would behoove me to select a higher angle of radiation for radio conditions represented in Fig. 7-2 so that the signal will land in those regions. But if I wanted to work stations in Europe or Africa or South America, then a low angle of radiation is required. Because of the difference between performance of high and low angles of radiation, some stations have two antennas for each band: one each for high and low angles of radiation.

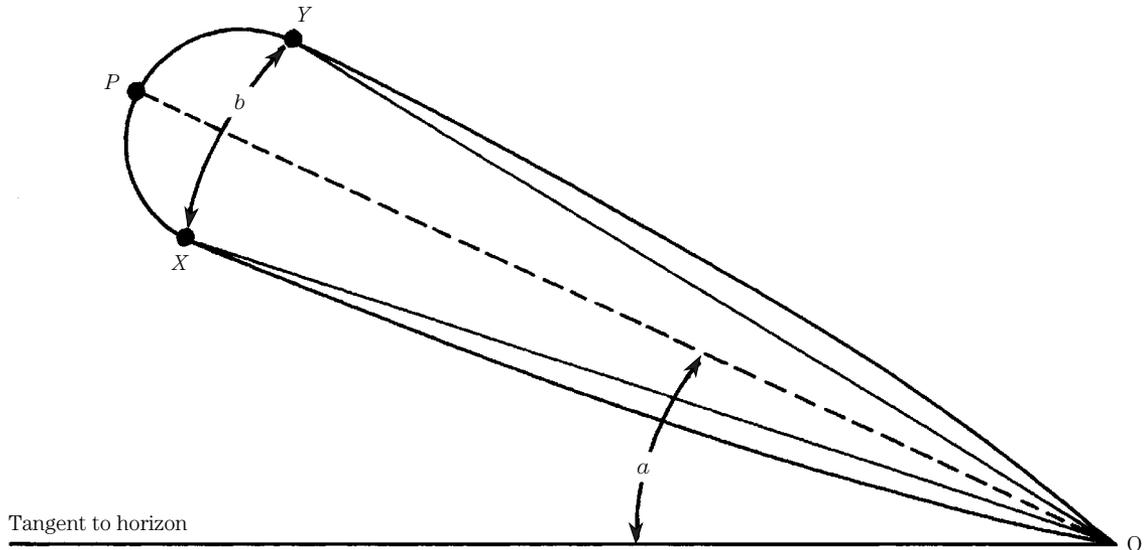
Figure 7-3 shows a signal from a hypothetical antenna located at point  $O$ , in order to show what angle is meant by “angle of radiation.” The beam from the antenna is elevated above the horizon (represented by the horizontal “tangent to horizon” line). The angle of radiation  $a$  is the angle between the tangent line and the center of the beam. This angle is not to be confused with the beamwidth, which is also an angle. In the case of beamwidth, we are talking about the thickness of the main lobe of the signal between points where the field strength is  $-3$  dB down from the maximum signal (which occurs at point  $P$ ); these points are represented by points  $X$  and  $Y$  in Fig. 7-3. Thus, angle  $b$  is the beamwidth, and angle  $a$  is the angle of radiation.

## Gain in vertical antennas

Vertical antennas are considered *omnidirectional* because they radiate equally well in all directions. “Gain” in an antenna is not the creation of power, but rather a simple refocusing of energy from all directions to a specific direction; gain therefore infers directivity. According to the convention, then, the vertical antenna cannot have any gain because it radiates in all directions equally . . . gain infers directivity. Right? No, not really. Let's develop the theme more carefully.

Again consider the idea of an isotropic radiator (the word *isotropic* means equal power in all directions). Consider a spherical point source radiator located at point  $O$  in Fig. 7-4. Whatever the level of power available from the transmitter, it will be spread equally well over the entire surface of the sphere as it radiates out into space away from point  $O$ . If you measure the power distributed over some area  $A$  at a distance  $R$  from the source, then the power available will be a fraction of the total power:

$$P_{\text{avail}} = \frac{\text{total available power} \times A}{\text{total surface area of sphere}}$$



*X, Y*: -3 dB points  
*P*: center beam  
 Angle *a*: elevation angle  
 Angle *b*: 3-dB beam width

7-3 Side view of vertical extent radiation pattern.

or, in math symbols:

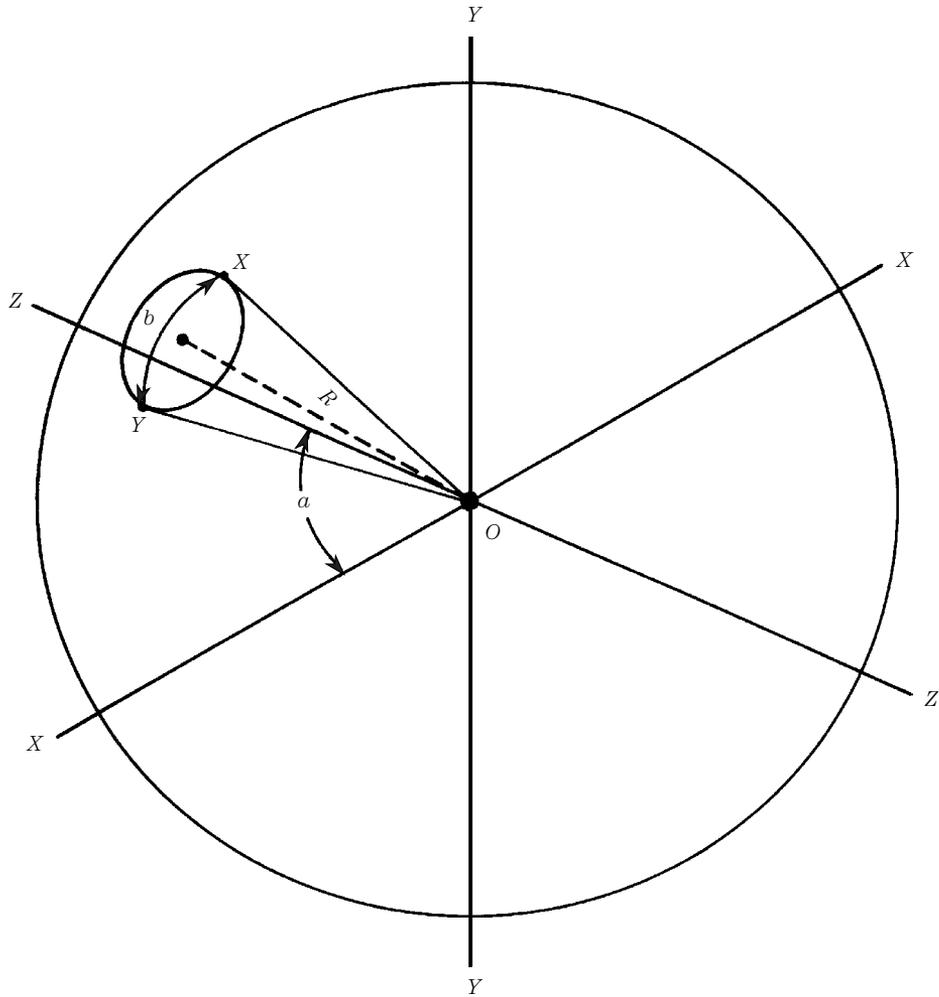
$$P_a = \frac{P_s}{4\pi r^2} \tag{7.1}$$

where

- $P_a$  is the power available per solid degree
- $P_s$  is the total radiated power, in watts
- $r$  is the radius of the sphere (i.e., the distance from  $O$  to  $P$ ).

A practical rule of thumb for this problem is to calculate from the surface area of the sphere. If you perform the right calculations, you will find that there are approximately 41,253 square degrees on the surface of a sphere. By calculating the surface area of the beam front (also in square degrees), you can find the power within that region.

Now for the matter of gain in a vertical antenna. The vertical is not gainless because it does not, in fact, radiate equally well in all directions. In fact, the vertical is quite directional, except in the horizontal (azimuth) plane. Figure 7-5 shows the radiation pattern of the typical free-space vertical radiator. The pattern looks like a giant doughnut in free space (see solid pattern in Fig. 7-5). When sliced like a bagel, the pattern is the familiar circular “omnidirectional” pattern. When examined in the vertical plane, however, the plane looks like a sliced figure-8. The “gain” comes from the fact that energy is not spread over an entire sphere, but rather it is concentrated

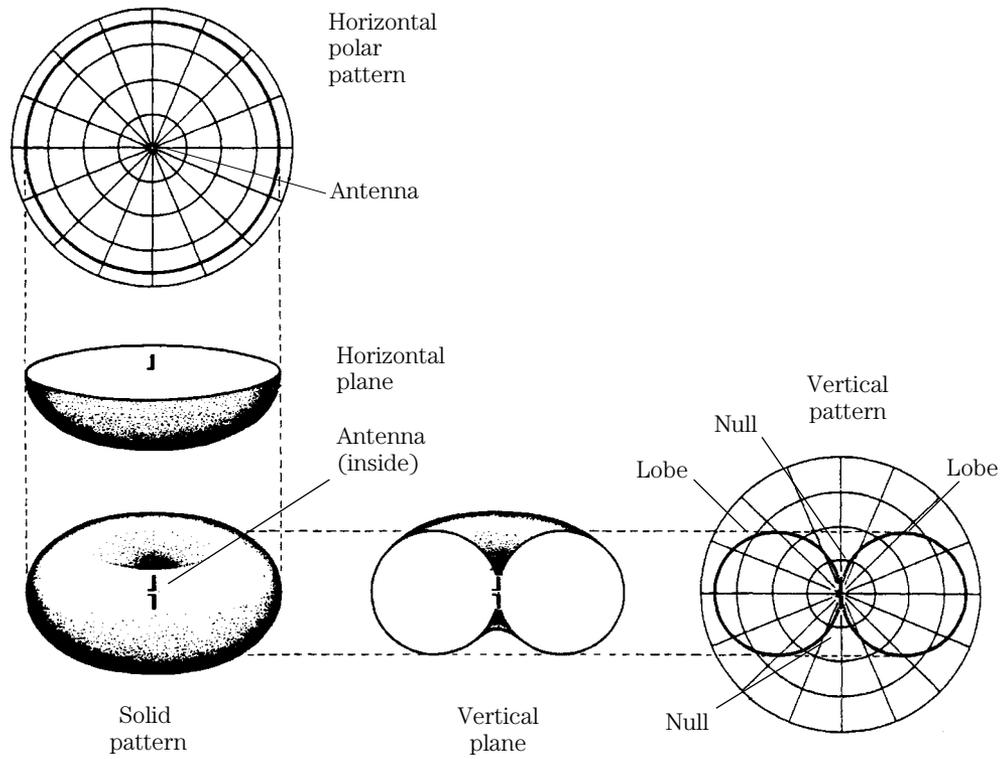


7-4 Spherical or “isotropic” radiation pattern.

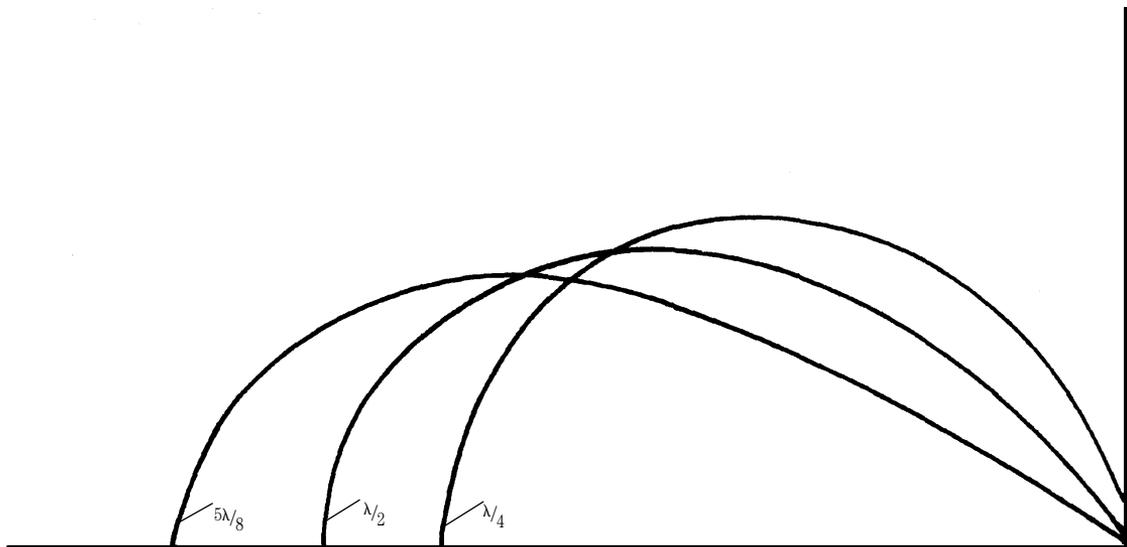
to the toroidal doughnut-shaped region shown. Therefore, the power per unit area is greater than for the isotropic (truly omnidirectional) case.

## Non-quarter-wavelength verticals

The angle of radiation for a vertical antenna, hence the shape of the hypothetical doughnut radiation pattern, is a function of the length of the antenna. (*Note:* Length in terms of vertical antennas is the same as height, and is sometimes expressed in degrees of wavelength, as well as feet and/or meters.) Figure 7-6A shows the approximate patterns for three different-length vertical antennas: quarter-wavelength, half-wavelength, and  $\frac{3}{4}$ -wavelength. Note that the quarter-wavelength antenna has the highest angle of radiation, as well as the lowest gain of the three cases.



7-5 Vertical antenna radiation pattern.



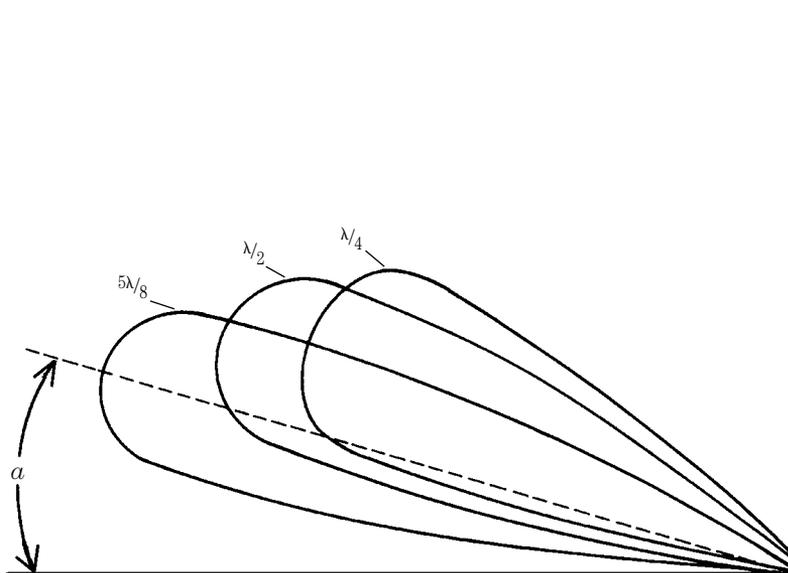
7-6A Vertical radiation pattern as a function of element length: Ideal.

The  $\frac{5}{8}$ -wavelength antenna is both the lowest angle of radiation and the highest gain (compared with isotropic).

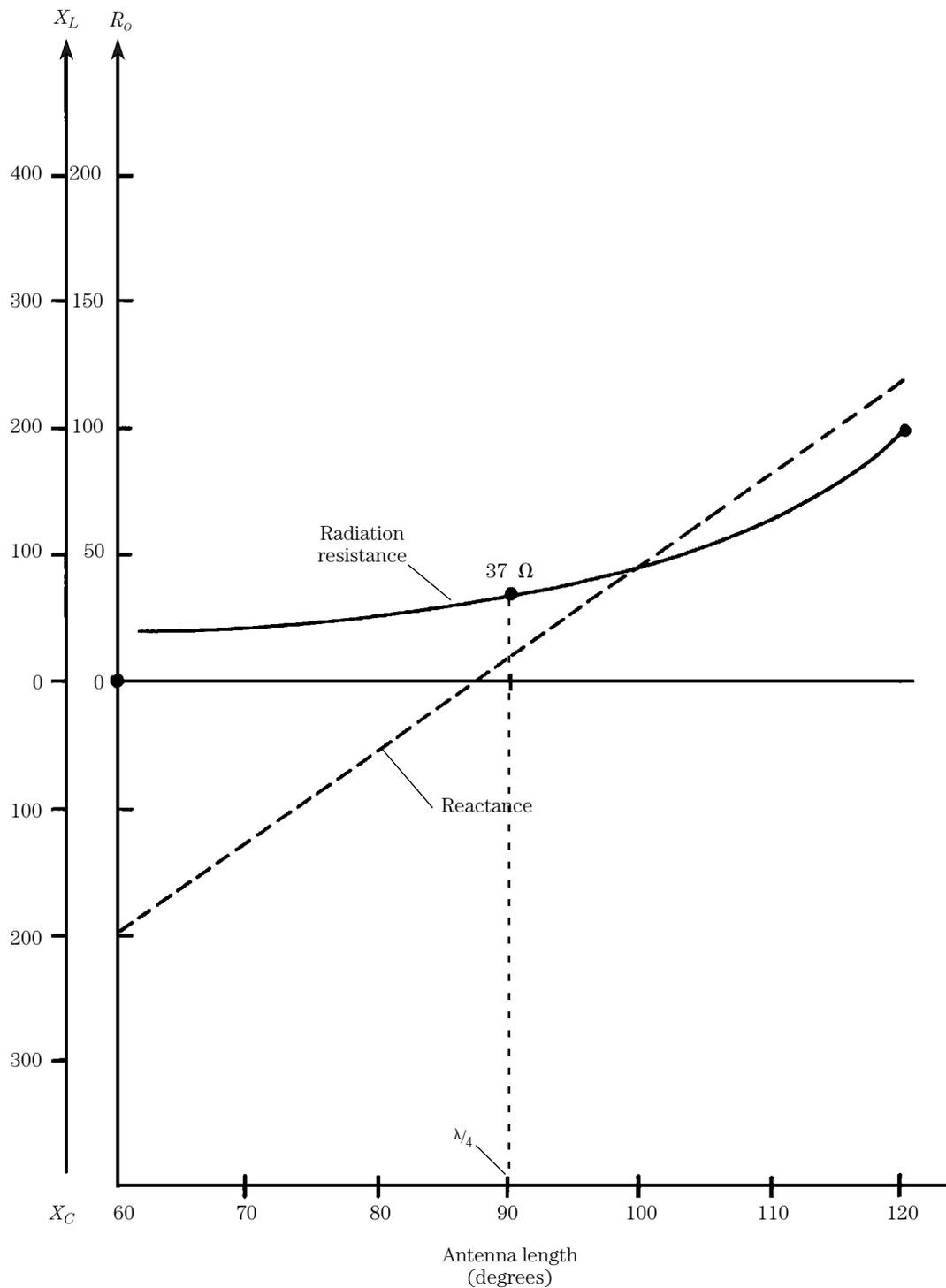
The patterns shown in Fig. 7-6A assume a perfectly conducting ground underneath the antenna. However, that is not a possible situation for practical antennas; all real grounds are lossy. The effect of ground losses is to pull in the pattern close to the ground (Fig. 7-6B). Although all of the patterns are elevated from those of Fig. 7-6A, the relationships still remain: the  $\frac{5}{8}$ -wavelength radiator has the lowest angle of radiation and highest gain.

The feedpoint impedance of a vertical antenna is a function of the length of the radiator. For the standard quarter-wavelength antenna, the feedpoint radiation resistance is a maximum of  $37 \Omega$ , with only a very small reactance component. Figure 7-7 shows the approximate feedpoint impedances for antennas from nearly zero effective length to  $120^\circ$  of length.

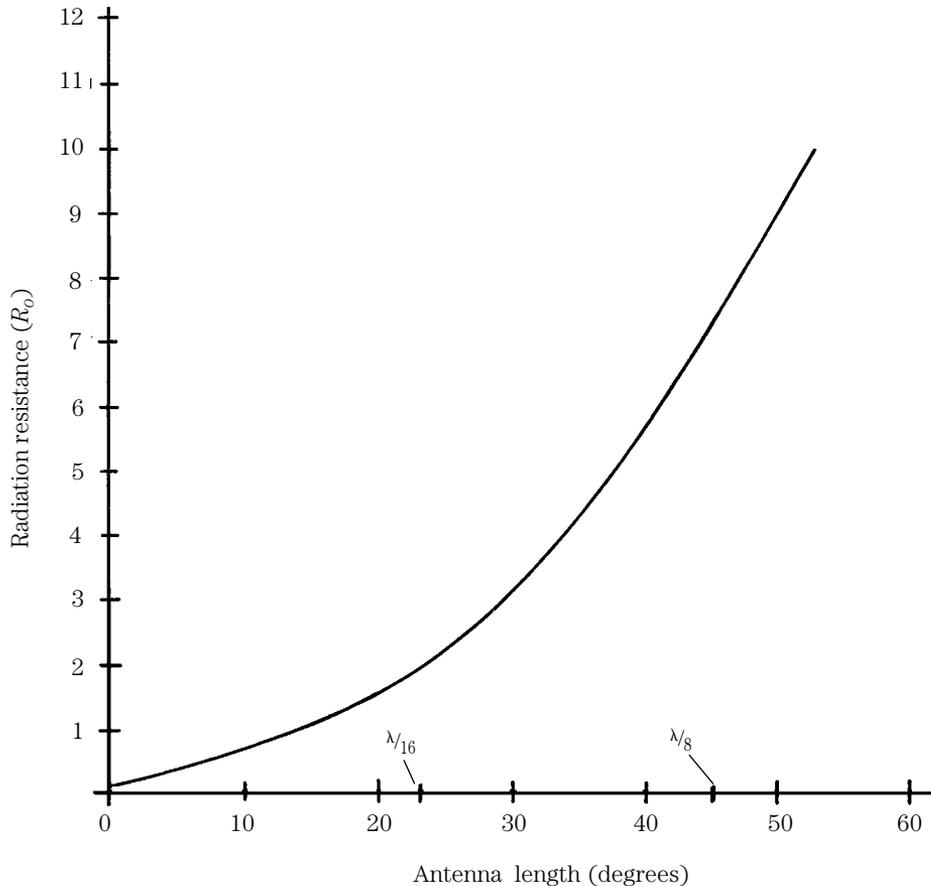
Antenna length expressed in degrees derives from the fact that one wavelength is  $360^\circ$ . Thus, a quarter-wavelength antenna has a length of  $360^\circ/4 = 90^\circ$ . To convert any specific length from degrees to wavelength, divide the length in degrees by 360. Thus, for a  $90^\circ$  antenna:  $90^\circ/360^\circ = \frac{1}{4}$  wavelength. The graph in Fig. 7-7A shows the antenna feedpoint impedance, both reactance and radiation resistance, for antennas from  $60$  to  $120^\circ$ ; Fig. 7-7B shows the radiation resistance for antennas from near zero to  $60^\circ$ . Note that the radiation resistance for such short antennas is extremely small. For example, an antenna that is  $30^\circ$  long ( $\frac{30}{360} = 0.083$  wavelengths) has a resistance of approximately  $3 \Omega$ . It is generally the practice on such antennas to use a broadband impedance-matching transformer to raise the impedance of such antennas to a higher value (Fig. 7-8).



7-6B Accounting for ground losses.



7-7A Antenna impedance as a function of antenna length.

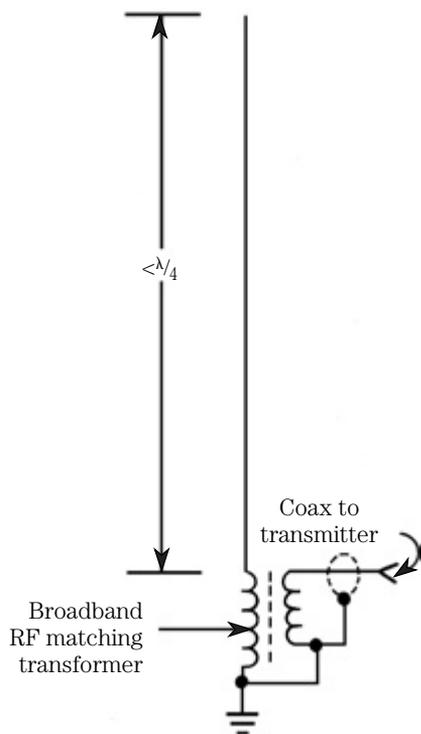


7-7B Radiation resistance as a function of antenna length.

## The ground system for the vertical antenna

The vertical antenna works well only when placed over a good ground system. Chapter 28 gives details of proper ground systems for all radio antennas, verticals included, so only certain specifics are included here. The reader should also examine Chap. 28, however, in order to get a firmer grasp of the problem.

The usual way to provide a good ground for a vertical is to use a system of radials such as Fig. 7-9A. This case shows a view (from above) of 16 quarter-wavelength radials arranged to cover the full circle around the antenna. Each radial is a quarterwavelength, so it will have a length (in fact) of  $246/F_{\text{MHz}}$ . All of the radials are connected together at the base of the antenna, and the ground side of the transmission line is connected to this system. The radials can be placed either on the surface or underground. One friend of the author built an extensive radial system on the bare dirt when his house was built, so when the sod was installed he had a very high-quality underground radial system.

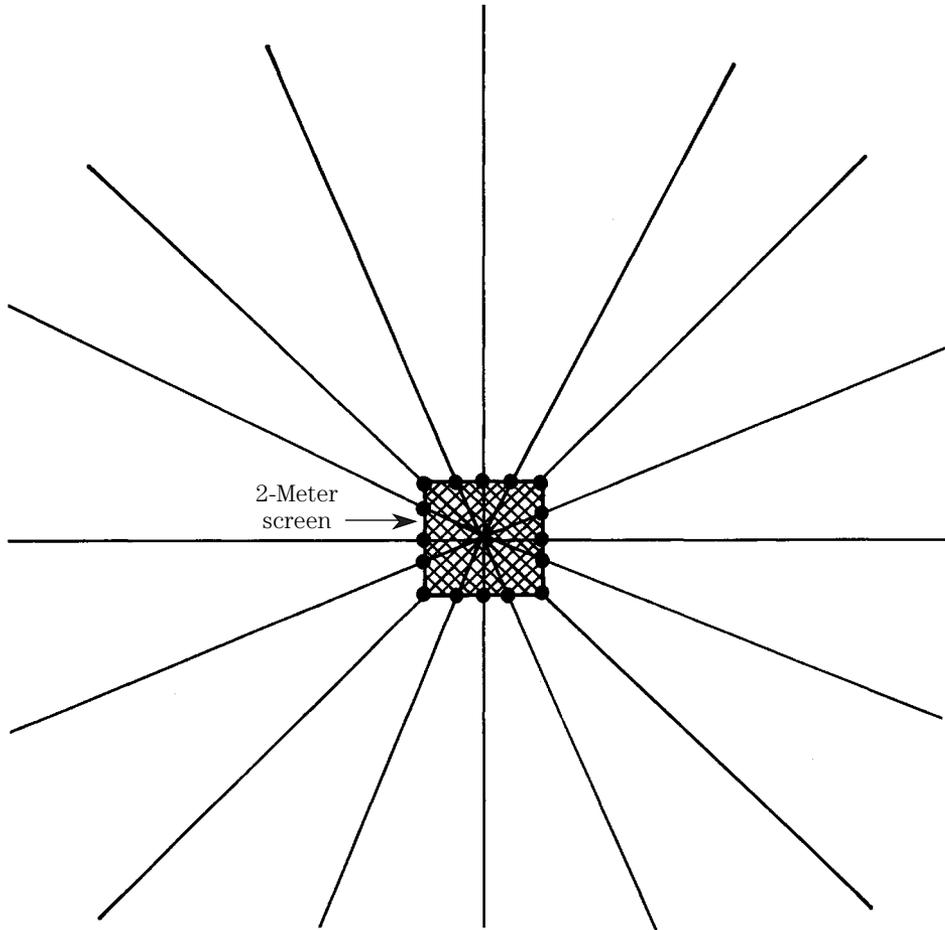


7-8 Impedance matching through broadband transformer.

If you decide to use an aboveground radial system, however, be sure to prevent people from tripping over it. There might be liability implications for people who trip and injure themselves, even when the person is an intruder or trespasser.

Some experts prefer to place a copper wire screen at the center of the radial system. The minimum size of this screen is about 2m square ( $6 \times 6$  ft). Connect it to the radials at the points shown using solder. Other experts will drive ground stakes into the ground at these points. Still another method is shown in Fig. 7-9B. This case shows a "spider web" of conductors shorting the radials at points a meter or two from the antenna. Again, some authorities recommend that ground rods be driven into the earth at the indicated points.

The exact number of radials to use depends in part on practical matters (how many can you physically install?). Use at least two radials per band, with four per band preferred for simple, low-cost systems. However, even four is considered a compromise case. The general rule is: *the more the better*. But it's also true that there is a law of diminishing returns as the number of radials is increased. Figure 7-9C shows the approximate field intensity (mV/m) as a function of the number of radials. Notice that the field intensity does not increase as rapidly with the number of radials above 20 or so. Note that the Federal Communications Commission requires AM band (550 to 1700 kHz) stations to use 120 radials, but that number is not necessary for amateur stations. A practical upper limit of 16 radials is usually accepted for amateur radio work, and four radials work quite well.

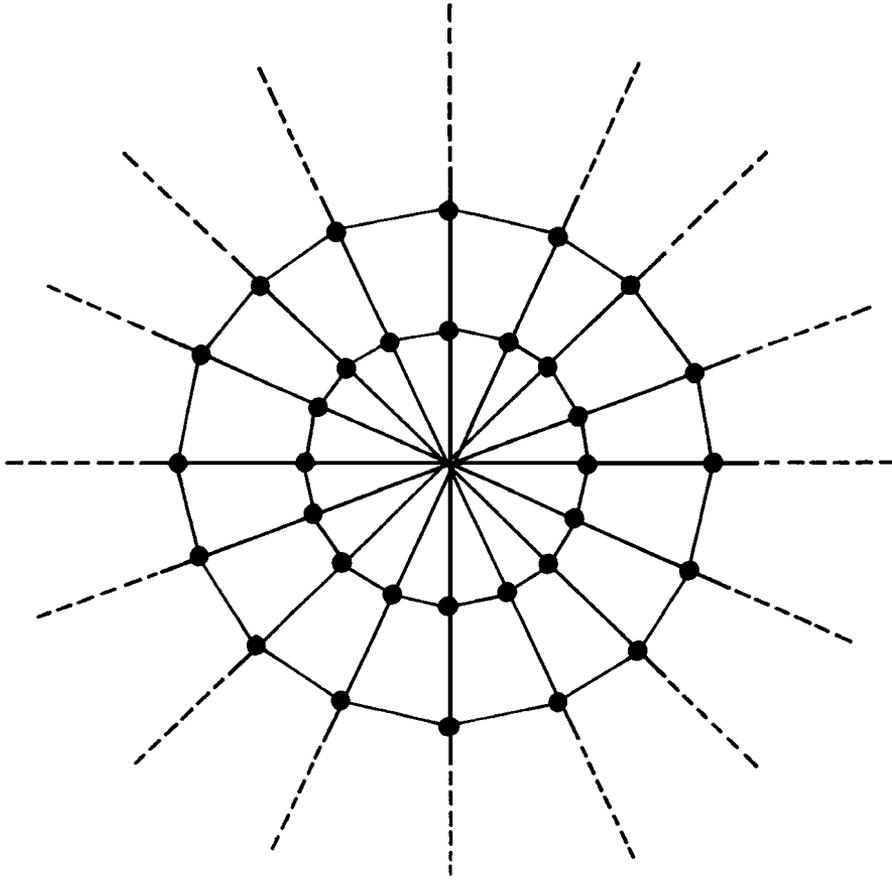


7-9A Comprehensive ground system for vertical antenna.

For vertical antennas mounted above ground, there is an optimum height above ground for the base of the antenna. This height is a quarter-wavelength above the actual ground plane. Unfortunately, that distance might not be the height above the surface. Depending upon ground conductivity and groundwater content, the height can be exactly a quarter-wavelength above the surface or slightly lower. The point is found from experimentation, and will, unfortunately, vary over the year if climatic changes are usual.

## Variations on the vertical antenna theme

Thus far, the vertical antennas have been standard quarter- or  $\frac{5}{8}$ -wavelength models. This section looks at several variations on the theme. Consider Fig. 7-10. This antenna is the vertical half-wavelength dipole. The vertical dipole is constructed in exactly the



7-9B “Spider web” ground system.

same manner as the horizontal dipole, but is mounted in the vertical plane. In general, the section of the radiator that is closest to the ground should be connected to the shield end of the coaxial cable transmission line.

Like the horizontal dipole, the approximate length of the vertical dipole is calculated from

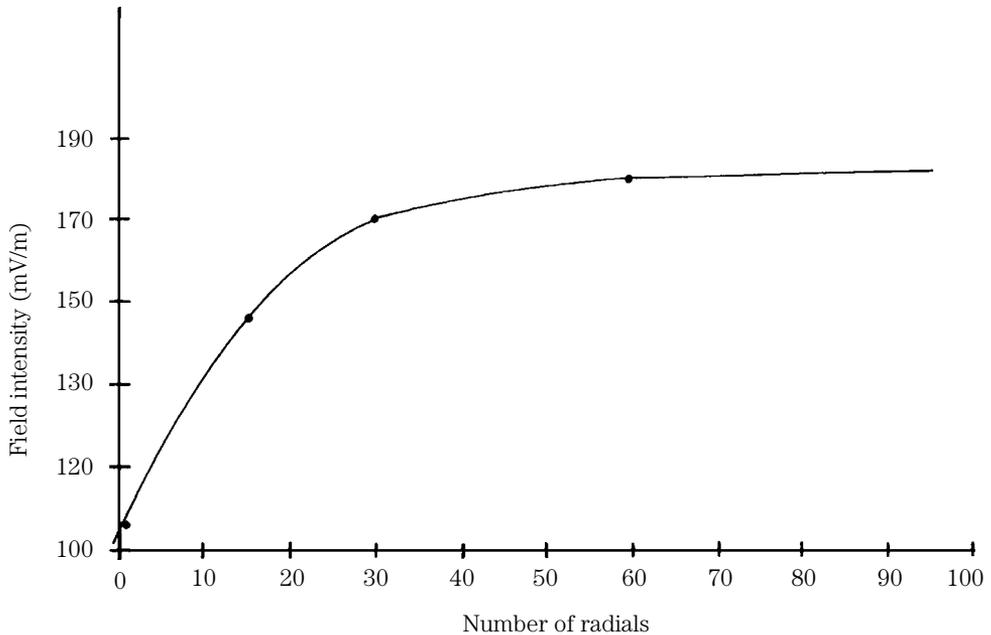
$$L_{\text{ft}} = \frac{468}{F_{\text{MHz}}} \quad [7.2]$$

where

$L_{\text{ft}}$  is the length, in feet

$F_{\text{MHz}}$  is the operating frequency, in megahertz

**Example** Calculate the length of a half-wavelength vertical dipole for operation on a frequency of 14.250 MHz in the 20-m amateur radio band.



7-9C Effect of increasing the number of radials on field strength.

**Solution:**

$$L_{ft} = \frac{468}{F_{MHz}}$$

$$L_{ft} = \frac{468}{14.250} = 32.8 \text{ ft}$$

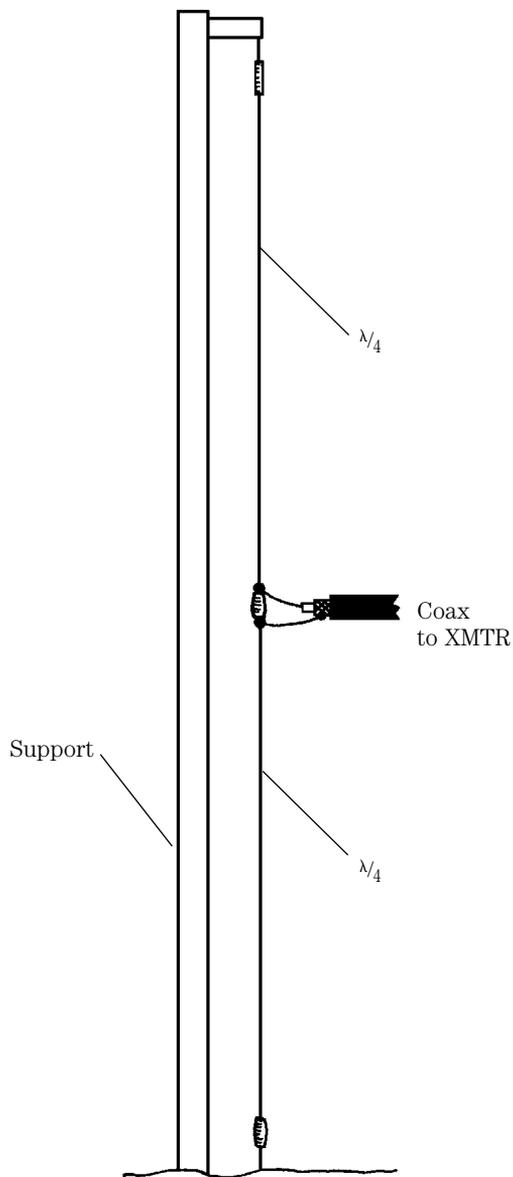
*Note:* The 0.8 ft part of this calculated length can be converted to inches by multiplying by 12:  $0.8 \times 12 = 9.6$  in.

Each leg of the vertical dipole is one half of the calculated length, or

$$\frac{32.8 \text{ ft}}{2} = 16.4 \text{ ft}$$

The vertical dipole antenna is used in many locations where it is impossible to properly mount a horizontal dipole, or where a roof- or mast-mounted antenna is impossible to install because of logistics, a hostile landlord, and/or a homeowners' association. Some row house and town house dwellers, for example, have been successful with the vertical dipole. In the 1950s and 1960s, the vertical dipole was popular among European amateurs because of space restrictions found in many of those locations.

The construction of the vertical dipole is relatively straightforward. You must find or build a vertical support structure. In the case shown in Fig. 7-10, the support



**7-10** Vertical half-wavelength dipole antenna.

is a wooden or PVC mast erected for that purpose. Ropes and insulators at either end support the wire elements from the ends and keep the antenna taut. If the neighbors are a problem, then try to find some white PVC pipe that will make a fine flagpole and be patriotic—with a vertical dipole hidden inside of the pipe. In other cases, if your home is not metal-sided, and if it is high enough, then a support from the roof structure (or soffits) will make a proper mount.

One problem with the vertical dipole, and one that liability-conscious people need to consider, is that a high-impedance current node is found at the ends of a half-wavelength dipole. Anyone touching the antenna will likely receive a nasty RF burn (or shock) from this antenna.

A *coaxial vertical* is similar to the vertical dipole (and, in fact, it can be argued that it is a form of vertical dipole) in that it uses a pair of vertical radiator elements. In the case of the coaxial vertical antenna, however, the radiator that is closest to the ground is coaxial with the transmission line and the main radiator element. An example is shown in Fig. 7-11A. An insulator at the feedpoint separates the two halves of the radiator; in most cases, it is of smaller diameter than the coaxial sleeve (also called the *shield pipe* in some publications). The reasons for this arrangement are not entirely electrical, for the most part, but mechanical. The coaxial cable transmission line passes through the sleeve, and is itself coaxial to the sleeve.

The overall length of the coaxial vertical antenna is a half-wavelength, which consists of two quarter-wavelength sections. Both the radiator and the sleeve are quarter-wavelength. The length of each is found (approximately) from

$$L_{\text{ft}} = \frac{234}{F_{\text{MHz}}} \quad [7.3]$$

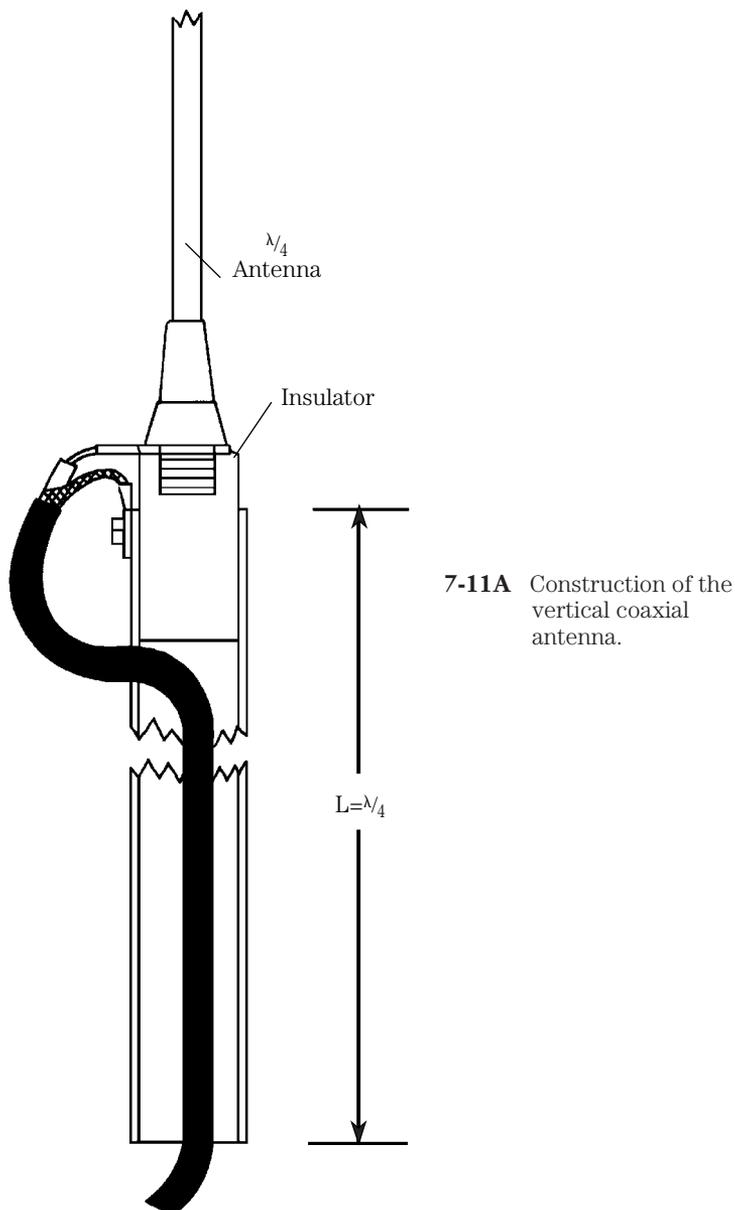
or

$$L_{\text{meters}} = \frac{72}{F_{\text{MHz}}} \quad [7.4]$$

You should recognize these as similar to the equation used previously to calculate half-wavelength antennas, but reduced by a factor of 2.

The coaxial vertical antenna was once popular with CB operators, and as such was called the *collinear antenna*. In some cases, you can find hardware from these antennas on the hamfest or surplus markets, and the pieces can be modified for amateur radio use. In the situation where a 10-m-band antenna is being built, it is a simple matter to cut the 11-m CB antenna for operation on a slightly higher frequency. In the case of the lower-frequency bands, however, it is a little more difficult and it is likely that only the insulator and mounting assembly are salvageable. Keep in mind, however, the fact that adjacent sizes of aluminum tubing are designed such that the inside diameter (ID) of the larger piece is a slip-fit for the outside diameter (OD) of the smaller piece. You can, therefore, connect adjacent sizes of aluminum tubing together without the need for special couplers, etc. With that in mind, salvaged insulator assemblies can be cut off with just 6 to 10 in of the former radiator and sleeve, and new radiators from “adjacent size” tubing installed.

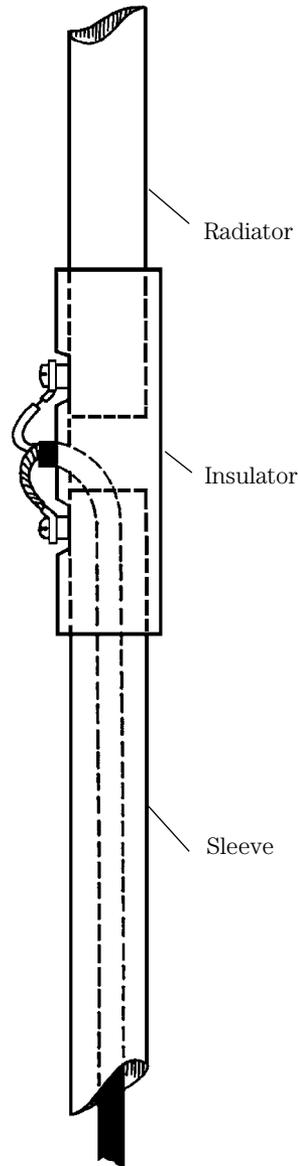
The configuration shown in Fig. 7-11A is the manner of construction used by commercial antenna manufacturers for VHF and CB collinear vertical dipoles, but is a little difficult for amateurs (unless they happen to own a machine shop) to make the center insulator. For those people, some other method is indicated before this antenna is practical.



**7-11A** Construction of the vertical coaxial antenna.

Figure 7-11B shows a construction method that has been used by amateurs with good results. The radiator and shield pipe (sleeve) are joined together in an insulating piece of thick-wall PVC plumbing pipe, Lucite, or plexiglass tubing; 6 to 10 in of tubing are needed.

A gap of about 2 in is left between the bottom end of the radiator pipe and the top end of the shield pipe in order to keep them electrically insulated from each



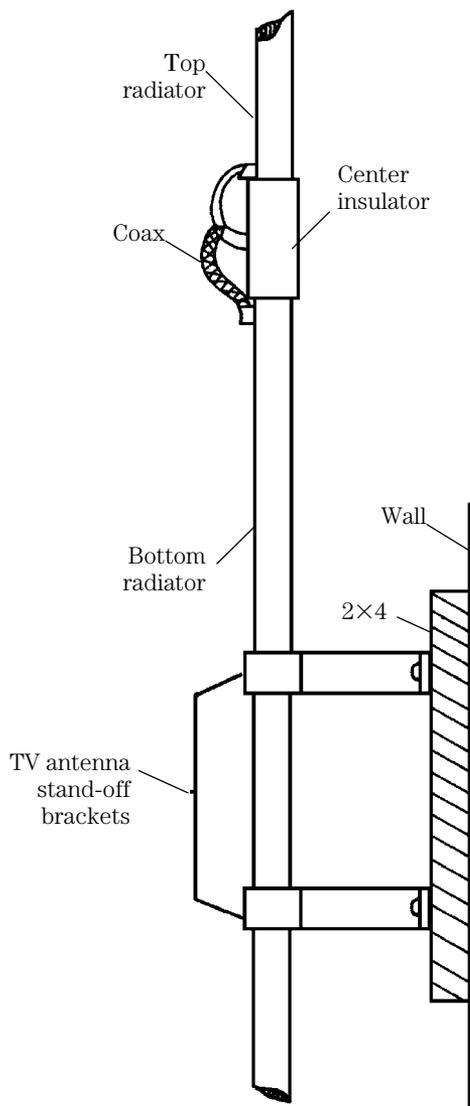
**7-11B** Vertical dipole made from tubing.

other, and to allow the coaxial cable to be passed through to the outside world. A hole in the insulator pipe is drilled for this purpose.

The aluminum tubing pieces for the radiator and the sleeve are fastened to the insulator with at least two heavy machine screws each. One of the machine screws on each can be used as the electrical connection between the coaxial cable and the pipes, provided that a larger hole is cut in the insulator at that point to admit the washer that provides the electrical pathway between the screw head and the aluminum pipe.

If you depend upon the machine screw touching the pipe at the edges of the hole cut for it, then there will probably be intermittent connection and all of the aggravation that ensues.

Mounting of the homebrew coaxial vertical antenna can be a “pain in the neck.” Normally, this antenna is mounted high in the air above ground, so some form of support is needed. Fortunately, you can use small-area metal supports connected to the sleeve for this purpose. Figure 7-11C shows one method for mounting that is popular. A pair of television antenna standoff mounting brackets are used to support the sleeve. Those brackets can be bought in sizes from 6 to 24 in. Note that a 2 × 4 piece of lumber is used between the building wall and the brackets. This wood serves as an



7-11C Mast mounting system for verticals.

insulator, so it should be varnished or painted. It is held to the wall with lag bolts, wing bolts, or some other effective method of anchoring. Keep in mind that the forces on the brackets increase tremendously during wind storms!

It is possible for the two vertical antennas shown previously to become a shock hazard to anyone who touches them. Both of these antennas are half-wavelength radiators and are of the dipole form of construction. The center point is used for feeding the antenna, so it forms the low-impedance point in the antenna. As a result, the ends of the antenna, one of which is close to the ground, are the high-impedance points—hence the voltages at those points within reach of prying hands playing in the yard can be high. It is wise to either mount the antennas so far above the ground that they cannot be reached, or build a small nonconductive fence around the bottom end of the antenna.

## Vertical antenna construction

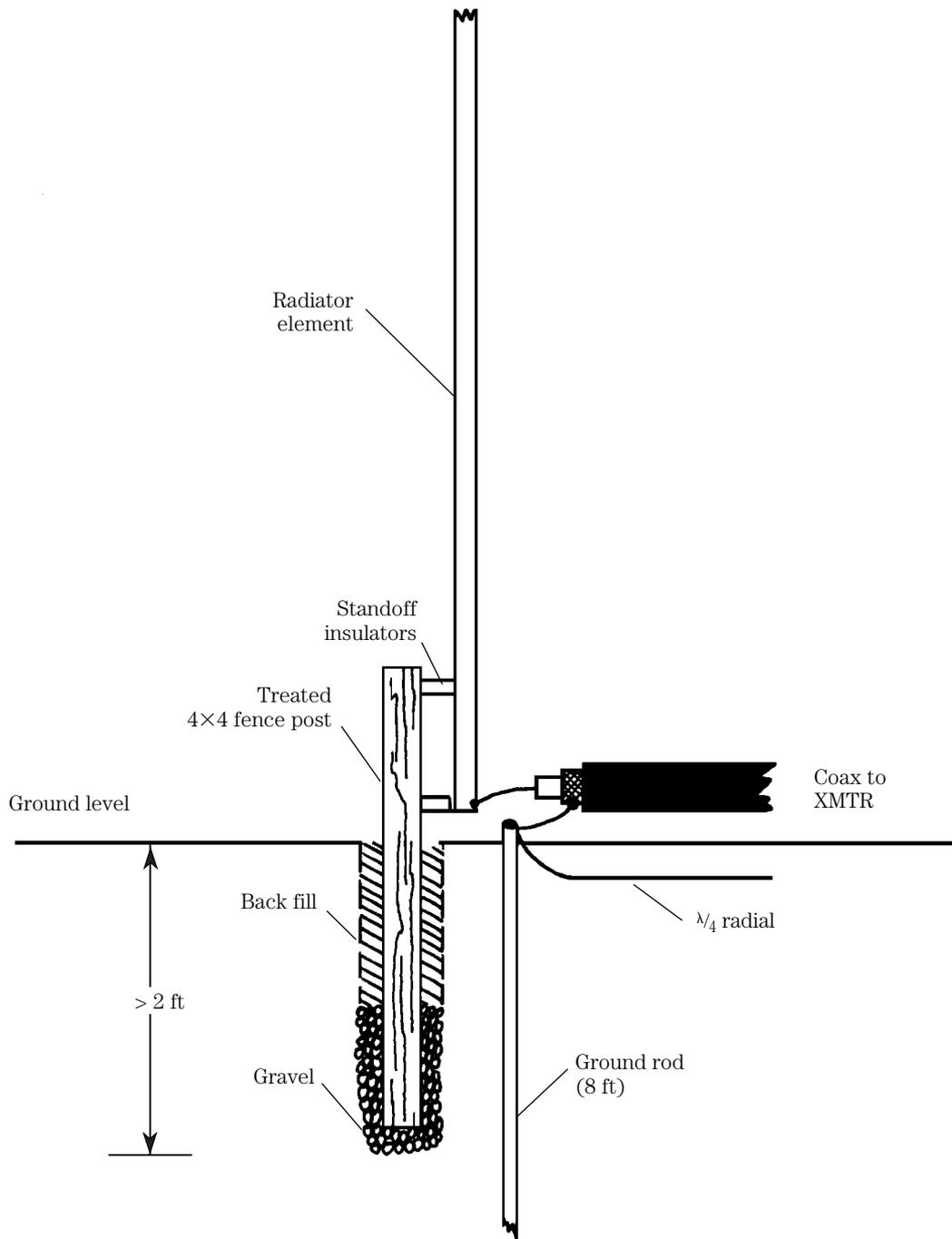
There are two general cases for installing vertical antennas: ground-level-mounted and non-ground-mounted. This section takes a brief look at both forms of mounting. We will concentrate on installation of homebrew verticals rather than commercial, because it is assumed that the vendors of such antennas will provide their own instructions.

The ground-level-mounted vertical is shown in Fig. 7-12. The typical vertical antenna is 8 to 40 ft high. Thus, although the actual weight of the antenna is small, the forces applied to the mounting structure (especially during windstorms) can be quite high. Don't be fooled by the apparent light weight of the antenna in this respect.

The mounting structure for the vertical antenna can be a metal or wooden fence post buried in the ground. At least 2 ft of the fence post should be above ground. In the case of Fig. 7-12, a 4 × 4 wooden fence post is used as the mounting, but the principles are similar for all forms of post. A fence post hole is provided that is at least 2 ft deep. In some cases, it might be possible to use a 1-ft-deep gravel fill topped with back-filled dirt. In other cases, especially where a steel fence post is used, a concrete plug is placed at the bottom of the hole over a 4-in layer of gravel.

The antenna radiator element is installed onto the fence post using standoff insulators. Unfortunately, these insulators are difficult to find, so they might have to be omitted. Given that varnished or painted wood is not a terribly good conductor, it is not unreasonable to bolt the radiator directly to the 4 × 4 fence post. Use  $\frac{5}{16}$ -in (or larger) bolts, and make them long enough to fit completely through both the antenna element and the 4 × 4 post. Thus, 6-in-, 7-in-, or 8-in-long  $\frac{5}{16}$ -in bolts are the candidates for this job. Use at least two bolts, one at the bottom of the antenna radiator element, and one near the top of the fence post. A third bolt, halfway between the other two, would not be out of order.

If the antenna is quarter-wavelength, then no matching is generally necessary. Although the feedpoint impedance is not exactly 52  $\Omega$ , it is close enough (37  $\Omega$ ) to form a reasonable match for 52- $\Omega$  coaxial cable (with VSWR = 1.4:1). The center conductor of the coaxial cable is connected to the radiator element, while the shield is connected to the ground system. In the example shown in Fig. 7-12, two ground methods are used. First, an 8-ft ground rod is driven into the earth at the base of the



7-12 Ground-mounted vertical.

antenna; second, a system of quarter-wavelength radials is used. Remember that the ground system is absolutely essential.

A method for installing a vertical antenna above ground is shown in Fig. 7-13. In this case, a wooden support ( $2 \times 4$  or  $4 \times 4$ ) is installed in a manner similar to Fig. 7-12, but with a deeper hole to counter the longer length. Alternatively, the wooden support is affixed to the side of a building wall, shed, or other preexisting structure. Once the support is arranged, however, the method of attachment of the radiator element is the same for the previous case, so that will not be repeated here.

Electrical connections to the antenna are also shown in Fig. 7-13. Because the antenna is above ground level, an electrical counterpoise ground consisting of a system of radials is absolutely essential; at least two radials per band must be provided. A small L bracket is used to support the radials and to provide an SO-239 coaxial connector for the coax. This connector is a chassis-mounted type with its center conductor connected to the radiator element. The shield of the connector is connected to the bracket, so it is also connected to the radial system.

In some installations, the antenna support structure will require guy wires to keep the structure stable. Do not use the radials as guy wires. The type of wire that normally works well for radials is too soft, and too easily stretched, for guy wire service. Use regular steel guy line, available where TV antenna supplies are sold, for this antenna. Make the lengths nonresonant, and break the guy lines up with egg insulators, if necessary, to achieve the nonresonance.

## 5/8-Wavelength verticals

Figure 7-14 shows the configuration for the 5/8-wavelength vertical antenna. Such an antenna generally gives a lower angle of radiation than the more common quarter-wavelength radiator, so presumably it works better for long distance.

The radiator of this antenna is made from 0.5-in to 1.5-in aluminum tubing. Again, remember that adjacent sizes fit together snugly to form longer sections. The physical length of the 5/8-wavelength radiator is found from

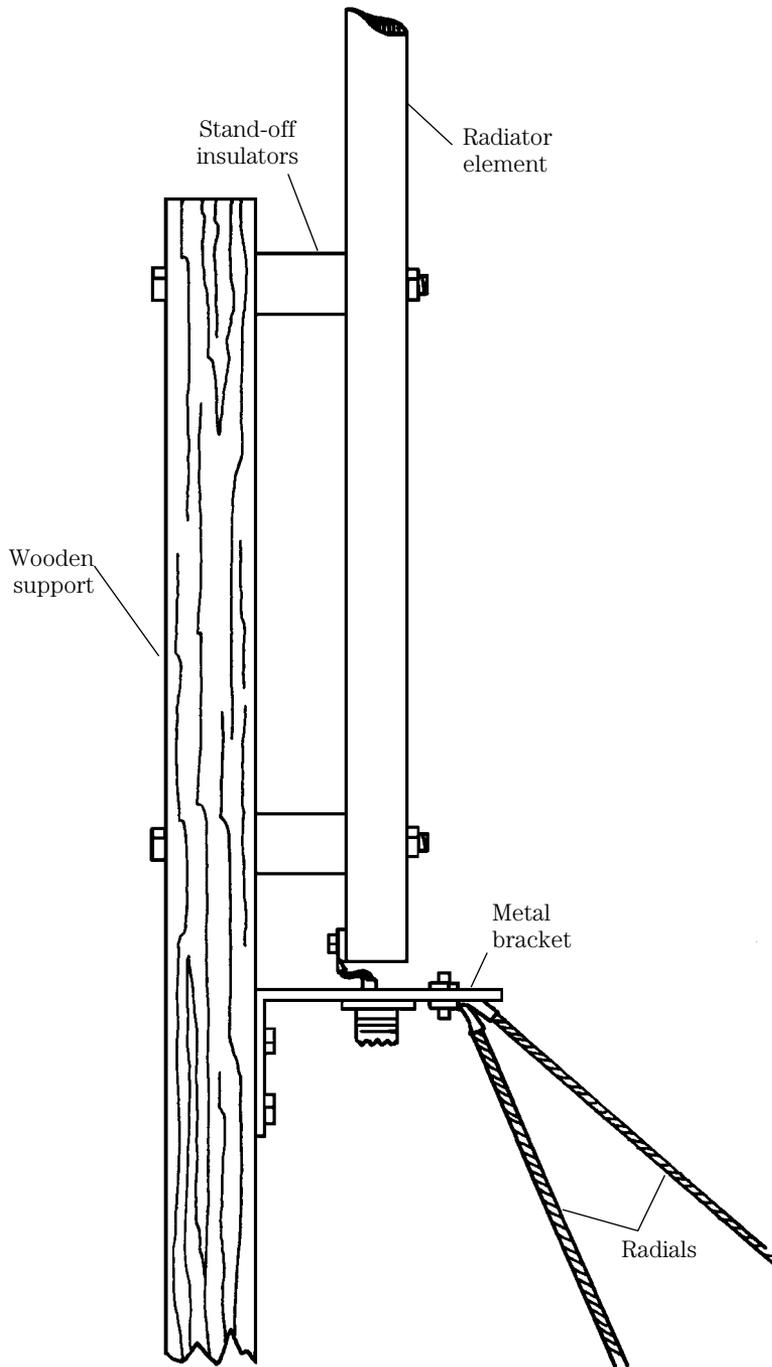
$$L_{\text{ft}} = \frac{585}{F_{\text{MHz}}} \quad [7.5]$$

or, in meters,

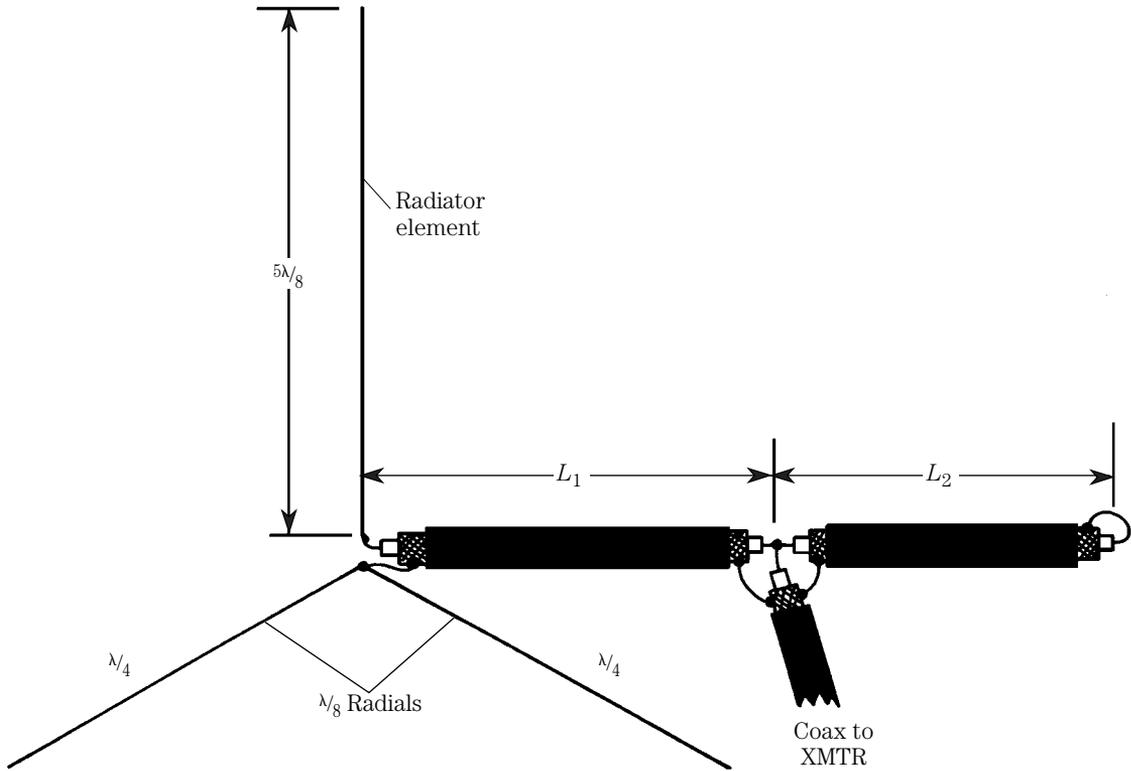
$$L_{\text{meters}} = \frac{180}{F_{\text{MHz}}} \quad [7.6]$$

The radials are the usual quarter-wavelength, and are made of no. 12 or no. 14 copper wire. These lengths are found from:

$$L_{\text{ft}} = \frac{246}{F_{\text{MHz}}} \quad [7.7]$$



7-13 Details of feedpoint circuit on vertical.



**7-14** Using a Q section to match the feedpoint impedance of a vertical antenna.

or, in meters,

$$L_{\text{meters}} = \frac{75}{F_{\text{MHz}}} \quad [7.8]$$

The feedpoint impedance of the  $\frac{5}{8}$ -wavelength antenna is about  $1600 \Omega$ , not a good match for the ordinary coaxial cables that are routinely available on the amateur market. Some form of impedance matching is needed.

One option is to use a broadbanded RF transformer. These transformers will work throughout the HF spectrum, and match a wide variety of impedances to the  $50\text{-}\Omega$  standard system impedance.

Another option, especially for a single-band antenna, is to use a coaxial cable impedance transformer, such as shown in Fig. 7-14. The transformer consists of two sections of coaxial cable joined together, shown as  $L_1$  and  $L_2$  in Fig. 7-14. The lengths are found from

$$L_1 = \frac{122}{F_{\text{MHz}}} \text{ ft} \quad [7.9]$$

or

$$L_1 = \frac{38}{F_{\text{MHz}}} \text{ m} \quad [7.10]$$

and

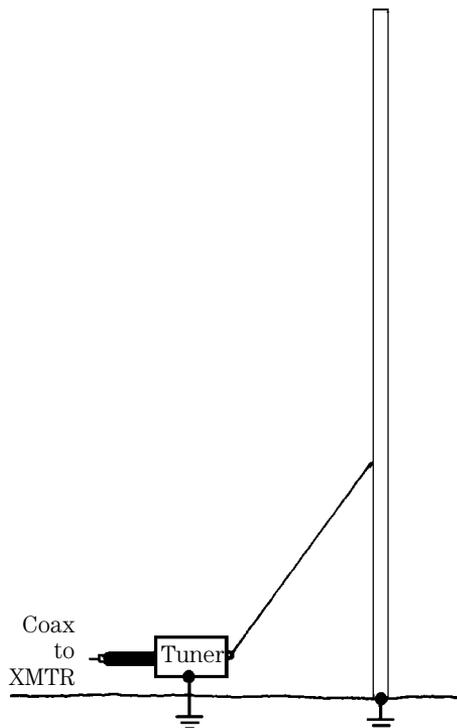
$$L_2 = \frac{30}{F_{\text{MHz}}} \text{ ft} \quad [7.11]$$

$$L_2 = \frac{9}{F_{\text{MHz}}} \text{ m} \quad [7.12]$$

## Grounded vertical antennas

The vertical antennas presented thus far in this chapter are called *series-fed verticals*, because the generator is essentially in series with the radiator element. Such an antenna must be insulated from ground. The other class of vertical is the *shunt-fed vertical*, which is grounded at one end (Fig. 7-15). There are three methods of shunt-feeding a grounded vertical antenna: *delta*, *gamma*, and *omega*. All three matching systems have exactly the same function: to form an impedance transformation between the antenna radiation resistance, at the feedpoint, and the coaxial cable characteristic impedance, as well as cancelling any reactance in the system.

The delta feed system is shown in Fig. 7-15A. In this case, a taut feed wire is connected between a point on the antenna, which represents a specific impedance on



7-15A Delta-fed grounded vertical.

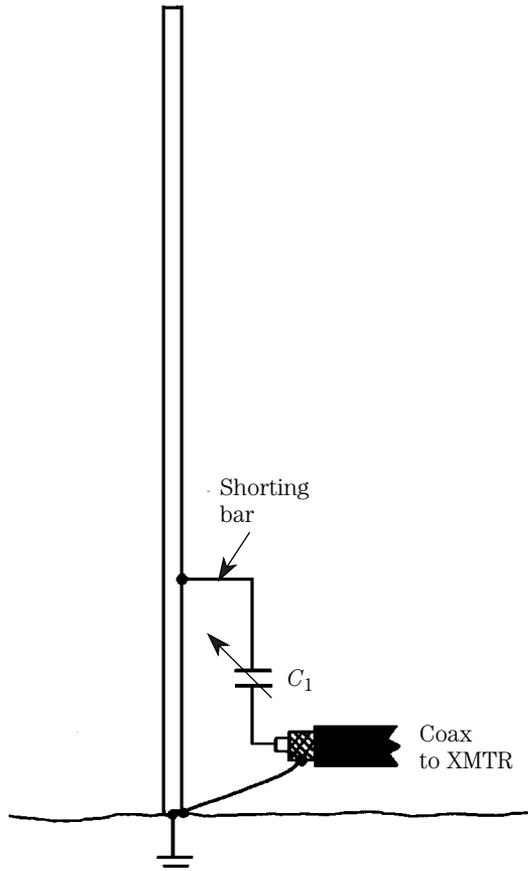
the antenna, and an antenna tuner. This method of feed is common on AM broadcast antennas (which are usually—perhaps always—verticals). Although you would think that the sloping feed wire would distort the pattern, that is not the case. The distortion of the pattern, if any, is minimal, hence it can be neglected.

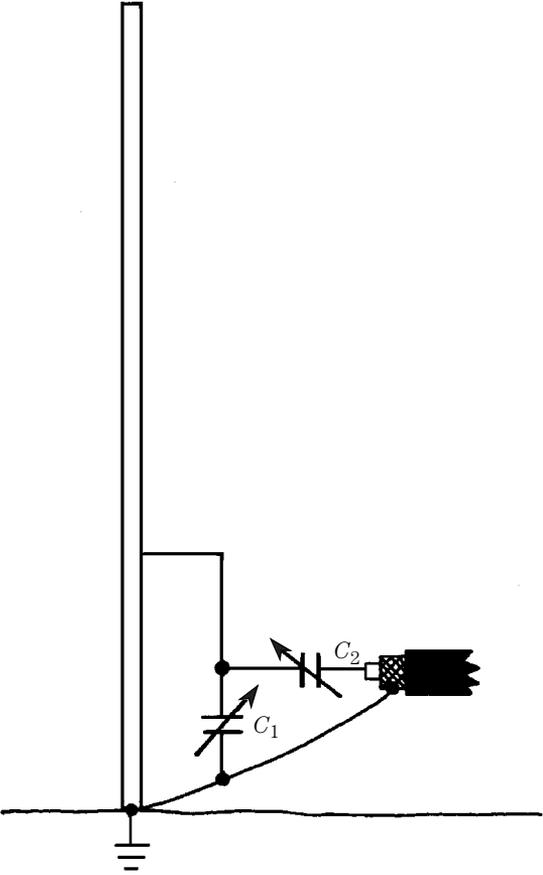
The gamma feed system is shown in Fig. 7-15B. This method is commonly used by amateurs to feed Yagi beam antennas, so it is quite familiar in the amateur radio world. The feed system consists of a variable capacitor to tune the system, and a matching rod that parallels the antenna radiator element. It is important that the rod not be anywhere near a quarter-wavelength, or it would become a vertical antenna in its own right, and in fact would resemble the so-called *J-pole antenna*. A review of the gamma match is given in Chap. 12. The omega feed (shown in Fig. 7-15C) is similar to the gamma match except that a shunt capacitor is used.

## Conclusion

The vertical antenna is a viable alternative for many situations, especially where real estate is at a premium. Contrary to popular opinion, the vertical antenna works well when installed properly and when due consideration has been given to matters such as the grounding and angle of radiation desired.

7-15B Gamma-fed grounded vertical.





7-15C Omega-fed grounded vertical.

## 8 CHAPTER

# Multiband and tunable-wire antennas

MOST COMMUNICATIONS OPERATORS REQUIRE MORE THAN ONE BAND, AND THAT MAKES the antenna problem exactly that—a problem to be solved. Amateur radio, commercial, and military operators are especially likely to need either multiple antennas for different bands, or a multiband antenna that operates on any number of different bands. This situation is especially likely on the high-frequency (HF) bands from 3.5 to 29.7 MHz.

Another problem regards the tunability of an antenna. Some amateur bands are very wide (several hundred kilohertz), and that causes any antenna to be highly variable from one end of the band to another. It is typical for amateurs to design an antenna for the portion of the band that they use most often, and then tolerate a high VSWR at the other frequencies. Unfortunately, when you see an antenna that seems to offer a low VSWR over such a wide range, it is almost certain that some problem exists that reduces the  $Q$ , and the antenna efficiency, to broaden the response. However, it is possible to tune an antenna for a wide band. It is also possible (now that amateurs have new HF bands) to use a single antenna between them, and then tune the difference out. For example, designing a single antenna for 21/24 MHz, 14/18 MHz, or 7/10 MHz should prove possible.

In this chapter we will take a look at both problems: the multiband and the tunable antenna.

## Multiband antennas

Although a triband Yagi or quad beam antenna will undoubtedly work better than a wire antenna (when installed correctly!), the low-budget amateur operator need not lament any supposed inability to “get out” on wire antennas. To quote an old saying: “‘Better’ is the enemy of ‘good enough.’” Or, to put it in terms of *Carr’s law*: “If it’s ‘good enough’ then don’t waste a lot of energy fretting over making it ‘better’ unless you really want to make it *a lot better*.”

## Trap dipoles

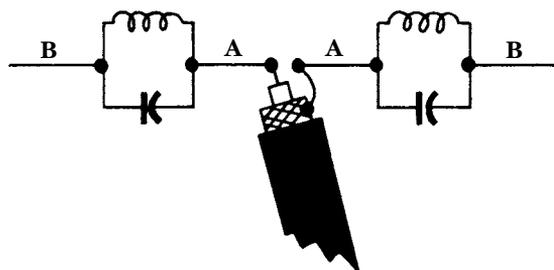
Perhaps the most common form of multiband wire antenna is the *trap dipole* shown in Fig. 8-1A. In this type of antenna, one (or more) pairs of parallel resonant traps are placed in series with the quarter-wavelength elements of the dipole. The purpose of the traps is to block their own resonant frequency, while passing all other frequencies.

In the example of Fig. 8-1A, a 10-m trap isolates the first 8 ft or so (quarter-wavelength on 10 m) so that the antenna resonates on that band. A 40-m or 15-m signal, on the other hand, passes through the traps and uses the whole length of the antenna. (*Note:* A half-wavelength 40-m dipole works as a  $3/2$ -wavelength antenna on 15 m.)

The overall length of the trap dipole will be a little less than the natural “non-trap” length for the lowest frequency of operation. At the low frequencies, the traps add a little inductance to the circuit so that the resonant point is lower than the natural resonant frequency. In general, most trap dipoles are just a few percent shorter than nontrap dipoles at the same band. The actual amount of shortening depends upon the values of the components in the traps, so consult the data for each trap purchased. Where more than one pair of traps is used in the antenna, make sure they are of the same brand and are intended to work together.

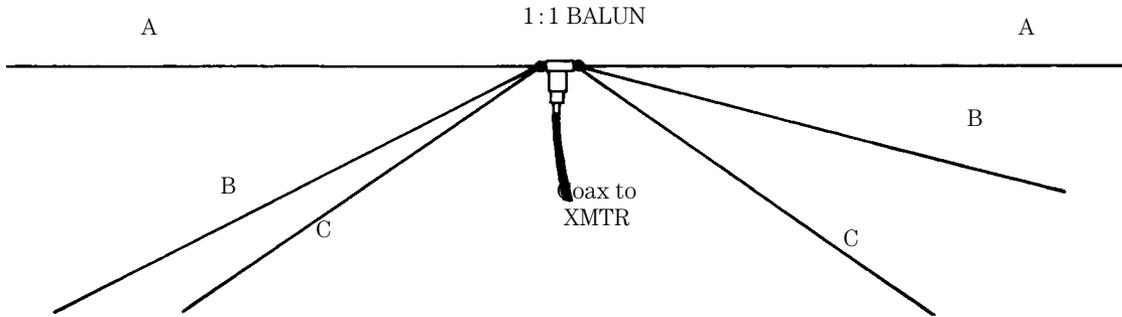
Another solution to this problem is shown in Fig. 8-1B. This type of antenna actually has two or more half-wavelength dipoles fed from the same transmission line. In this illustration, a total of three dipoles are fed from the same  $75\text{-}\Omega$  transmission line. There is no theoretical limit to how many dipoles can be accommodated, although there is certainly a practical limit. For one thing, there is a mechanical limit to how many wires are supportable (or desirable) hanging from any given support. There is also an electrical limit, although it is less defined. Having a lot of dipoles increases the possibility of radiating harmonics and other spurious emissions from your transmitter.

The  $75\text{-}\Omega$  coaxial cable is connected to the center feedpoint of the multidipole either directly or through a 1:1 balun transformer, as shown in Fig. 8-1B. Each an-



A=10-meter segment  
A+B=15-meter/40-meter segment

**8-1A** Trap dipole for multiband operation. (Courtesy of Hands-On Electronics and Popular Electronics)



**8-1B** Multiband dipole consists of several dipoles fed from a common feedline.

tenna (*A-A*, *B-B*, or *C-C*) is a half-wavelength. Therefore, the overall length is found approximately from the standard dipole expressions:

Overall length (*A + A*, *B + B*, or *C + C*):

$$L_{ft} = \frac{468}{F_{\text{MHz}}} \quad [8.1]$$

or, for each element alone (*A*, *B*, or *C*):

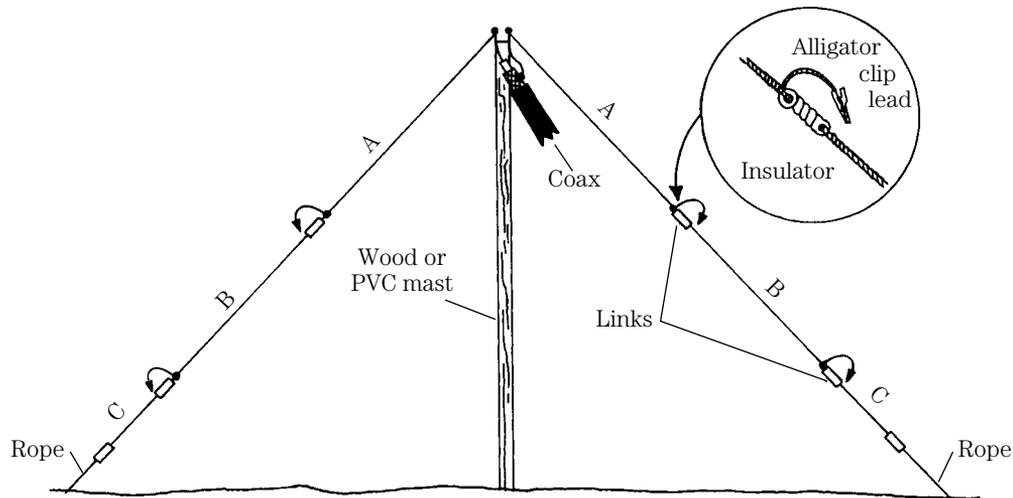
$$L_{ft} = \frac{234}{F_{\text{MHz}}} \quad [8.2]$$

As always, close to the earth's surface, these equations are approximations, and are not to be taken too literally. Some experimentation will probably be necessary to optimize resonance on each band. Also, be aware that the drooping dipoles (*B* and *C* in this case) may act more like an inverted-vee antenna (see Chap. 7) than a straight dipole, so the equation length will be just a few percent too short. In any event, a little spritzing with this antenna will yield results.

Some amateurs build the multiple dipole from four- or five-wire TV rotator lead. That type of wire is used to control antenna rotators, and has either four or five parallel wires in a flat arrangement similar to lamp cord. Cut each wire to the length required for a band, and strip off any unused portions.

Another possibility is the link-tuned dipole shown in Fig. 8-1C. In this situation, a single conductor is used for each half of the dipole, or actually inverted vee. The conductors are broken into segments *A*, *B*, and *C* (or more, if desired). Each segment is separated from the two adjacent sections by inline insulators (standard end insulators are suitable). Segment *A* is a quarter-wavelength on the highest frequency band of operation, *A + B* is a quarter-wavelength on the next highest band of operation, and *A + B + C* is a quarter-wavelength on the lowest frequency band of operation.

The antenna is "tuned" to a specific band by either connecting, or disconnecting, a similar wire (see inset) jumper across the insulator that breaks the connection between the segments. Either a switch or an alligator clip jumper will short out the insulator to effectively lengthen the antenna for a lower band. Some amateurs use single-pole 110-Vac power line switches to jumper the insulator. Although I have not tried this method, it should work.



**8-1C** Multiband inverted vee uses shorting links to change bands.

A big disadvantage to this type of multiband antenna is that you must go out into the yard and manually switch the links to change bands, which probably explains why other antennas are a lot more popular, especially in northern latitudes.

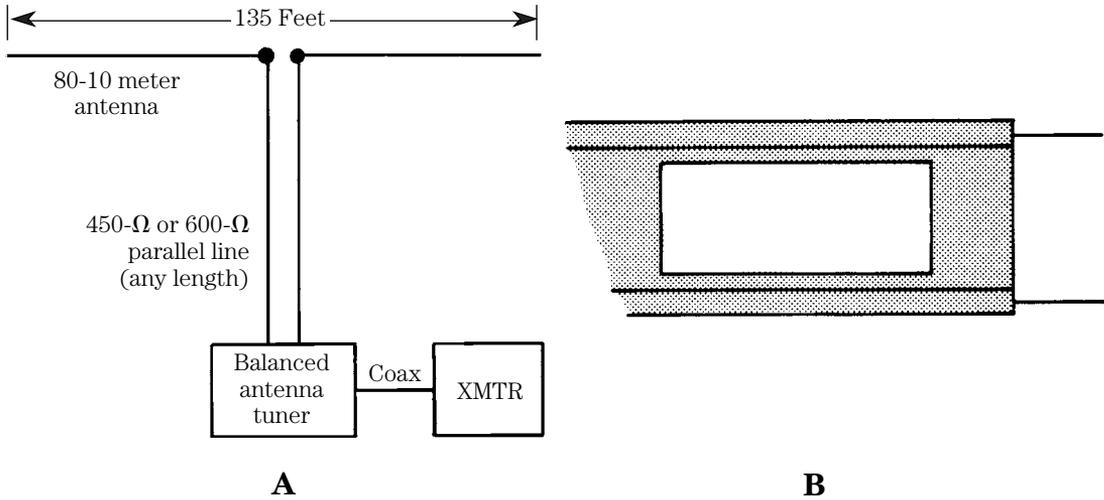
## Tuned feeder antennas

Figure 8-2A shows the tuned feeder type of antenna. This antenna can be used from 80 through 10 m, but it requires a special tuner and a length of parallel transmission line. There are two ways to get parallel transmission line: make it or buy it. Using no. 14 or 12 wire, and specially made insulators (also called *spreaders*), you can make 300-, 450-, or 600- $\Omega$  parallel transmission line. But that's a pain in the *ptusch* because you can also buy parallel line rather cheaply. I paid \$16 for 100 ft of 450- $\Omega$  line recently.

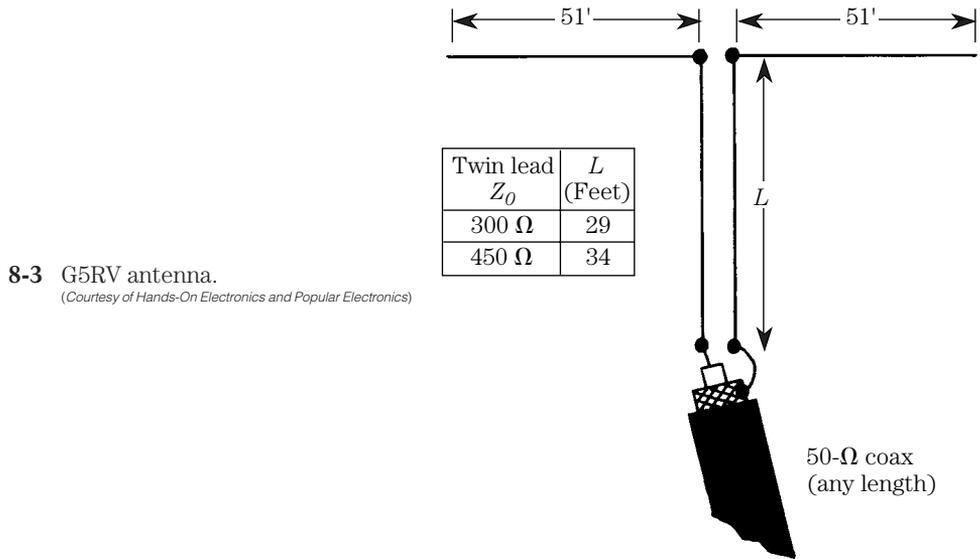
One form of parallel line is ordinary TV-type twin lead, which has an impedance of 300  $\Omega$ . This line will take up to about 250 W, although some people use it at higher powers (not recommended!). The antenna of Fig. 8-2A uses 450- $\Omega$  parallel line. You can buy insulated 450- $\Omega$  twin lead (see Fig. 8-2B) that can be handled as easily as TV twin lead—and a lot more easily than open (uninsulated) parallel line.

## The G5RV multiband dipole

Figure 8-3 shows the popular G5RV antenna. Although not without some problems, this antenna is very popular. It can be used either as a horizontal dipole, a sloper, or an inverted-vee antenna (which is how I used it). The dipole elements are each 51 ft long. The feedline can be either 300- or 450- $\Omega$  twin lead. For 300- $\Omega$  cases, use 29 ft of line, and for 450- $\Omega$  line, use 34 ft. One end of the parallel transmission line is connected to the antenna, and the other end is connected to a length of 50- $\Omega$  coaxial



8-2 Tuned feeder antenna can be used on several bands. (Courtesy of Hands-On Electronics and Popular Electronics)



8-3 G5RV antenna. (Courtesy of Hands-On Electronics and Popular Electronics)

cable. Although most articles on the G5RV claim that any length of 50- $\Omega$  line will work, J. M. Haerle (*HF Antenna Systems: The Easy Way*) recommends that the 50- $\Omega$  segment be at least 65 ft long.

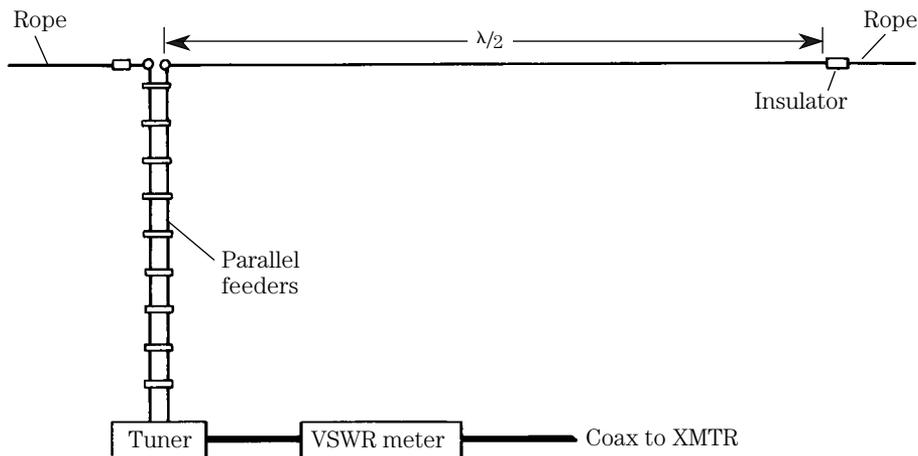
Haerle is a little caustic in his comments on the G5RV, but his criticism is well taken. If you don't have a parallel transmatch, then the G5RV will work (especially if your rig can tolerate a 3:1 VSWR on some selected frequencies). Otherwise, use the antenna of Fig. 8-2A (or a transmatch with the G5RV).

Figure 8-4 shows the once-popular end-fed Zepp antenna. This antenna is a “monopole” in that it uses a half-wavelength radiator, but it is fed at a voltage loop rather than a current loop (i.e., the end of the antenna rather than the center). A 450- or 600- $\Omega$  parallel transmission line is used to feed the Zepp antenna. Although the line can theoretically be any length, practicality dictates a quarter-wavelength because of the fact that the antenna is fed at a high RF voltage point.

The Zepp (as shown) is a single-band antenna unless the transmission line is fed with a good, widerange transmatch, or other antenna tuner unit. The antenna tuner required is a balanced type, although a standard transmatch with a 4:1 balun transformer at the output will also work well. Using the Zepp on many bands is easy, but keep in mind that it is voltage-fed, and (at frequencies for which the parallel line is not an odd integer multiple of a quarter-wavelength) there will be a high-voltage node at the transmatch. That raises the possibility of “RF in the shack”—“hot” snip-pets of RF on the grounded chassis of station equipment (including microphones, where it is uncomfortable when touched to the lips).

## Feeding parallel transmission line

Parallel transmission line is *balanced* with respect to ground, but coaxial cable is *unbalanced* to ground (i.e., one side of the coax is usually grounded). As a result, the standard amateur transmitter output will not drive parallel (balanced) feedline properly. You need to do one of two things: (1) buy (or build) an antenna tuner that is balanced on the output and unbalanced on the input or (2) convert a standard “coax-to-coax” transmatch or other form of antenna coupler to “coax-to-balanced” configuration. In some cases you can use a 4:1 balun transformer at the output of the coaxial cable tuner. Keep in mind, however, that some tuner manufacturers recommend against this practice. If you use the standard tee-network (or SPC) transmatch



8-4 End-fed Zepp antenna.

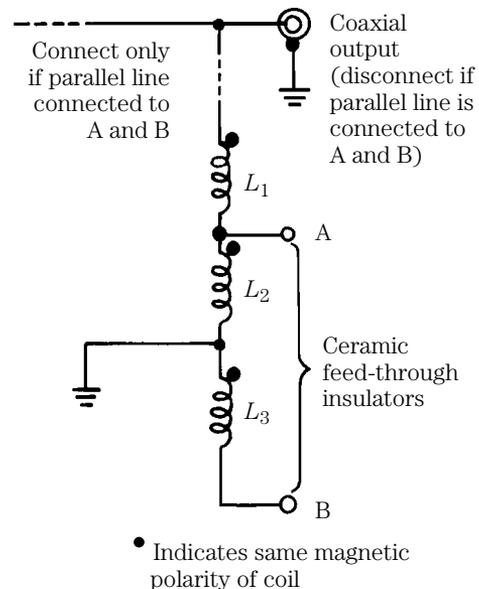
then it is possible to make a balun for this purpose from some no.12 formvar- or enamel-covered wire and either a ferrite toroid or a short piece of PVC plumbing pipe. Figure 8-5 shows a balun coil construction project. Use 12 turns of no.12 formvar or enameled wire over a 1-in outside diameter piece of PVC pipe or tubing.

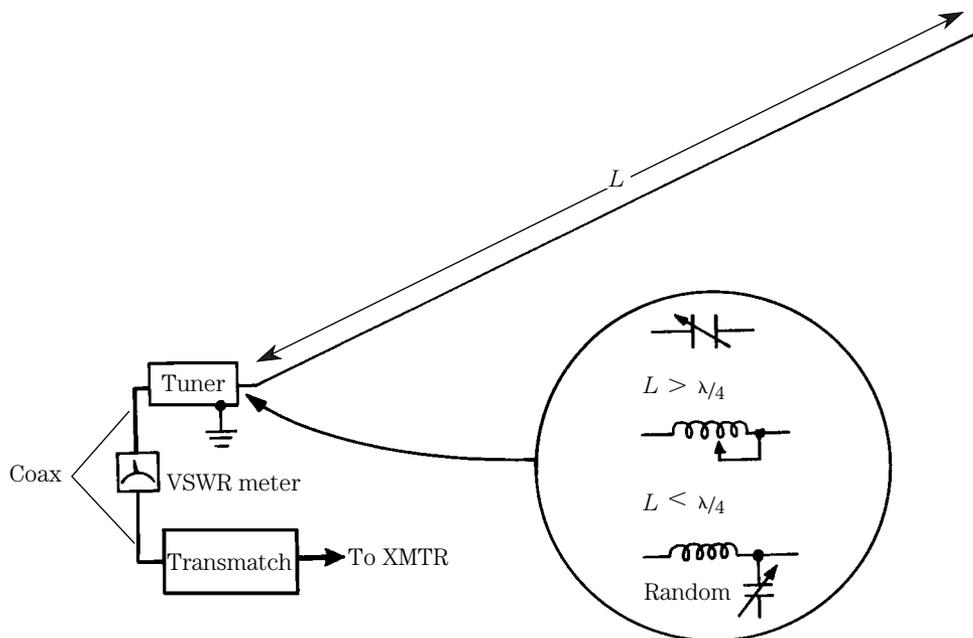
Next, we have the so-called longwire antenna. I use the term *so-called* because not all long wire antennas are truly longwires. A true longwire is longer than two wavelengths at the lowest frequency of operation. In Fig. 8-6 we see a longwire, or “random-length,” antenna fed from a tuning unit. If the antenna length  $L$  is greater than a quarter-wavelength, then the tuner consists of a single series capacitor (see below); and if it is shorter than a quarter-wavelength, the tuner is a series inductor. The standard tuner for this type of antenna, regardless of length, is a simple L-section coupler (also shown), which can be selected for L-section, series-L or series-C operation.

The Windom antenna (Fig. 8-7) has been popular since the 1920s. Although Loren Windom is credited with the design, there were actually a number of contributors. Coworkers with Windom at the University of Illinois were John Byrne, E. F. Brooke, and W. L. Everett, and they are properly cocredited. The designation of Windom as the inventor was probably due to the publication of the idea (credited to Windom) in the July 1926 issue of *QST* magazine. Additional (later) contributions were rendered by G2BI and GM1IAA (Jim MacIntosh). We will continue the tradition of crediting Loren Windom, with the understanding that others also contributed to this antenna design.

The Windom is a roughly half-wavelength antenna that will also work on even harmonics of the fundamental frequency. The basic premise is that the antenna radiation resistance varies from about  $50\ \Omega$  to about  $5000\ \Omega$ , depending upon the

**8-5** Balun transformer converts coaxial line to parallel line feed system.





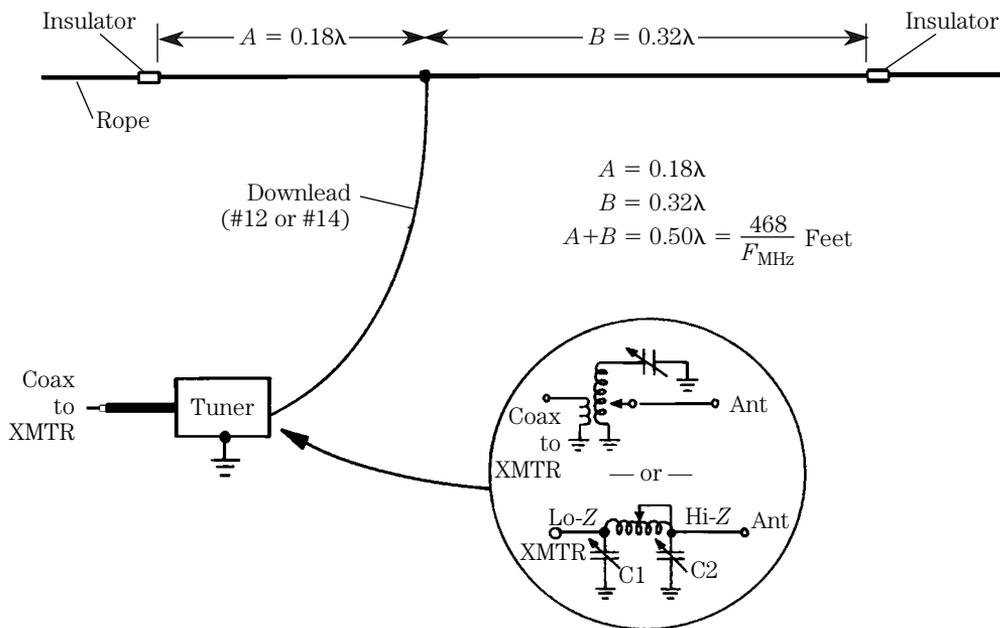
8-6 Random-length (a.k.a., if erroneously, “longwire”) antenna.

selected feedpoint. When fed in the exact center, a current loop, the feedpoint impedance will be  $50\ \Omega$ ; similarly, end feeding the antenna finds a feedpoint impedance of about  $5000\ \Omega$ . In Fig. 8-7A the feedpoint is tapped away from the center at a point that is about one-third the way from one end, at a point where the impedance is about  $600\ \Omega$ .

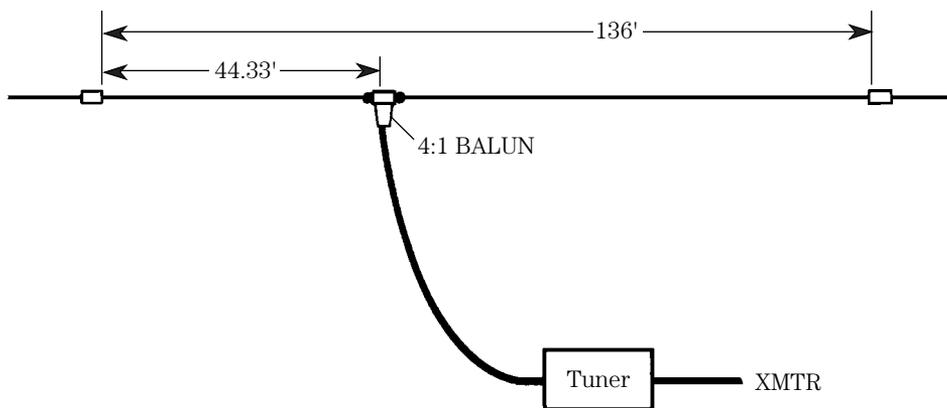
The Windom antenna works well—but with some caveats. For example, the antenna has a tendency to put “RF in the shack” because of the fact that it is voltage-fed. Second, there is some radiation loss from the feedline. Finally, the antenna works poorly on odd harmonics of the fundamental frequency.

The antenna tuning unit can be either a parallel resonant, link-coupled  $LC$  tank circuit (see inset to Fig. 8-7A); or a reversed pi network. In the case of the Windom, the pi network is turned around backward from the usual configuration:  $C_1$  is at the low-impedance end of the network, so it is larger than  $C_2$ . Design a pi network (see programs at end of book) to match  $50\ \Omega$  on the transmitter end and  $600\ \Omega$  on the antenna end.

A reasonable compromise Windom that reduces feedline radiation losses is shown in Fig. 8-7B. In this antenna a 4:1 balun transformer is placed at the feed point, and this in turn is connected to  $75\text{-}\Omega$  coaxial transmission line to the transmitter. A transmatch, or similar antenna tuner, is then connected between the transmitter and the transmission line.



8-7A Windom antenna.



8-7B Coaxial-fed Windom.

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## 9 CHAPTER

# Longwire directional antennas

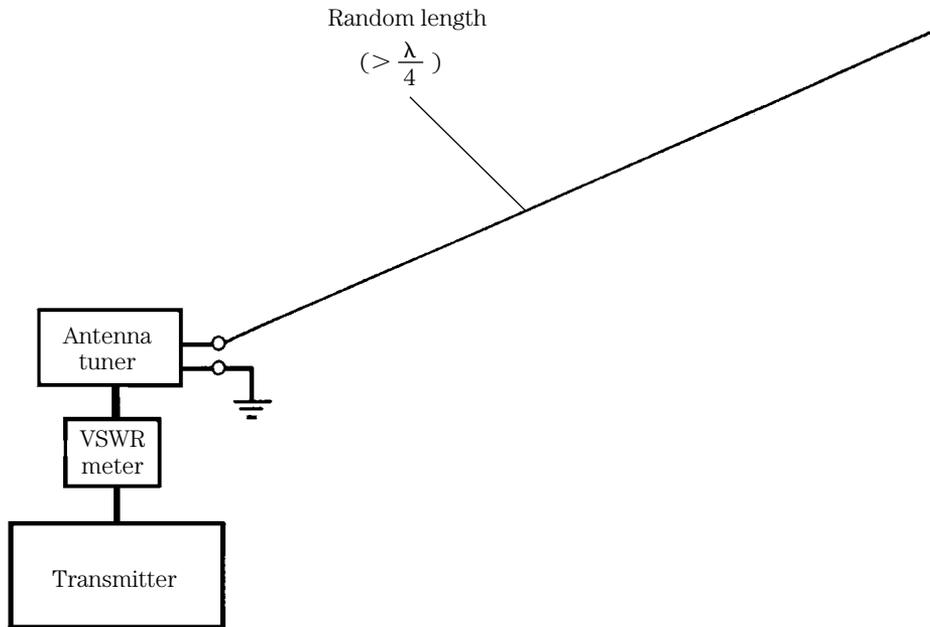
YOU'VE HEARD IT DOZENS OF TIMES: THE RANDOM-LENGTH LONGWIRE ANTENNA IS THE “perfect” solution to awkward antenna problems. Whether it’s a lack of real estate, cranky landlords, or a profound lack of dollars, the random-length longwire will do the job for you. Right? Well, now, that depends on who you ask, and what they did to make it work— or not work. One person says the longwire is not worth a plugged nickel; another is very ho-hum about it because his “kinda works”; still another is enthusiastic because hers is installed correctly and it works better than anything “since sliced pickle and liverwurst sandwiches.” Over the years my various living arrangements have forced me to use longwires at many QTHs, as well as on Field Day. Why does the random-length longwire have such a varied reputation? Before we answer that question let’s find out (for those who came in late) just what is a longwire antenna.

As an aside, let me point out that random-length wire antennas (less than  $2\lambda$ ) are not true longwire antennas. However, common (if erroneous) usage compels the inclusion of both types.

## Longwire antennas

Longwire antennas are any of several types of resonant and nonresonant antennas. Any given longwire antenna may be both resonant and nonresonant, depending upon the operating frequencies used. In the “old days,” when I was first starting in amateur radio, most resonant longwires were resonant over all HF bands because those bands were harmonically related to each other. But with the addition of the 10-, 18-, and 24-MHz band segments, that relationship no longer holds true for all cases.

Figure 9-1 shows the classic random length, nonresonant longwire antenna. It consists of a wire radiator that is at least a quarter-wavelength long, but is most often longer than a quarter-wavelength. The specific length is not critical, but it must be *greater than a quarter-wavelength at the lowest frequency of operation anticipated*. If you have a 90-ft wire, it will work on all HF bands above 3.5 MHz. In

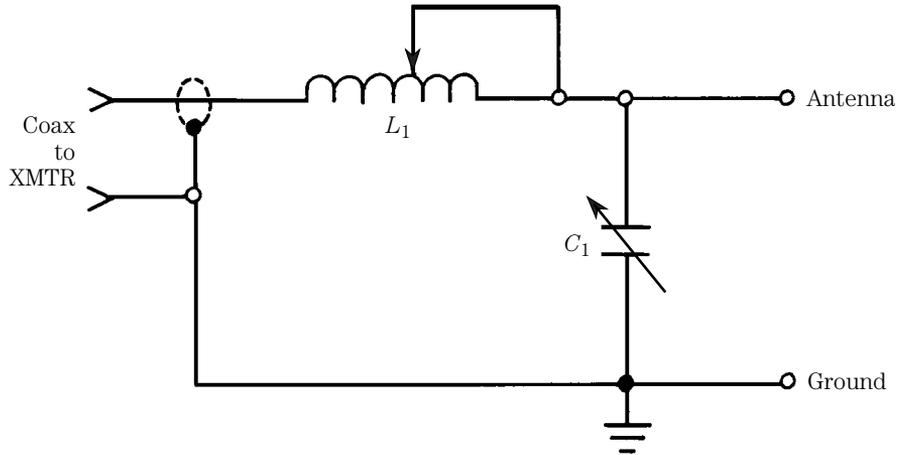


9-1 Random-length antenna.

most installations, the wire is no.12 and no.14 *copperweld*, or hard-drawn copper wire. I have successfully used both no.12 and no.14 house wire, but because it is solid (not stranded) it is not the best material. Stranded wire lasts longer in the wind, because solid wire fatigues and breaks quicker.

The longwire antenna is capable of providing gain over a dipole and a low angle of radiation (which is great for DX operators!). But these advantages are only found when the antenna is several wavelengths long, so it occurs only on typical HF antennas in the 21- through 29-MHz bands.

The longwire is end-fed, and therefore has a high impedance, except in those cases where the radiator happens to be quarter-wavelength. Because of this fact, it is necessary to use an antenna tuner between the low-impedance transmitter output (usually limited to 50- to 75- $\Omega$  impedances) and the antenna. You can buy any of several commercial antenna tuners, or make one yourself. Figure 9-2 shows a typical antenna tuner for longwires. It is an L-section coupler consisting of a series inductance and a shunt capacitance, both variable. The inductor can be homemade, or it can be made from a B&W miniductor (3029 is suitable). Alternatively, you can buy a used roller inductor from a hamfest. If you opt for the homemade or miniductor alternatives, then an alligator clip lead (short!) connected to one end of the coil can be used to short out unneeded turns when adjusting the inductance. I prefer the roller inductor method, because it allows the whole “shootin’ match” to be installed inside a shielded cabinet, helping the TVI/BCI (television interference/broadcast interference) situation.



9-2 L-section coupler.

The capacitor (also available from the same sources) should be 150-pF to 250-pF maximum capacitance high-voltage variable. So-called transmitting variables are usually ok if the plate spacing is at least  $\frac{1}{8}$  in.

Tuning of the longwire is simplicity itself. If the tuner does not have a built-in VSWR meter, then install one in the line between the transmitter output and the tuner input. Adjust both  $L_1$  and  $C_1$  (they are a bit interactive so do it several times) for the lowest VSWR. If you use the alligator clip method, *turn off the darn transmitter before adjusting the clip position—RF burns are nasty!*

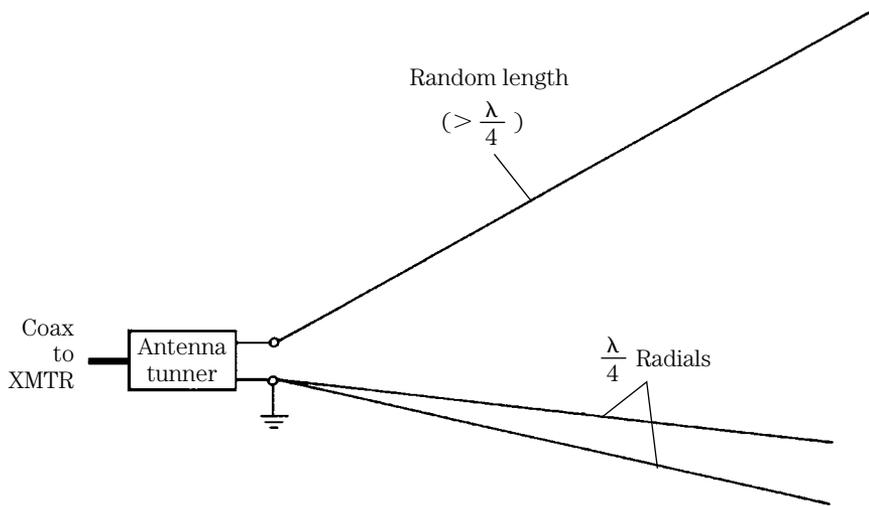
### So what's the problem?

Ok, so we have decided to install a longwire. How do we make it work? First, make sure that it is long enough. The bare minimum lengths for HF bands are 70 ft for 3.5 MHz and up, 34 ft for 7 MHz and up, and so on. In general, the longer the better. The second problem is a *good ground*. The importance of a good ground cannot be underestimated, and it accounts for about 99 percent of the difference in reported performance of longwires.

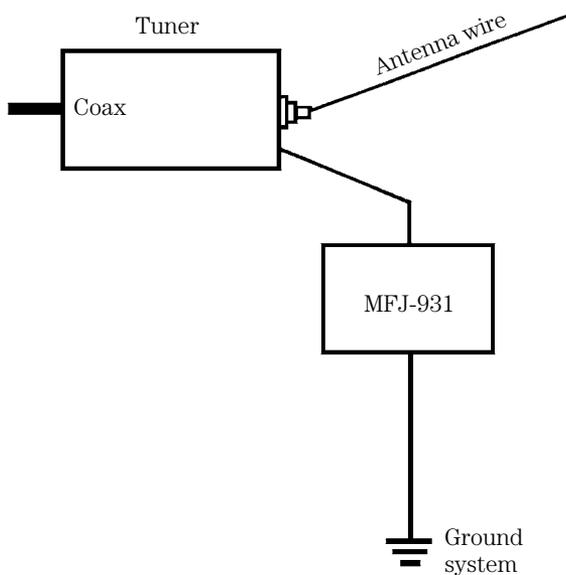
A good ground consists of a short wire to either one very long ground rod or multiple ground rods spaced a couple feet apart. An important factor is the length of the ground wire. It must be considerably less than a quarter-wavelength. Use no.10 stranded wire (several parallel lengths) or braid.

When I was in college I lived on the second floor of a whacky student boarding house in Norfolk, VA. The ground was 24 ft away, and that made "short ground wire" a joke—and my longwire nearly inoperative. But I figured out a way around the problem. A quarter-wave radial was dropped out the window (see Fig. 9-3). Use more than one radial if possible. In one case, when the landlord was particularly cranky, I tacked the radial to the baseboard of my room (insulated on standoffs at the far end to protect against the high RF voltages present). It worked well!

A new product also comes to the rescue of those reader-hams who cannot put up a good longwire. The MFJ Enterprises (Box 494, Mississippi State, MS, 39762, 1-800-647-1800) model MFJ-931 artificial RF ground is installed in the ground line (see Fig. 9-4) and is used to tune the ground wire. Adjust the capacitance and inductance controls for maximum ground current as shown on the built-in meter. Be sure to follow the instructions, however, because high RF voltages can appear on a nonresonant ground wire. I wish I'd had one of these when I was in that whacky boarding house; discussing existentialist poetry can't hold a candle to DXing.



9-3 Radials improve the "ground" of random-length antenna.



9-4 MFJ ground line tuner installation.

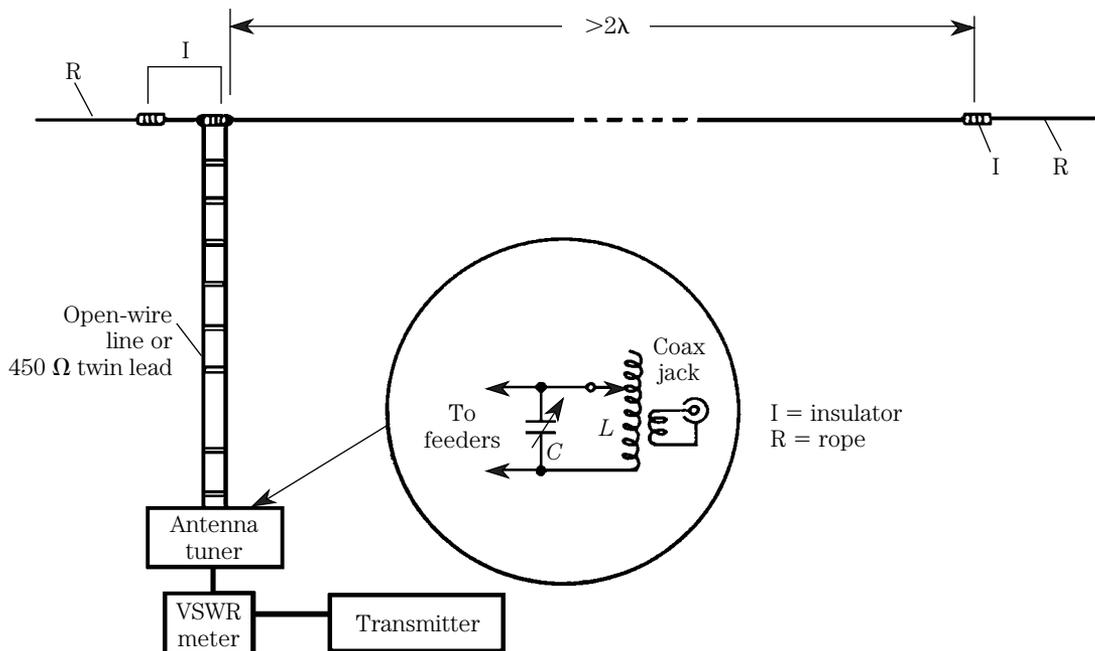
## When is a long wire a longwire?

For many years, the longwire has been a popular form of antenna. It is cheap, it is easy to construct, and—although reports vary—it has the potential to perform well. Properly constructed, it has a lot of utility. But what is a *real* longwire antenna? In past columns, I have used the term to mean an antenna such as Fig. 9-1. This form of antenna is popularly called a *longwire* if it is more than a quarter-wavelength. I and other authors have used the term longwire to mean this antenna, but that’s not rigorously correct. A *true* longwire is an antenna that is many wavelengths long, or to be a little more rigorous, an antenna that is *more than two wavelengths long*. Although I will still use the term *longwire* for both forms of longwire antenna, proper rigor requires Fig. 9-1 to be called a “random-length antenna.”

## True longwire antennas

Figure 9-5 shows the true resonant longwire antenna. It is a horizontal antenna, and if properly installed, it is not simply attached to a convenient support (as is true with the random length antenna). Rather, the longwire is installed horizontally like a dipole. The ends are supported (dipole-like) from standard end insulators and rope.

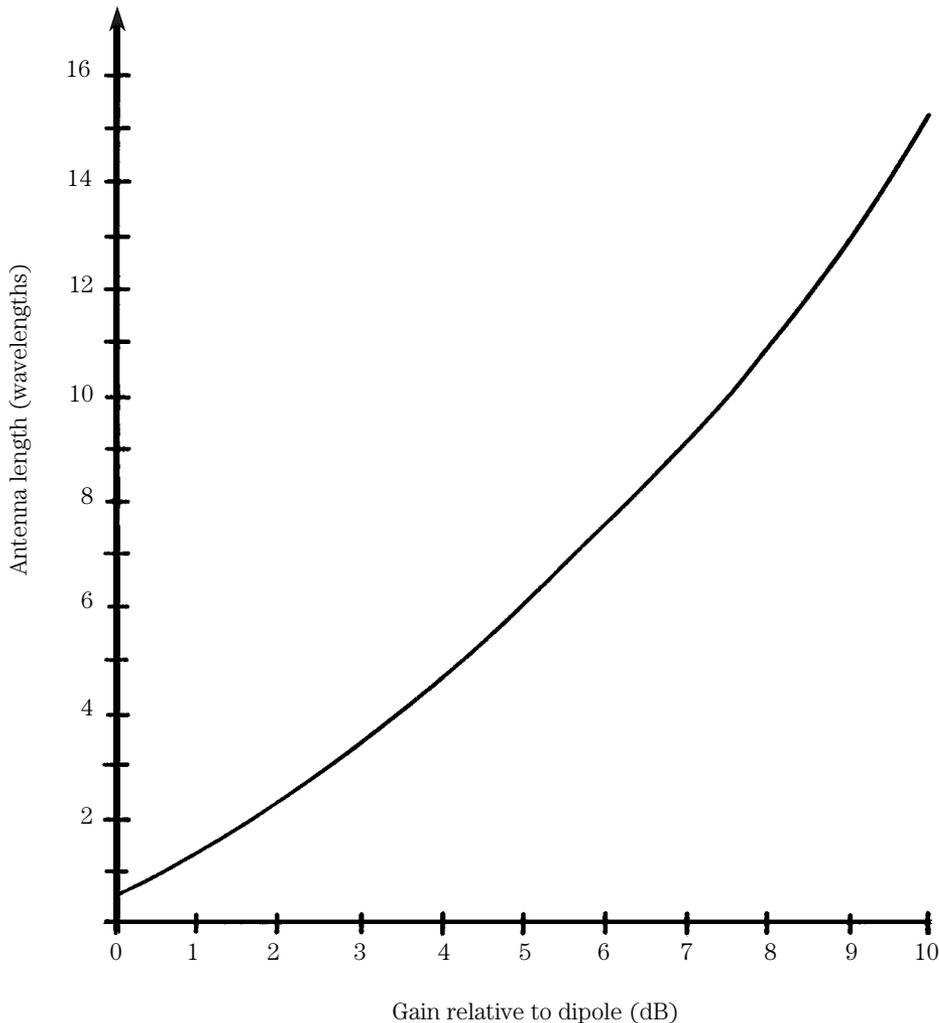
The feedpoint of the longwire is one end, so we expect to see a voltage antinode where the feeder is attached. For this reason we do not use coaxial cable, but rather either parallel transmission line (also sometimes called *open-air* line or some such



9-5 End-fed longwire with suitable tuner.

name), or 450- $\Omega$  twin lead. The transmission line is excited from any of several types of balanced antenna tuning unit (see Fig. 9-5). Alternatively, a standard antenna tuning unit (designed for coaxial cable) can be used if a 4:1 balun transformer is used between the output of the tuner and the input of the feedline.

What does “many wavelengths” mean? That depends upon just what you want the antenna to do. Figure 9-6 shows a fact about the longwire that excites many users of longwires: It has gain! Although a two-wavelength antenna only has a slight gain over a dipole; the longer the antenna, the greater the gain. In fact, it is possible to obtain gain figures greater than a three-element beam using a longwire, but only at nine or ten wavelengths.



**9-6** Antenna length versus gain over dipole.

What does this mean? One wavelength is  $984/F_{\text{MHz}}$  ft, so at 10 m (29 MHz) one wavelength is about 34 ft; at 75 m (3.8 MHz) one wavelength is 259 ft long. In order to meet the two-wavelength criterion a 10-m antenna need only be 68 ft long, while a 75-m antenna would be 518 ft long! For a ten-wavelength antenna, therefore, we would need 340 ft for 10 m; and for 75 m, we would need nearly 2,600 ft. Ah me, now you see why the longwire is not more popular. The physical length of a nonterminated resonant longwire is on the order of

$$L_{\text{feet}} = \frac{492(N - 0.025)}{F_{\text{MHz}}} \quad [9.1]$$

( $N$  is number of half-wavelengths in the radiator elements.)

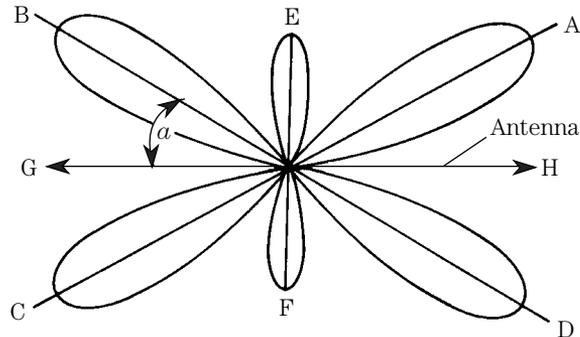
Of course, there are always people like my buddy (now deceased) John Thorne, K4NFU. He lived near Austin, TX on a multiacre farmette that has a 1400-ft property line along one side. John installed a 1300 ft longwire and found it worked excitingly well. He fed the thing with homebrew 450- $\Omega$  parallel (open-air) line and a Matchbox antenna tuner. John's longwire had an extremely low angle of radiation, so he regularly (much to my chagrin on my small suburban lot) worked ZL, VK, and other Southeast Asia and Pacific basin DX, with only 100 W from a Kenwood transceiver.

Oddly enough, John also found a little bitty problem with the longwire that textbooks and articles rarely mention: electrostatic fields build up a high-voltage dc charge on longwire antennas! Thunderstorms as far as 20 mi away produce serious levels of electrostatic fields, and those fields can cause a buildup of electrical charge on the antenna conductor. The electric charge can cause damage to the receiver. John solved the problem by using a resistor at one end to ground. The "resistor" is composed of ten to twenty 10-M $\Omega$  resistors at 2 W each. This resistor bleeds off the charge, preventing damage to the receiver.

A common misconception about longwire antennas concerns the normal radiation pattern of these antennas. I have heard amateurs, on the air, claim that the maximum radiation for the longwire is

1. Broadside (i.e., 90°) with respect to the wire run or
2. In line with the wire run

Neither is correct, although ordinary intuition would seem to indicate one or the other. Figure 9-7 shows the approximate radiation pattern of a longwire when viewed from above. There are four main lobes of radiation from the longwire ( $A$ ,  $B$ ,  $C$ , and  $D$ ). There are also two or more (in some cases many) minor lobes ( $E$  and  $F$ ) in the antenna pattern. The radiation angle with respect to the wire run ( $G-H$ ) is a function of the number of wavelengths found along the wire. Also, the number and extent of the minor lobes is also a function of the length of the wire.



9-7 Radiation pattern of longwire antenna.

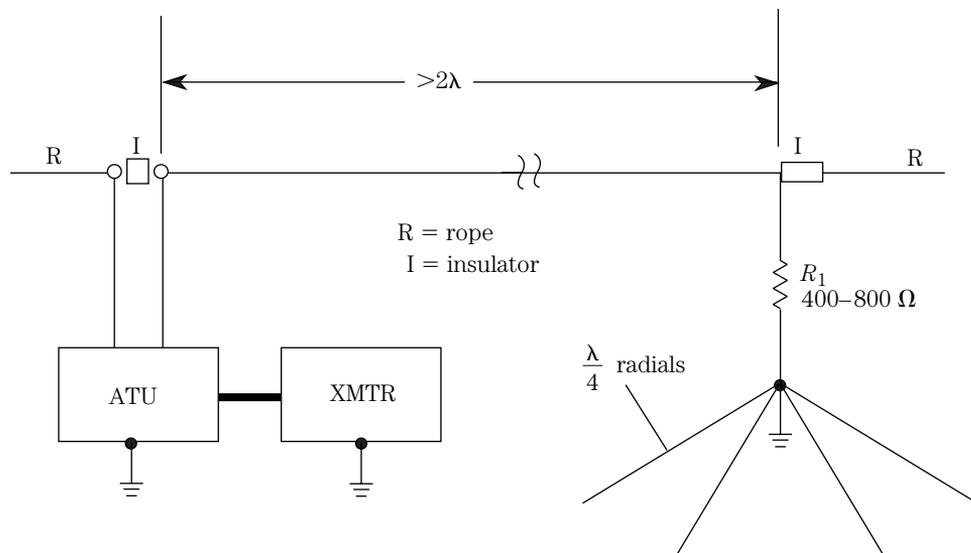
## Nonresonant single-wire longwire antennas

The resonant longwire antenna is a *standing wave antenna*, because it is unterminated at the far end. A signal propagating from the feedpoint, toward the open end, will be reflected back toward the source when it hits the open end. The interference between the forward and reflected waves sets up stationary standing current and voltage waves along the wire.

A *nonresonant longwire* is terminated at the far end in a resistance equal to its characteristic impedance. Thus, the incident waves are absorbed by the resistor, rather than being reflected. Such an antenna is called a *traveling wave antenna*. Figure 9-8 shows a terminated longwire antenna. The transmitter end is like the feed system for other longwire antennas, but the far end is grounded through a terminating resistor  $R_1$  that has a resistance  $R$  equal to the characteristic impedance  $Z_o$  of the antenna (i.e.,  $R = Z_o$ ). When the wire is 20 to 30 ft above the ground,  $Z_o$  is about 500 to 600  $\Omega$ .

The radiation pattern for the terminated longwire is a unidirectional version of the multilobed pattern found on the unterminated longwires. The angles of the lobes vary with frequency, even though the pattern remains unidirectional. The directivity of the antenna is partially specified by the angles of the main lobes. It is interesting to note that gain rises almost linearly with  $n\lambda$ , while the directivity function changes rapidly at shorter lengths (above three or four wavelengths the rate of change diminishes considerably). Thus, when an antenna is cut for a certain low frequency, it will work at higher frequencies, but the directivity characteristic will be different at each end of the spectrum of interest.

A two-wavelength ( $2\lambda$ ) pattern is shown in Fig. 9-7. There are four major lobes positioned at angles of  $\pm 36^\circ$  from the longwire. There are also four minor lobes—the strongest of which is  $-5$  dB down from the major lobes—at angles of  $\pm 75^\circ$  from the longwire. Between all of the lobes, there are sharp nulls in which little reception is possible. As the wire length is made longer, the angle of the main lobes pulls in tighter (i.e., toward the wire). As the lobes pull in closer to the wire, the number of minor lobes increases. At  $5\lambda$ , there are still four main lobes, but they are at angles of  $\pm 22^\circ$  from the wire. Also, the number of minor lobes increases to 16. The minor lobes are located at  $\pm 47^\circ$ ,  $\pm 62^\circ$ ,  $\pm 72^\circ$ , and  $\pm 83^\circ$  with respect to the wire. The minor lobes tend to be  $-5$  to  $-10$  dB below the major lobes. When the longwire gets very



**9-8** Longwire (greater than  $2\lambda$ ) antenna is terminated in a resistance (typically 400 to 800  $\Omega$ ). Radials under the resistor improve the grounding of the antenna.

much longer than  $5\lambda$ , the four main lobes begin to converge along the length of the wire, and the antenna becomes bidirectional. This effect occurs at physical lengths greater than about  $20\lambda$ .

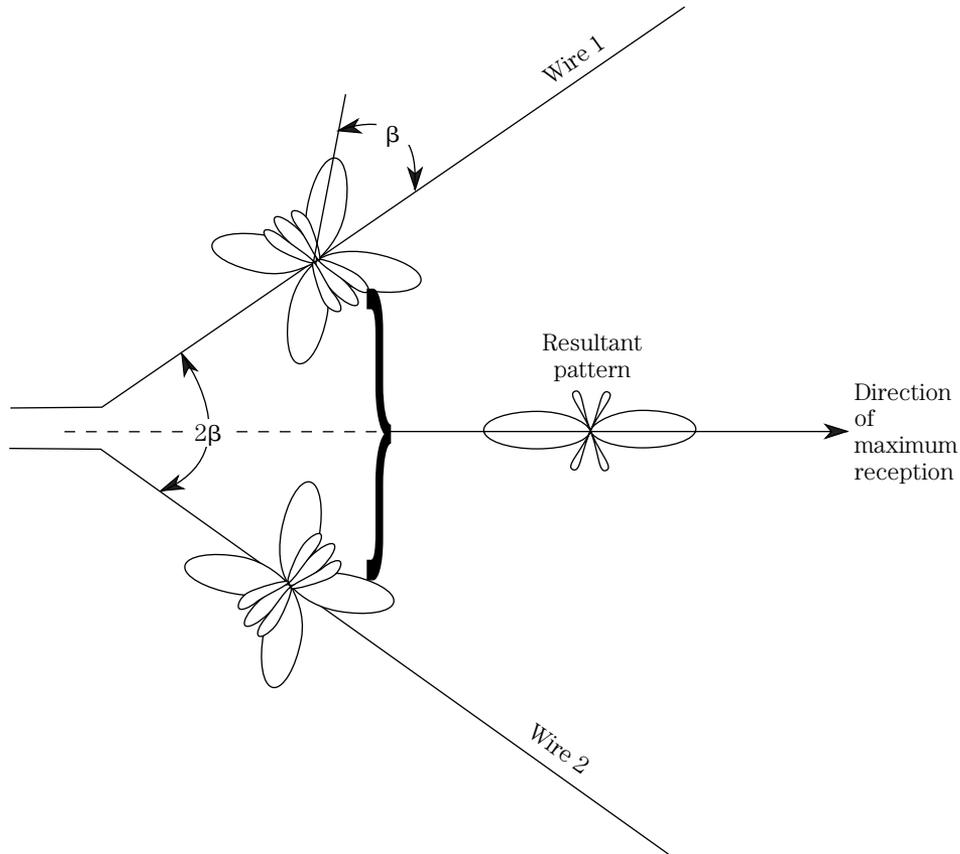
In general, the following rules apply to longwire antennas:

- On each side of the antenna, there is at least one lobe, minor or major, for each half-wavelength of the wire element. For the overall element, there is one lobe for every quarter-wavelength.
- If there are an even number of lobes on either side of the antenna wire, then half of the total number of lobes are tilted backward, and half are tilted forward; symmetry is maintained.
- If there are an odd number of lobes on either side of the wire, then one lobe on either side will be perpendicular to the wire, with the other lobes distributed either side of the perpendicular lobe.

### Vee beams and rhombic beams

Longwire antennas can be combined in several ways to increase gain and sharpen directivity. Two of the most popular of these are the *vee beam* and the *rhombic* antennas. Both forms can be made in either resonant (unterminated) or nonresonant (terminated) versions.

**Vee beams** The *vee beam* (Fig. 9-9A) consists of two equal-length longwire elements (wire 1 and wire 2), fed  $180^\circ$  out of phase with each other, and spaced to produce an acute angle between them. The  $180^\circ$  phase difference is inherent in connecting the two wires of the vee to opposite conductors of the same parallel conductor feedline.



**9-9A** Radiation pattern of vee beam antenna consists of the algebraic sum of the two longwire patterns that make up the antenna.

The unterminated vee beam (Fig. 9-9A) has a bidirectional pattern that is created by summing together the patterns of the two individual wires. Proper alignment of the main lobes of the two wires requires an included angle, between the wires, of twice the radiation angle of each wire. If the radiation angle of the wire is  $\beta$ , then the appropriate included angle is  $2\beta$ . To raise the pattern a few degrees, the  $2\beta$  angle should be slightly less than these values. It is common practice to design a vee beam for a low frequency (e.g., 75-/80- or 40-m bands), and then to use it also on higher frequencies that are harmonics of the minimum design frequency. A typical vee beam works well over a very wide frequency range *only* if the included angle is adjusted to a reasonable compromise. It is common practice to use an included angle that is between  $35^\circ$  and  $90^\circ$ , depending on how many harmonic bands are required.

Vee beam patterns are based on an antenna height that is greater than a half-wavelength from the ground. At low frequencies, such heights will not be practical, and you must expect a certain distortion of the pattern because of ground reflection effects.

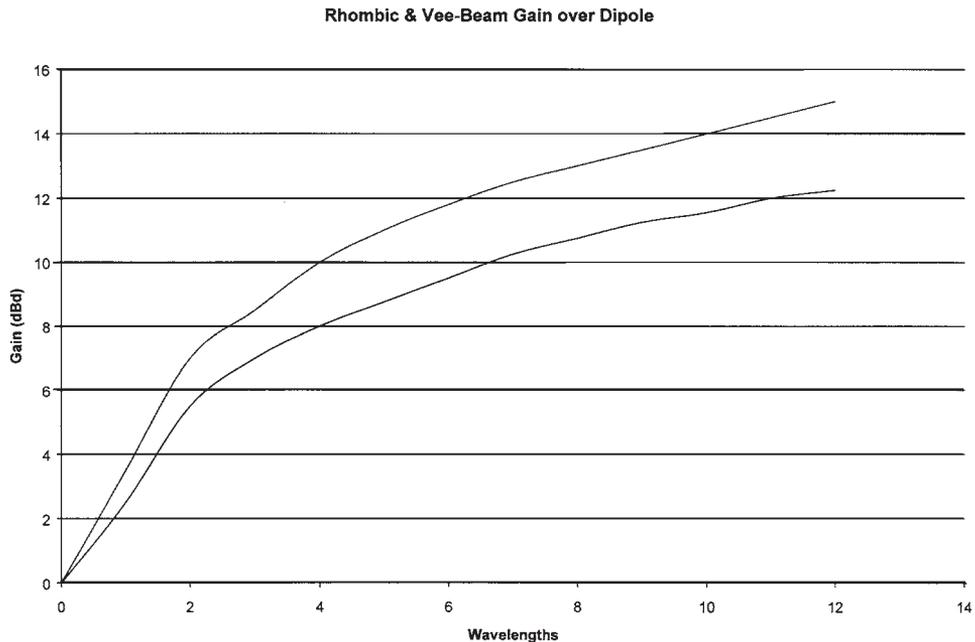
Gain on a vee beam antenna is about 3 dB higher than the gain of the single-wire longwire antenna of the same size, and it is considerably higher than the gain of a

dipole (see Fig. 9-9B). At three wavelengths, for example, the gain is 7 dB over a dipole. In addition, there may be some extra gain because of mutual impedance effects—which can be about 1 dB at  $5\lambda$  and 2 dB at  $8\lambda$ .

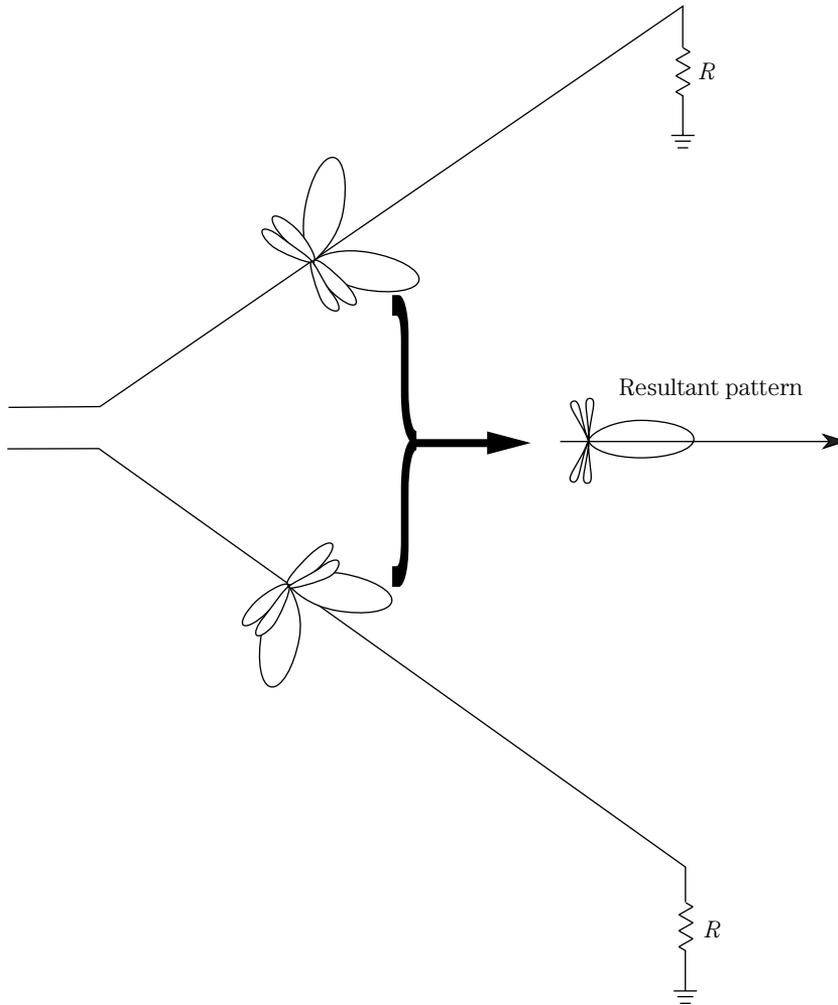
**Nonresonant vee beams** Like the single-wire longwire antennas, the vee beam can be made nonresonant by terminating each wire in a resistance that is equal to the antenna's characteristic impedance (Fig. 9-10). Although the regular vee is a standing wave antenna, the terminated version is a traveling wave antenna and is thus unidirectional. Traveling wave antennas achieve unidirectionality because the terminating resistor absorbs the incident wave after it has propagated to the end of the wire. In a standing-wave antenna, that energy is reflected backwards toward the source, so it can radiate oppositely from the incident wave.

**Rhombic beams** The *rhombic beam antenna*, also called the *double vee*, consists of two vee beams positioned end-to-end with the tips connected. The unidirectional, nonresonant (terminated) rhombic is shown in Fig. 9-11. The unterminated resonant form gives approximately the same gain and directivity as a vee beam of the same size. The nonresonant rhombic has a gain of about 3 dB over a vee beam of the same size (see Fig. 9-9 again).

Two angles are present on the rhombic antenna. One-half of the included angle of the two legs of one wire is the *tilt angle* ( $\phi$ ), while the angle between the two wires is the *apex angle* ( $\theta$ ). A common rhombic design uses a tilt angle of  $70^\circ$ , a length of  $6\lambda$  for each leg (two legs per side), and a height above the ground of  $1.1\lambda$ .



**9-9B** Gain versus length of vee beam and rhombic beam antennas.

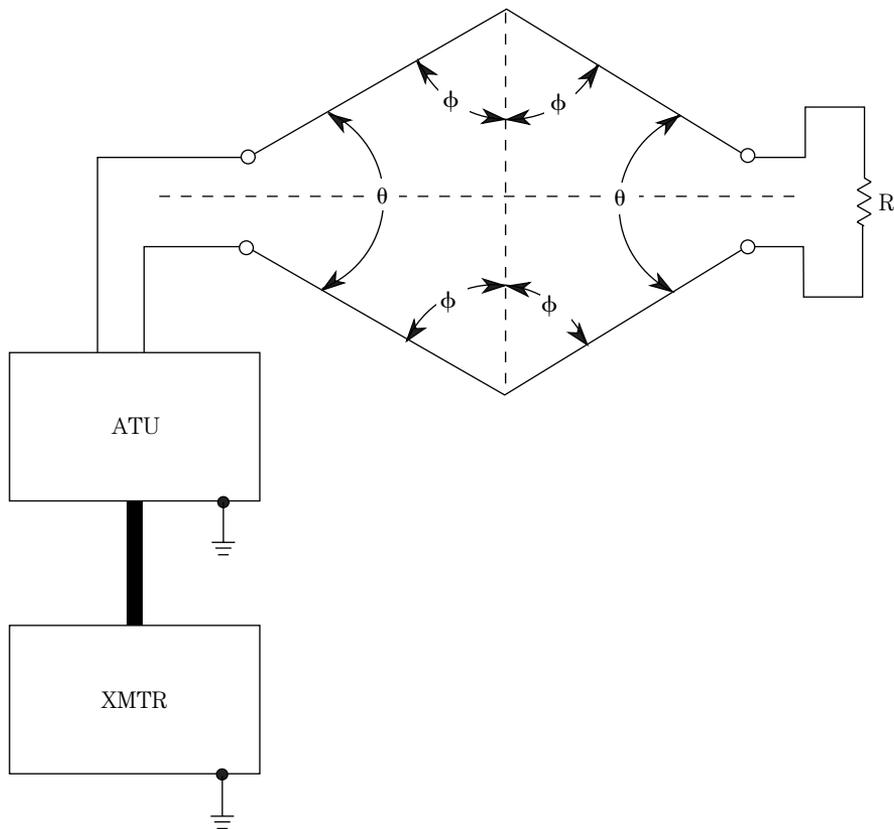


**9-10** Nonresonant vee beam antenna.

The termination resistance for the nonresonant rhombic is 600 to 800  $\Omega$ , and must be noninductive. For transmitting rhombics, the resistor should be capable of dissipating at least one-third of the average power of the transmitter. For receive-only rhombics, the termination resistor can be a 2-W carbon composition or metal-film type. Such an antenna works nicely over an octave (2:1) frequency range.

### **Beverage or “wave” antennas**

The *Beverage* or *wave antenna* is considered by many people to be the best receive antenna available for very low frequency (VLF), AM broadcast band (BCB), medium



9-11 Terminated rhombic antenna.

wave (MW), or tropical band (low HF region) DXing. The Beverage was used by RCA at its Riverhead, Long Island (NY) station in 1922, and a technical description by Dr. H. H. Beverage (for whom it is named) appeared in *QST* magazine for November 1922, in an article titled “The Wave Antenna for 200-Meter Reception.” In 1984, an edited and updated version of the 1922 article appeared in the same magazine. In 1921, Paul Godley, under sponsorship of the American Radio Relay League, journeyed to Scotland to erect a receiving station at Androssan. His mission was to listen for amateur radio signals from North America. As a result of politicking in the post-World War I era, hams were consigned to the supposedly useless shortwave ( $\lambda < 200$  m), and it was not clear that reliable international communications were possible. Godley went to Scotland to see if that could happen; he reportedly used a wave antenna for the task (today, called the *Beverage*).

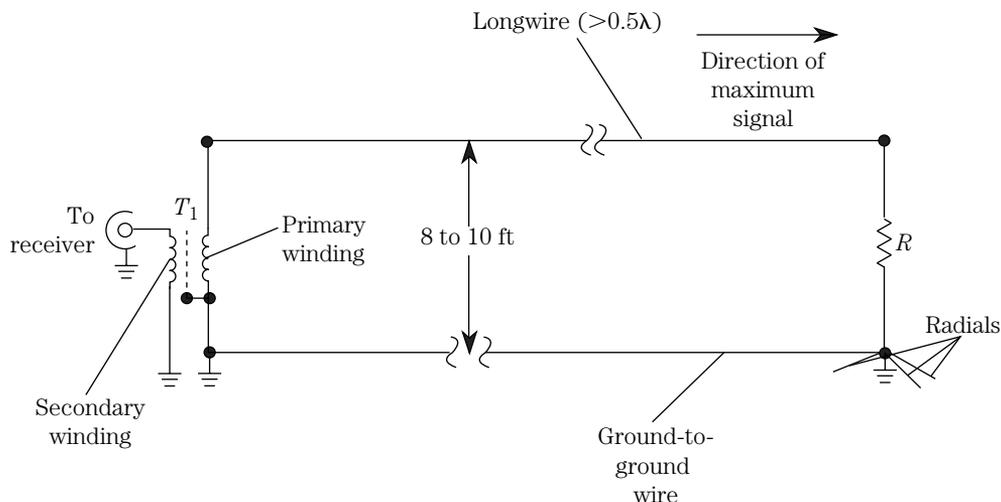
The Beverage antenna is a longwire of special design, more than one wavelength ( $1\lambda$ ) long (Fig. 9-12), although some authorities maintain that  $>0.5\lambda$  is minimally sufficient. The Beverage provides good directivity and good gain, but is not very efficient.

As a result, it is preferred for receiving, and is less useful for transmitting. This is an example of how different attributes of various antennas make the Law of Reciprocity an unreliable sole guide to antenna selection. Unlike the regular longwire, which is of a different design, the Beverage is intended to be mounted close to the earth's surface (typically  $< 0.1\lambda$ ); heights of 8 to 10 feet is the usual prescription.

Figure 9-12 shows the basic single-wire Beverage antenna. It consists of a single conductor (no.16 to no.8 wire, with no.14 being most common) erected about 8 to 10 ft above ground. Some Beverages are unterminated (and bidirectional), but most of them are terminated at the far end in a resistance  $R$  equal to the antenna's characteristic impedance  $Z_o$ . The receiver end is also terminated in its characteristic impedance, but generally requires an impedance matching transformer to reduce the antenna impedance to the  $50\text{-}\Omega$  standard impedance used by most modern transmitters.

The Beverage works best in the low-frequency bands (VLF through MW), although at least some results are reportedly relatively easy to obtain up to 25 m (11.5 MHz). Some questionably successful attempts have been made at making Beverage antennas work as high as the 11-m Citizens Band or the 10-m ham band (29.7 MHz).

The Beverage antenna works on vertically polarized waves arriving at low angles of incidence. These conditions are normal in the AM BCB, where nearly all transmitting antennas are vertically polarized. In addition, the ground- and sky-wave propagations found in these bands (VLF, BCB, and low HF) are relatively consistent. As the frequency increases, however, two factors become increasingly dominant. First, the likelihood of horizontal polarization increases because of the size of a wavelength at those frequencies. The polarization of the received signal not only changes in those bands, but does so constantly when conditions are unsettled. It is the strong dependence of the



9-12 Beverage antenna.

Beverage on relatively constant vertical polarization that makes me suspect the claims of Beverage-like performance above the 25- or even 31-m bands.

The Beverage depends on being erected over poorly conductive soil, even though the terminating resistor needs a good ground. Thus, one source claimed that sand beaches adjacent to salty marshes make the best Beverage sites (a bit of an overstatement). Figure 9-13 shows why poorly conductive soil is needed. The E-field vectors are launched from the transmitting antenna perpendicular to the earth's surface. Over perfectly conducting soil, the vertical waves would remain vertical. But over imperfectly conducting soil, the field lines tend to bend close to the point of contact with the ground. As shown in the inset of Fig. 9-13, the bending of the wave provides a horizontal component of the E-field vector, and this provides the means of generating an RF current in the conductor wire.

A debate among Beverage antenna fans regards the best length for the antenna. Some sources state that the length can be anything greater than or equal to  $0.5\lambda$ , yet others say greater than or equal to  $1\lambda$  is the minimum size. One camp says that the length should be as long as possible, while others say it should be close to a factor called the *maximum effective length* (MEL), which is

$$\text{MEL} = \frac{\lambda}{4 \left( \frac{100}{K} - 1 \right)} \quad [9.2]$$

where

MEL is the maximum effective length, in meters

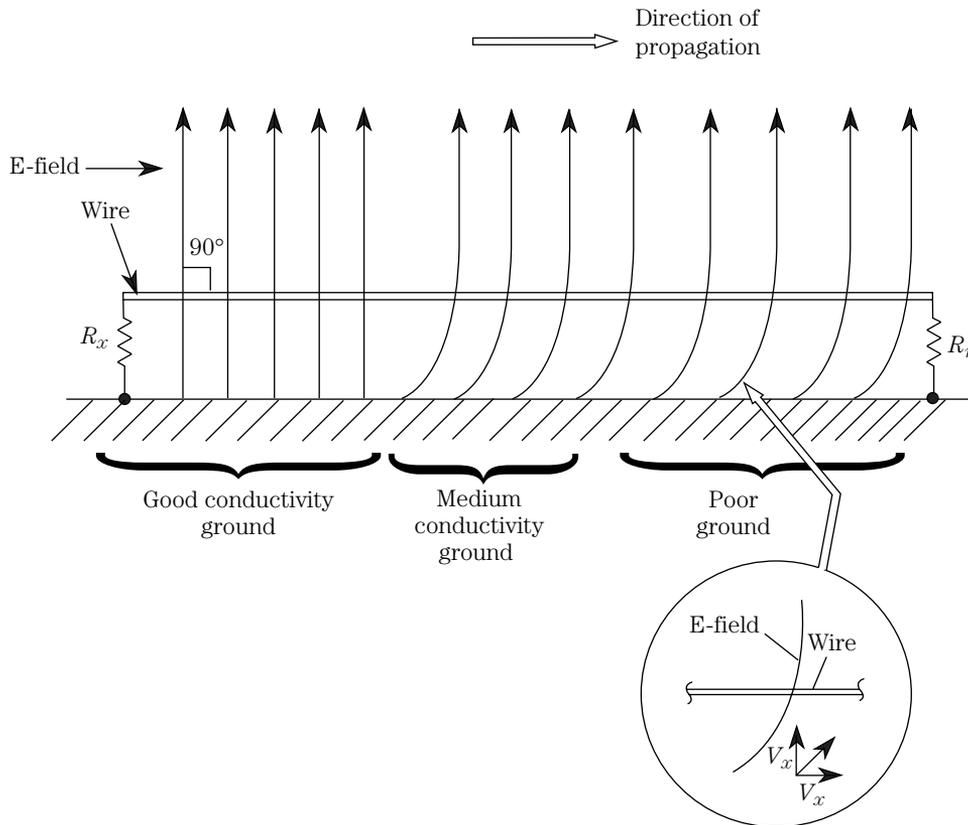
$\lambda$  is the wavelength, in meters

$K$  is the velocity factor, expressed as a percent

Misek, who may well be the leading exponent of the Beverage antenna, uses numbers like  $1.6\lambda$  to  $1.7\lambda$  over the 1.8- to 7.3-MHz region, and  $0.53\lambda$  to  $0.56\lambda$  on frequencies lower than 1.8 MHz. Doctor Beverage was once quoted as saying that the optimum length is  $1\lambda$ .

Like the longwire antenna, the Beverage needs a termination resistor that is connected to a good ground. This requirement might be harder to meet on Beverage antennas because they work best over lossy ground, which doesn't make a very good ground connection. As in the longwire case, insulated or bare wires, a quarter-wavelength long, make the best radials. However, a substantial improvement in the ground is possible using just bare wires measuring 15 to 20 ft long (which is much less than  $\lambda/4$ ), buried in the soil just below the surface (far enough to prevent erosion from bringing it to the surface). Many articles and books on Beverages show ground rods of 2 or 3 ft long, which borders on the ridiculous. Poor soil requires longer ground rods, on the order of 6 to 8 ft. Copper-clad steel makes the best rods.

In addition to the radials and ground rod, Misek also recommends using a wire connection between the ground connection at the termination resistor, and the



**9-13** Performance of Beverage antenna over different values of ground conductance.

ground connection at the receiver transformer (see again Fig. 9-12). According to Misk, this wire helps to stabilize the impedance variations at higher frequencies.

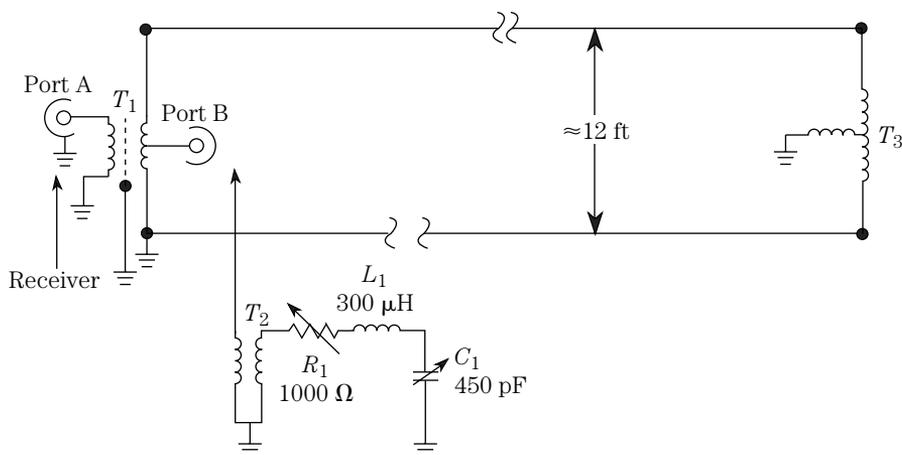
Installation of the Beverage antenna is not overly critical if certain rules are followed. The antenna should be installed at a height of 6 to 10 ft off the ground, and it should be level with the ground over its entire length. If the ground is not flat enough to make a level installation possible, then try to use a height that is 6 to 10 ft above the average terrain elevation along its run. A popular installation method is to erect 16-ft  $4 \times 4$  lumber, such that 3 to 4 ft are buried in a concrete-filled posthole. Use lumber that is treated for outdoor use (i.e., pressure-treated lumber sold for decks and porches). The wire can be fastened to the  $4 \times 4$  posts using either ceramic standoff ("beehive") insulators or electric fence insulators (which some people deem preferable). Try to use one contiguous length of wire for the antenna, if possible, in order to avoid soldered splices and joints.

One of the Beverage installation difficulties shared with the longwire is the need to slope down to a point where a termination resistor can be easily installed close to the ground. While the longwire can be sloped over a large portion of its length, the Beverage should be sloped downward only over the last 60 ft or so.

### Steerable notch Beverage antennas

A Beverage erected with two wires—parallel to each other, at the same height, spaced about 12 in apart (Fig. 9-14), with a length that is a multiple of a half-wavelength—is capable of *null steering*. That is, the rear null in the pattern can be steered over a range of  $40^\circ$  to  $60^\circ$ . This feature allows strong, off-axis signals to be reduced in amplitude so that weaker signals in the main lobe of the pattern can be received. There are at least two varieties of the *steerable wave Beverage* (SWB).

If null steering behavior is desired, then a phase control circuit (PCC) will be required—consisting of a potentiometer, an inductance, and a variable capacitor in series with each other. Varying both the “pot” and the capacitor will steer the null. You can select the direction of reception, and the direction of the null, by using a switch to swap the receiver and the PCC between port A and port B.



9-14 Beverage antenna with a steerable null.

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## 10 CHAPTER

# Hidden and limited-space antennas

ONE OF THE MOST SIGNIFICANT IMPEDIMENTS TO AMATEUR RADIO OPERATORS, CB operators, and shortwave listeners is the space available for their antennas. In many thousands of other cases, the limitation is less one of space, but of regulators. More and more subdivisions are built with covenants on the deed that prohibit the buyer from installing outdoor antennas. Once limited to townhouse developments, where that breed of contemptible vermin (called the Homeowners Committee) routinely intruded on the affairs of people who mistakenly think they “own” their townhouse (ownership implies right of use, which is limited by the covenants). These onerous covenants are now routinely placed on single-family dwellings as well. In fact, it is the single most serious threat to amateur communications people in the country today. Other homeowners are no longer angered by the restriction on antennas because their television reception is now carried to them via cable systems in most parts of the country. Even where cable is not available, most users can install a moderately sized television antenna in their attic, or use rabbit ears. In other cases, the townhouse community will install a single master TV antenna and then distribute signals to each unit. The result is that the amateur radio operator, CB operator, and shortwave listener are left to fend for themselves without assistance from neighbors.

In this chapter we will examine some of the alternatives available to those readers who have either a limited space situation (such as a small city lot), or are unable to move out of a subdivision where there are stupid rules against outdoor antennas. The suggestions contained in this chapter are not universal, and indeed the author recommends that you adapt, as well as adopt, these recommendations, and come up with some of your own. Creativity within the constraints of the laws of physics governing radio antennas is encouraged.

## Hidden antennas

A hidden antenna is one that is either completely shielded from view or disguised as something else. Alternatively, you could also include in this category antennas that are in semiopen view, but which are not too obvious (except to the trained and diligent eye). Some people have opted for “hidden” longwires made of very fine wire (no. 26 enameled wire is popular). The user will install the wire in the open, high off of the ground (as in an apartment installation), and operate without anyone knowing the difference. One chap used no. 22 wire suspended between two 16-story apartment house buildings that were 100 yards apart. He had one of the best working longwires in town until a windy day when the whole thing came down. No one was injured; but had someone been hurt, there might have been a lawsuit. Hidden antennas must be designed with an eye toward causing others no harm. It is neither ethical nor smart to place others at risk in enjoying our hobbies.

The dipole is a popular antenna with both shortwave listeners and amateur radio operators. Indeed, for the CB operator who wants to get on the air from an apartment or townhouse (or restricted single-family home), the dipole can represent a respectable alternative.

As you learned in Chap. 6, the dipole is a horizontal wire (or pipe) antenna that is a half-wavelength long and fed in the center (ideally) with 75- $\Omega$  coaxial cable. In the “townhouse” dipole, it is possible to build the antenna entirely inside the attic of the building. The length of the dipole is given approximately by

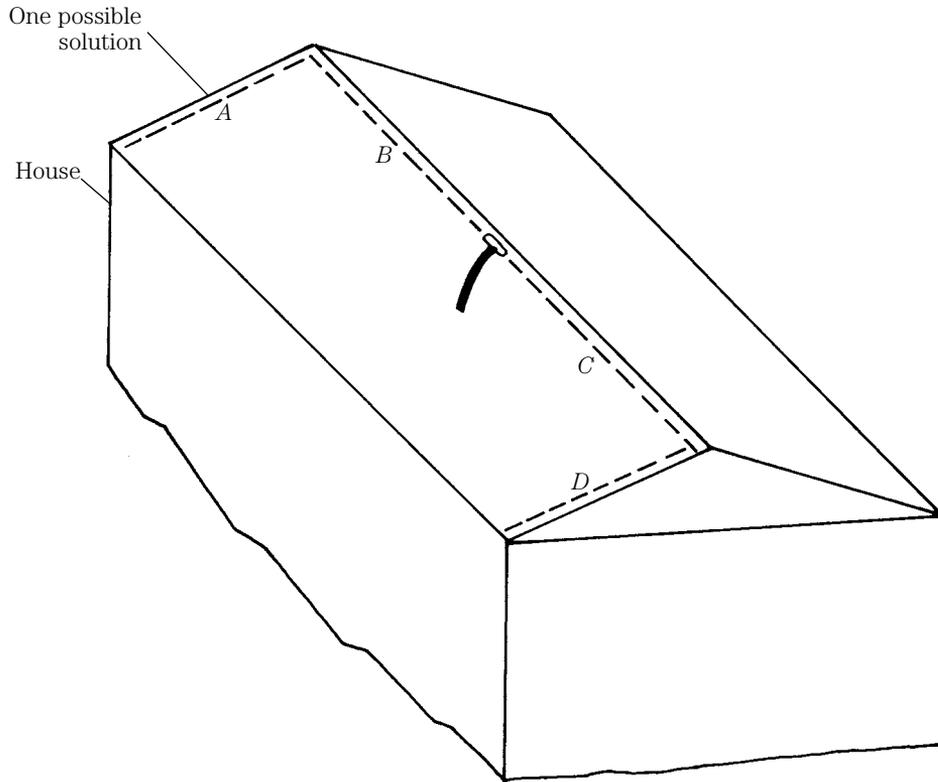
$$L_{ft} = \frac{468}{F_{\text{MHz}}} \quad [10.1]$$

If you do some quick calculations you will find that antennas for the 10-, 13-, 15-, and (possibly) even 18- and 20-m bands, will fit entirely inside the typical townhouse attic. This statement is also true of the 11-m citizen’s band antenna: It will fit inside the standard townhouse attic. But what about the lower frequency bands?

Figure 10-1 shows a possible solution to the use of the lower frequencies in the townhouse situation. The two quarter-wavelength arms of the dipole are ideally installed in line with each other, as was shown in the chapter on horizontal antennas. But in a sticky situation we can also install the dipole with the arms bent to accommodate the space available. In this example, only one of many possible methods for accomplishing this job is shown. Here each quarter-wavelength section is composed of two legs, *AB* and *CD*, respectively. Ideally, segments *B* and *C* are the longest dimensions. Also, if possible, make segments *A* and *D* equal lengths. Another method is to reverse the direction of one end leg, say for example *D*, and run it to the other corner of the building over the peak of the roof.

The author is almost hesitant to offer only one drawing, despite space constraints, in the fear that readers will take the offered pattern as the only authorized version, or solution. In reality, you might not be able (for a variety of reasons) to use the exact pattern shown (so ad lib a little bit).

How about performance? Will the constrained dipole of Fig. 10-1 work as well as a regular dipole installed a wavelength or two off the ground and away from objects?



10-1 Installing dipole in attic.

In a word: no. But that is not the problem being solved; getting on the air at all is the present problem. You will find that the pattern of the constrained dipole is distorted compared with that of the regular dipole. In addition, the feedpoint impedance is not going to be  $73 \Omega$  (except if by fluke), so you will be required to use an antenna tuner of some kind. You are well advised to read the chapter on antenna instruments and measurements so that you can be prepared to figure out any feed problems that crop up—as well they might.

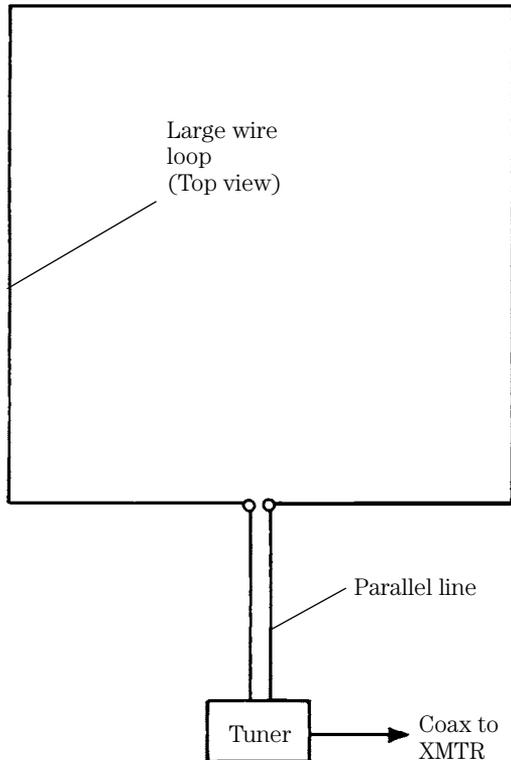
The wire used in the constrained dipole (or other forms of attic antenna) should be mounted on TV-type screw-in standoff insulators. Almost any outlet that sells TV antennas, or installation parts, will have them. These standoff insulators are also available in many hardware stores, and department stores that sell TV antennas or “Harry and Harriet Homeowner” supplies, in addition to electronics parts suppliers. Do not simply tape the wire to the wooden underside of your roof. The reason is simple: in a poorly tuned antenna, voltages can get high enough at the ends to produce *corona* effects; arcing could be a fire hazard. Also, use insulated wire to avoid accidental contact in the event that someone is in the attic when you are operating. Be aware that the insulation might affect the propagation velocity of signals in the wire, so it will slightly alter the required length.

Another alternative for the attic antenna, and one that avoids some of the problems of corona effects, is the nonresonant loop shown in Fig. 10-2. Although specified as a top view, the loop can be installed in the configuration that best uses the space. In fact, the best performance will be bidirectional when the loop is installed vertically. In this case, a large loop (as large as can be accommodated in the space available) is installed in the attic. Again, use standoff insulators and insulated wire for the installation.

The giant loop is fed with parallel line, and is tuned with a balanced antenna tuning unit. As was true with the constrained dipole, the performance is not to be equated with the performance of more regular antennas; but, with the prospect of not being on the air at all . . . ! Once again we have a compromise antenna for a compromising situation.

Another ploy is the old “flagpole trick” shown in Fig. 10-3. Some developments allow homeowners to vent their patriotism by installing flagpoles—and flagpoles can be disguised antennas. In the most obvious case, you can install a brass or aluminum flagpole and feed it directly from an antenna tuning unit. For single-band operation, especially on the higher frequencies, you can delta-feed the “flagpole” unobtrusively, and call your flagpole a vertical antenna. But that is not always the best solution.

Figure 10-3 shows two methods for creating a flagpole antenna, and both depend on using white PVC plumbing pipe as the pole. The heavier grades of PVC pipe

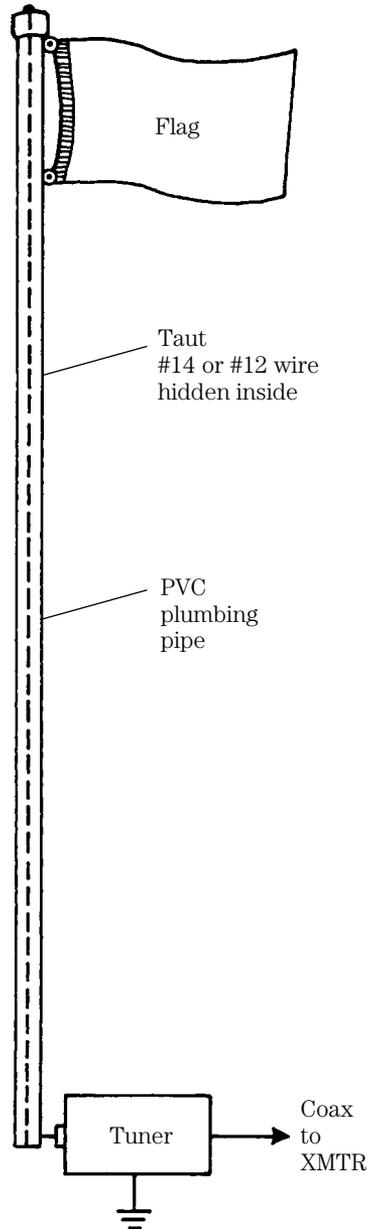


**10-2** Large loop antenna.

are self-supporting to heights of 16 to 20 ft, although lighter grades are not self-supporting at all (hence are not usable).

Figure 10-3A shows the use of a PVC flagpole in which a no. 12 (or no. 14) wire is hidden inside. This wire is the antenna radiator. For some frequencies, the wire will be resonant, and for others, it will surely be nonresonant. Because of this problem an antenna tuning unit is used either at the base of the antenna or inside at the transmitter.

**10-3A** Flagpole antenna made from thick-walled PVC pipe.



If the wire is too long for resonance (as might happen in the higher bands), then place a capacitor in series with the wire. Various settings may be required, so use a multi-section transmitting variable that has a total capacitance selectable to more than 1000 pF. Alternatively, use a vacuum variable capacitor of the same range.

In cases where the antenna is too short for resonance, as will occur in the lower bands, insert an inductance in series with the line to “lengthen” it. Another alternative is to use an L-section tuner at the feedpoint.

A good compromise situation is the use of a 16-ft length of “flagpole” pipe with a 16-foot wire embedded inside. The 16-foot wire is resonant at 20 m, so it will perform like a vertical antenna at those frequencies. The tuner will then accommodate frequencies above and below 20 m.

Another alternative is the version shown in Fig. 10-3B. In this case the wire radiator is replaced with a section of aluminum tubing. A wooden or plastic insert is fashioned with a drill and file to support the aluminum tubing inside the PVC tubing. One way to make the support is to use a core bit in an electric drill to cut out a disk that fits snugly inside the PVC tubing. Rat out the center hole, left by the core bit pilot, to the outside diameter of the aluminum tubing. The support can be held in place with screws from the outside, or simply glued in place.

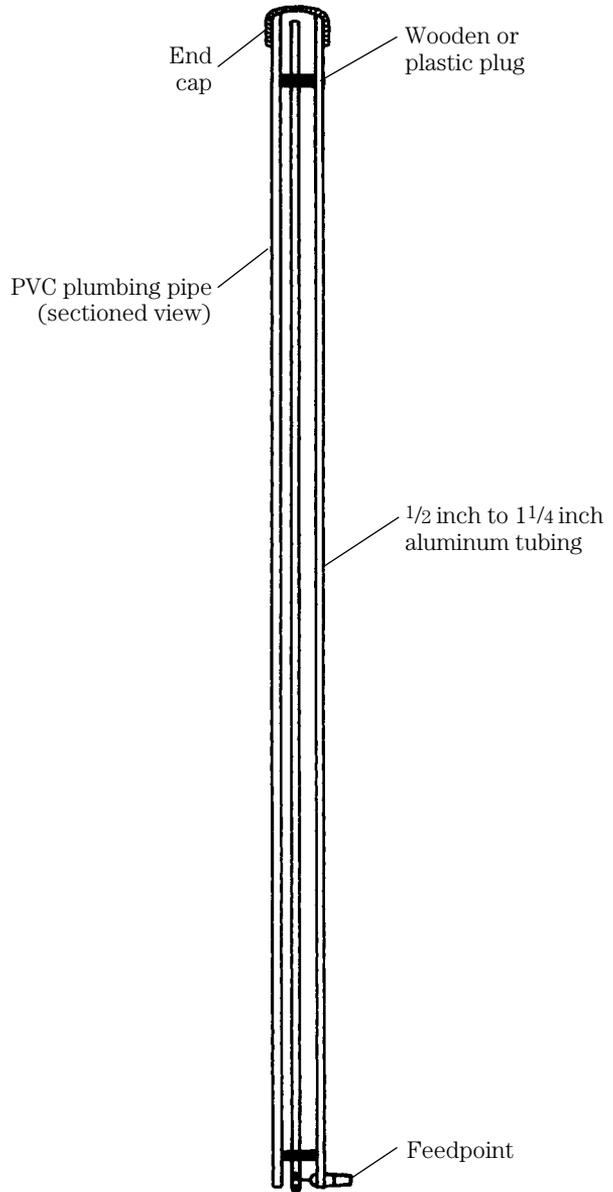
The problem of operating with a hidden antenna is a serious challenge. But with some of these guidelines and a little creativity, you can get on the air and enjoy your amateur radio hobby.

## Limited-space antennas

Many people live in situations where it is permissible to install an outdoor antenna, but it is not practical to install a full-size antenna. I once lived in a house that was 16 ft wide and 37 ft long, on a 33 × 100-ft lot. Very few full-sized antennas could be installed on that lot because of the space constraints. Beam antennas were out because county laws required that the antenna not hang over the property line. Although vertical antennas were possible, there was a period of time when I used other antennas that were easier to install. In this chapter, you will examine some of the options open to those with limited space for amateur radio, CB, or SWL antennas.

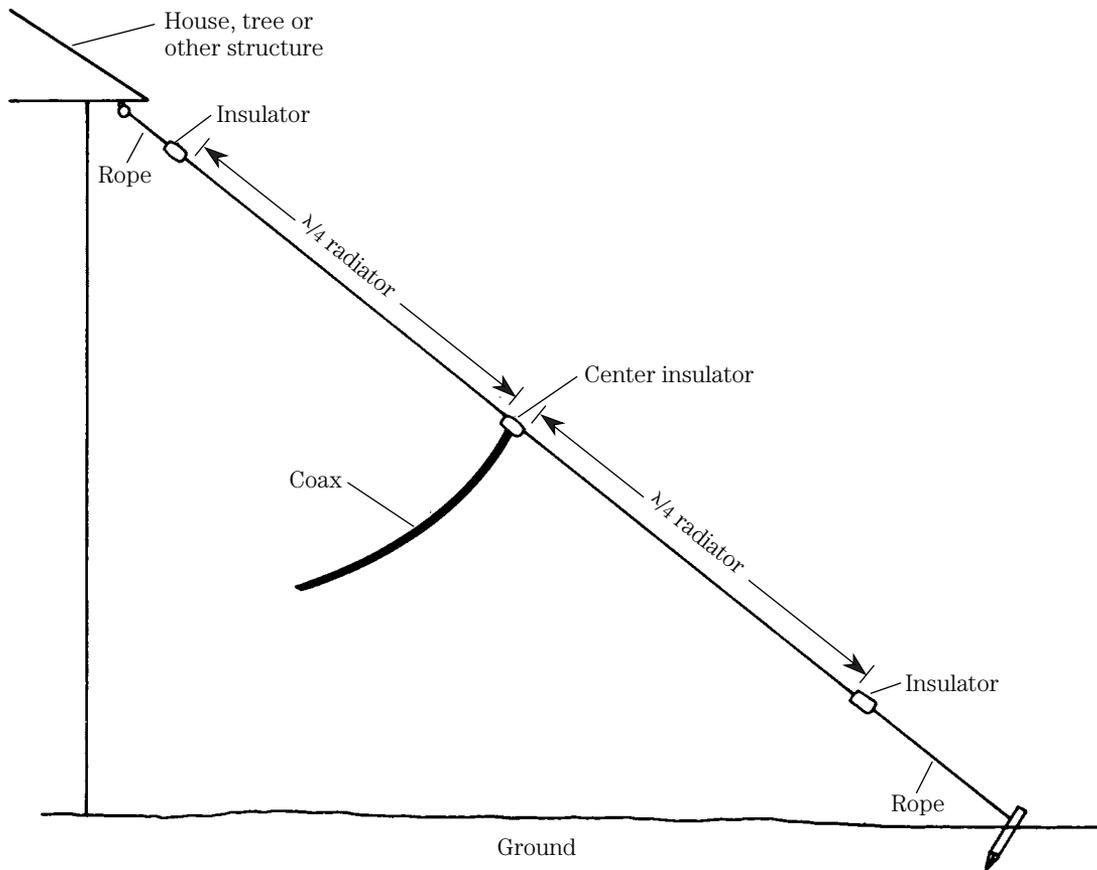
Once again we return the simple dipole as the basis for our discussion. In Fig. 10-4, you see several alternatives for installing an outdoor antenna in a limited space. In Fig. 10-4A, the slanted dipole (or slipole) antenna uses the standard dipole configuration, but one end is connected to the high point of the building, while the other is anchored near the ground. The coaxial cable is connected to the midpoint of the antenna in the usual manner for regular dipoles. If the end of the dipole is within reach of people on the ground, then they may get a nasty RF burn if the antenna is touched while you are operating. Take precautions to keep people and pets away from this lower end.

Another method is the vee dipole shown in Fig. 10-4B. In a regular dipole installation, the ends of the antenna are along the same axis (in other words, at an angle of 180°). In the example of Fig. 10-4B, however, the angle between the elements is less than 180°, but greater than 90°. In some cases we might want to bend the elements, rather than install them in a vee shape. Figure 10-4C shows an angled dipole with four segments. For the best performance (but not as good as that



**10-3B** PVC flagpole hiding an aluminum tubing vertical.

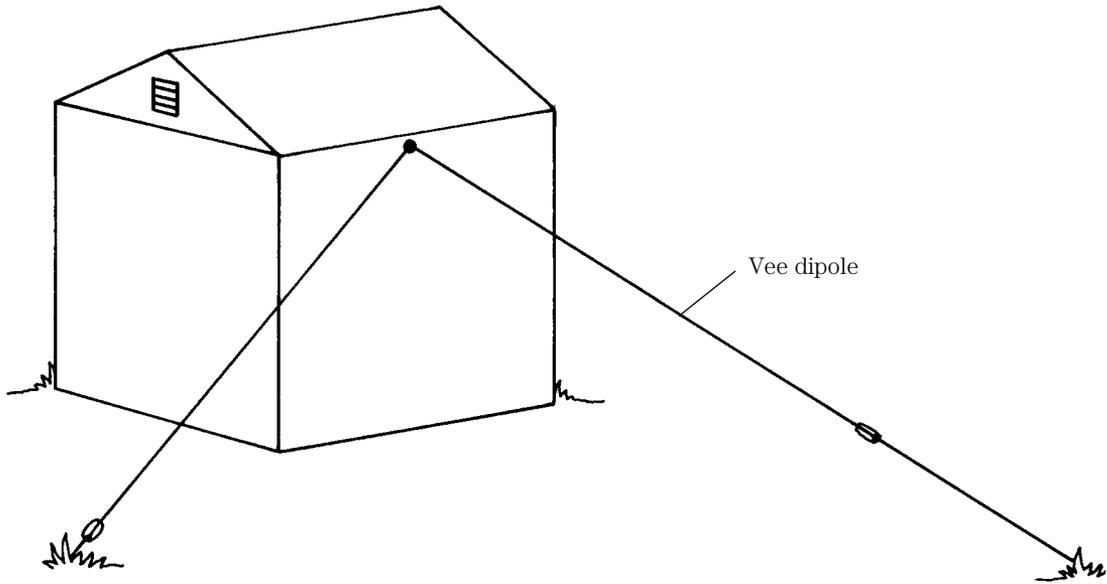
of a regular dipole), make  $A = D$  and  $B = C$ . In all three examples, Figs. 10-4A through 10-4C, you can expect to find the length needed for resonance varies somewhat from the standard  $468/F_{\text{MHz}}$  value, and that the feedpoint impedance is other than  $73 \Omega$ . Also, the pattern will be distorted with respect to the regular dipole. Although these antennas do not work as well as a properly installed dipole, the performance is sufficient to allow successful operation.

**10-4A** Sloper dipole.

Another limited-space wire antenna is the (so-called) half-slope shown in Fig. 10-4D. Although single-band versions are often seen, the example in Fig. 10-4D is a multiband version. Resonant traps separate the different band segments. This antenna operates like the vertical, but it is not omnidirectional. Also, the feedpoint impedance will be different from the regular vertical situation. If the impedance varies too much for comfort, insert an antenna tuner, such as the Transmatch, in the coaxial line between the transmitter and the antenna.

Figure 10-5 shows another antenna that's useful for limited-space situations. Although it is easily constructed from low-cost materials, the antenna is also sold by several companies under various rubrics including "cliff-dweller," "apartment house," "townhouse," or "traveler's" antennas. The antenna consists of a 4- to 16-ft section of aluminum or copper tubing. Some of the commercial antennas use a telescoping tubing that can be carried easily in luggage. As was true with the longwire, the window-sill antenna is tuned to resonance with an L-section coupler.

The L-section coupler must be tuned to produce the lowest possible VSWR, so either an RF power meter or a VSWR meter must be installed in the coaxial cable between the transmitter and the tuner. Again, a good ground (or radial system) will



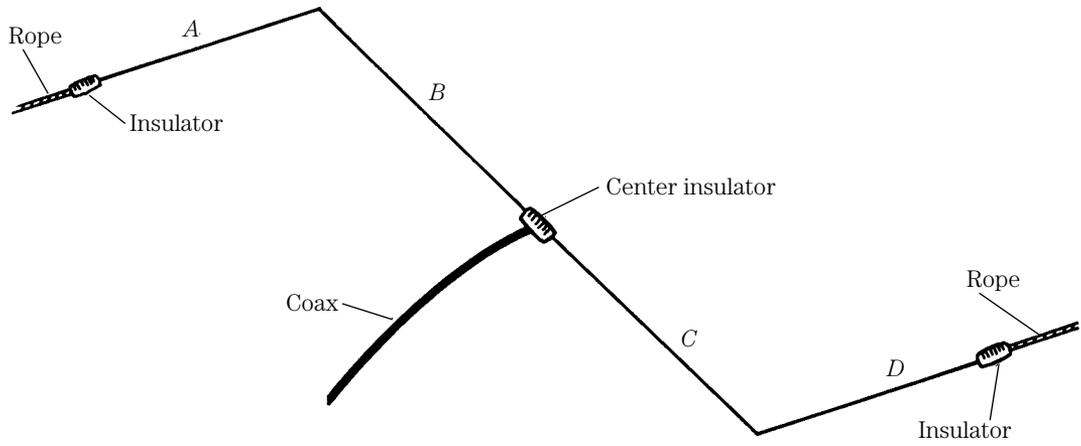
10-4B Vee dipole.

greatly improve the operation of this antenna. The performance should be considerably less than that of a good longwire, but it will work DX for you even on lower frequencies. Once again it is not the best antenna made, but it will get you on the air.

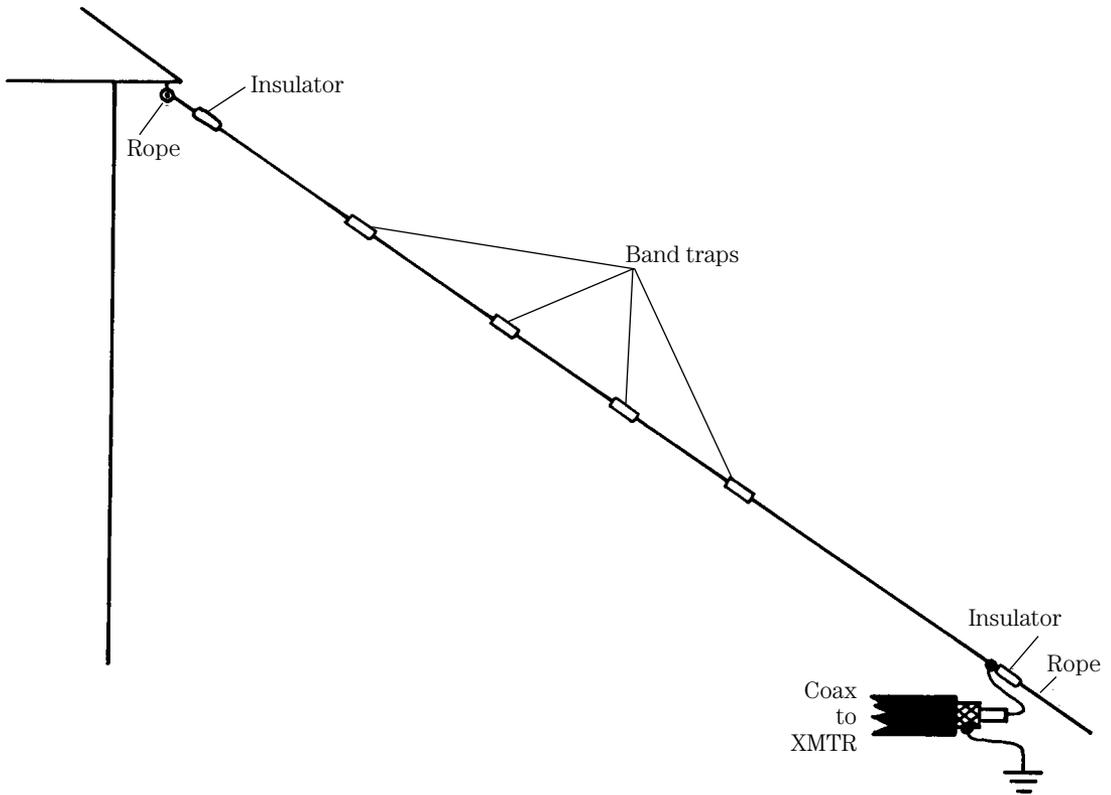
Figure 10-6 shows the use of a mobile antenna for a fixed or portable location. I used this type of antenna at one QTH to good effect. Mount a mobile antenna, such as the Hustler, on the window sill, or other convenient mount (at least one amateur radio operator uses one mounted on the roof of the house). Grounding is essential for this antenna, as was also true for the longwire and window-sill antenna (Fig. 10-5). The operation of this antenna is improved by installing at least two radials per band as a counterpoise ground. My installation worked well, even though only two radials were present.

A directional *rotatable* dipole is shown in Fig. 10-7. This antenna is made from a pair of mobile antennas connected “back-to-back” on a horizontal length of  $1 \times 2$  lumber and fed in the center with a coaxial cable. The end pieces of the mobile antenna set the resonance, and you must recognize that adjusting one requires a countervailing adjustment of the other as well.

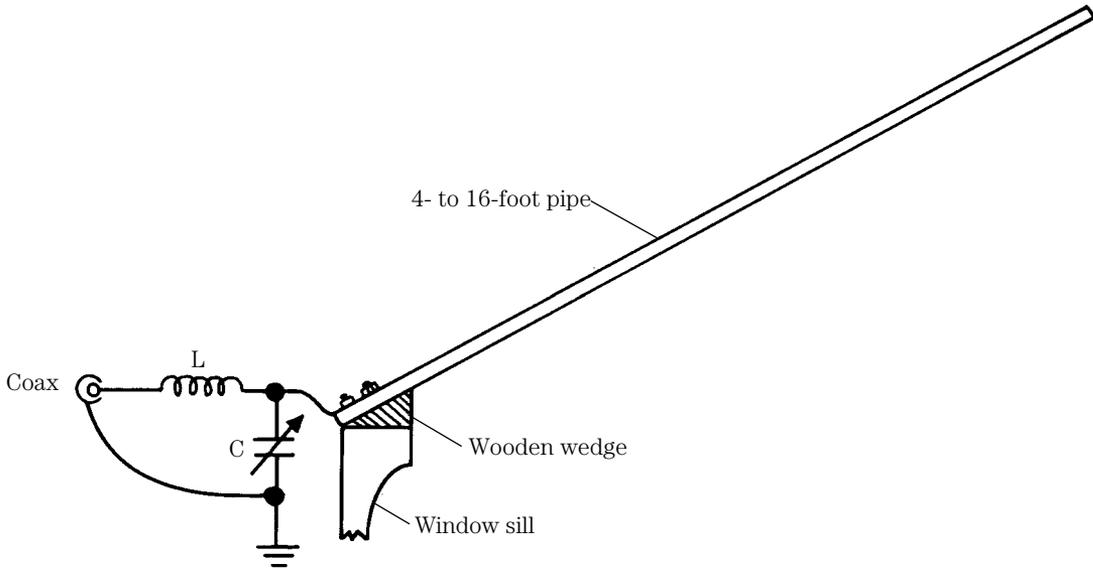
Two examples of *helically* wound antennas are shown in Fig. 10-8. In this type of antenna, an insulating mast is wound with a half-wavelength of antenna wire. The overall length of the antenna is considerably less than a half-wavelength, except at the highest frequencies. In order to dissipate the high voltages that tend to build up at the ends of these antennas, a capacitance hat is used. These “hats” are either disks (pie tins work well) or rods of conductor about 16 to 24 in long. The version shown in Fig. 10-8A is a vertical antenna, and like other verticals, it must be installed over either a good ground or a counterpoise ground. The version shown in Fig. 10-8B is horizontally polarized.



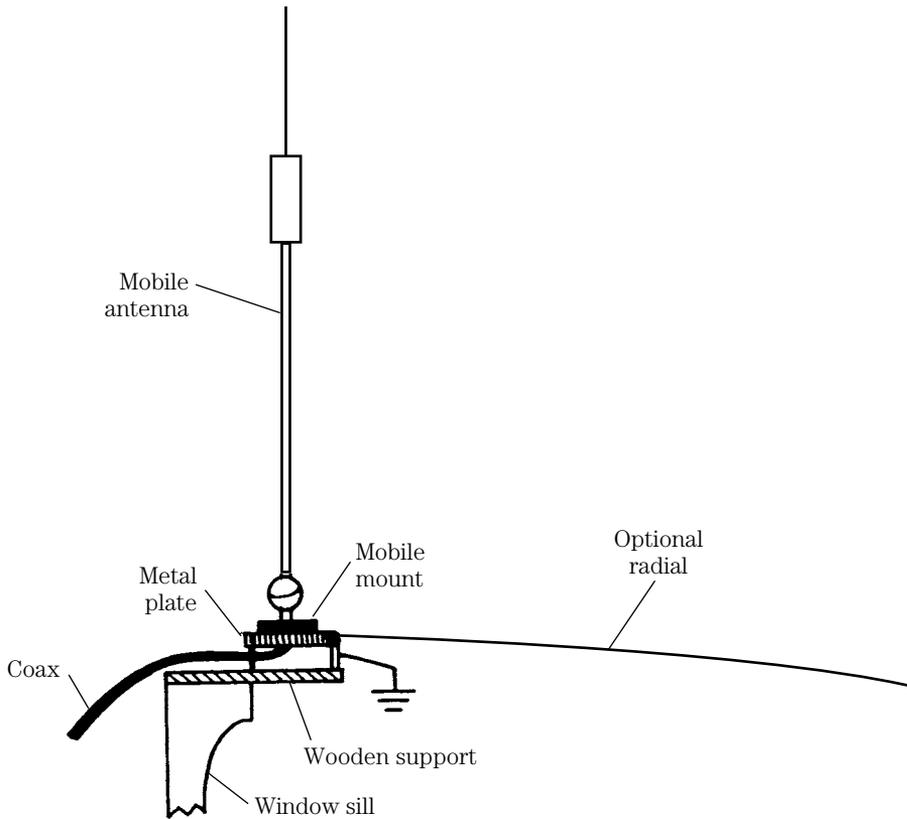
10-4C Bent dipole.



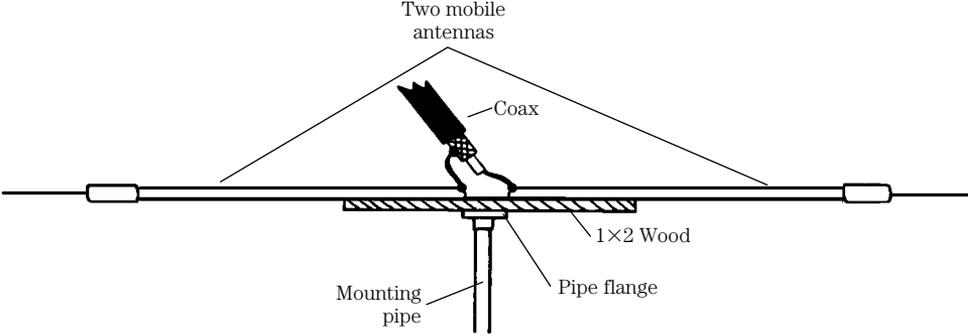
10-4D Sloper trap vertical.



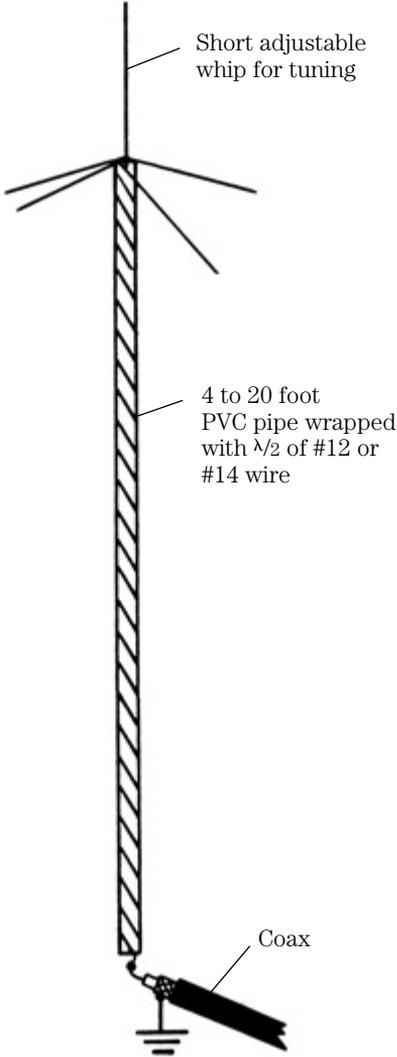
10-5 Window sill antenna.



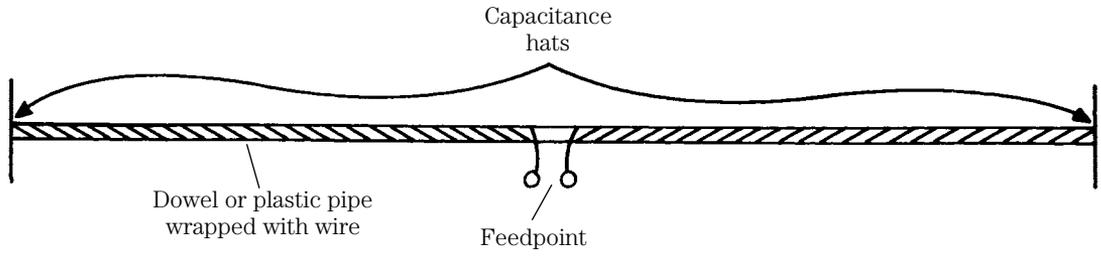
10-6 Use of mobile antenna as a window sill antenna.



**10-7** Use of two mobile antennas as a rotatable dipole.



**10-8A** Helically vertical antenna.



**10-8B** Helically wound dipole antenna.

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# 11

## CHAPTER

# Directional phased vertical antennas

THE VERTICAL ANTENNA IS A PERENNIAL FAVORITE WITH RADIO COMMUNICATIONS users. The vertical is either praised, or cursed, depending upon the luck of the owner. “DXability” is usually the criterion for judging the antenna’s quality. Some amateurs can’t get out of their backyards with a vertical, and they let everyone within earshot know that such and such a brand is no good. Yet, another person routinely works New Zealand or Australia on 15 m using exactly the same brand of vertical. The proper installation of vertical antennas is dealt with in another chapter, so, for the present, let’s look at another problem attributed to vertical antennas.

That problem is that vertical antennas are *omnidirectional* in the azimuth aspect; that is, they send out and receive equally well from all directions. Some people moan that this pattern dissipates their power, and gives them a weaker signal “out where it counts” (true). However, the main disadvantage of the omnidirectional pattern is noise (QRN and QRM). “QRN” is natural noise from thunderstorms and other sources. “QRM” is man-made noise, and can consist of other stations or the many assorted forms of electrical filth that pollute the airwaves. All forms of noise, however, have one thing in common: they are directional with respect to the station. In other words, if you could null signals coming from the direction of the noise source (or undesired station), you would be able to hear desired stations much better. A directional antenna performs this task, so let’s look at some vertically polarized directional antennas.

Although most amateurs seem to think that the effective radiated power (ERP) increase that the directional antenna gives them is the real reason to own one, the main benefit is actually on receive. Think about it for a moment. With anywhere from 100 to 1500 W available, the increase or decrease in signal strength (due to the directivity of the antenna) results in a minimal difference on the receive end, especially during good DX conditions. If we rotate the directional pattern, to null out interference, then we usually find that the change in our signal strength perceived by the other guy is small; the S meter reading of the desired station is minimally affected; but the amplitude of the interference source is greatly attenuated! The overall effect is an apparent increase

in the other guy's signal, even though the S meter tells a slightly different story. The improvement of signal-to-noise ratio (SNR) is tremendously improved.

## Directivity and phasing

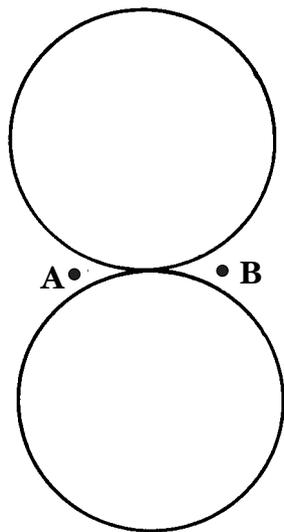
So, how does a vertical antenna owner get the benefit of directivity without the kilobuck investment that a beam or quad costs? The usual solution is to use phased verticals. AM broadcast stations, with more than one tower, are using this type of system (although for different reasons than hams). The idea is to place two or more antennas in close proximity and feed them at specific phase angles to produce a desired radiation pattern. A lot of material is available in the literature on phased vertical antenna systems, and it is far too much to be reproduced here. There are “standard patterns” dating from before World War II that are created with different spacings and different phase angles of feed current. In this chapter, we will consider only one system.

Figure 11-1 shows the patterns for a pair of quarter-wavelength vertical antennas spaced a half-wavelength ( $180^\circ$ ) apart. Without getting into complex phase shifting networks, there are basically two phasings that are easily obtained:  $0^\circ$  (antennas in phase) and  $180^\circ$  (antennas out of phase with each other).

When the two antennas (*A* and *B*) are fed in phase with equal currents (Fig. 11-1A), the radiation pattern (shown somewhat idealized here) is a bidirectional figure 8 that is directionally perpendicular to the line of centers between the two antennas; this pattern is called a *broadside* pattern. A sharp null exists along the line of centers (*A-B*).

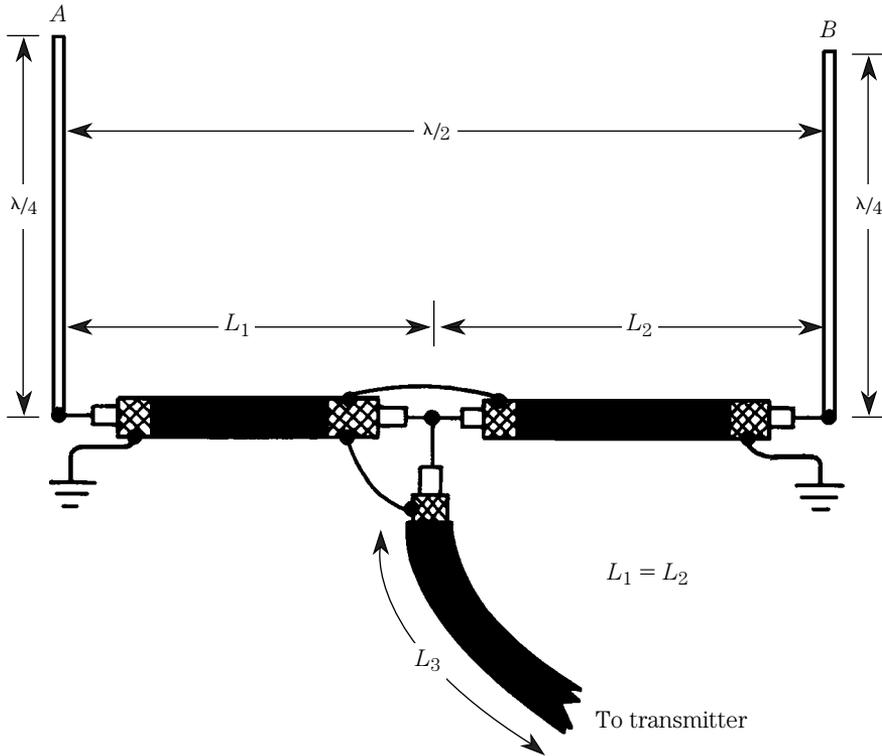
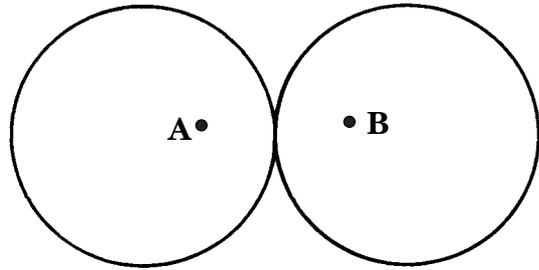
When the antennas are fed out of phase with each other by  $180^\circ$  (Fig. 11-1B), the pattern rotates  $90^\circ$  (a quarter way around the compass) and now exhibits directivity along the line of the centers (*A-B*); this is the “end fire” pattern. The interference cancelling null is now perpendicular to line *A-B*.

It should be apparent that you can select your directivity by selecting the phase angle of the feed currents in the two antennas. Figure 11-2 shows the two feeding



**11-1A** Pattern of two radiators in-phase, spaced a half-wavelength apart.

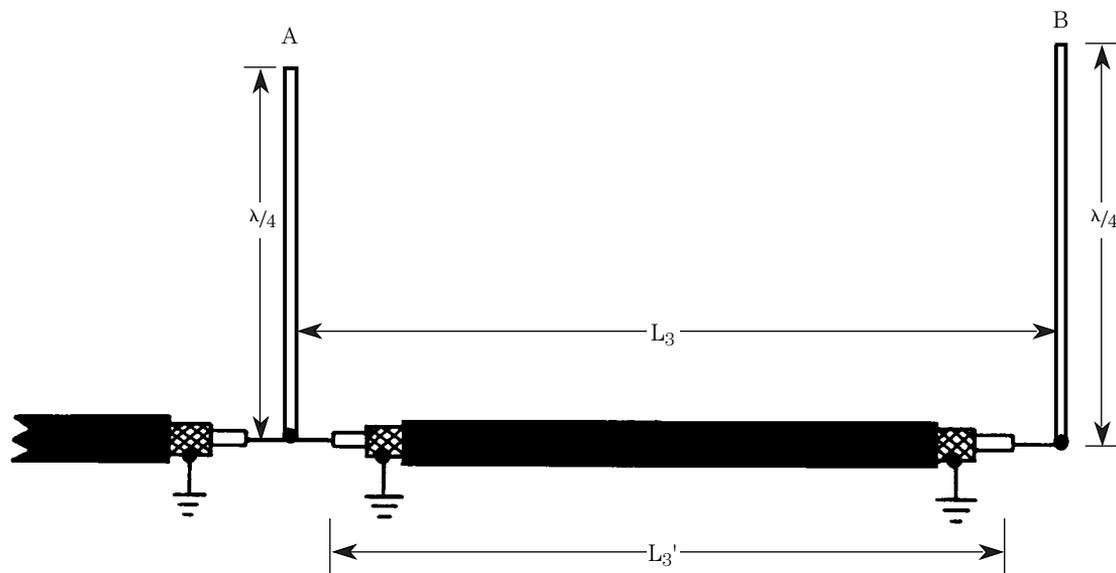
11-1B Pattern of two radiators out of phase, spaced a half-wavelength apart.



11-2A Feeding a phased array antenna in phase.

systems usually cited for in-phase (Fig. 11-2A) and out-of-phase (Fig. 11-2B) systems. Figure 11-2A shows the coax from the transmitter coming to a coax tee connector. From the connector to the antenna feedpoints are two lengths of coax ( $L_1$  and  $L_2$ ) that are equal to each other, and identical. Given the variation between coaxial cables, I suspect that it would work better if the two cables were not merely the same length ( $L_1 = L_2$ ), but also that they came from the same roll.

The second variation, shown in Fig. 11-2B, supposedly produces a  $180^\circ$  phase shift between antenna A and antenna B, when length  $L_3$  is an electrical half-wavelength. According to a much-publicized theory, the system of Fig. 11-2B ought to



**11-2B** Feeding a phased array antenna out of phase.

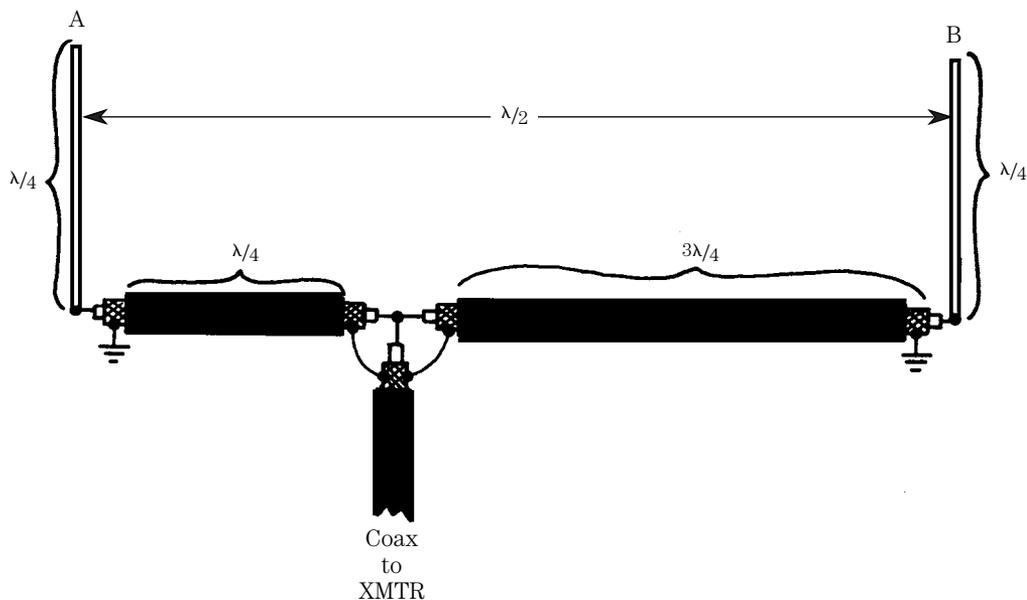
produce the pattern of Fig. 11-1B—yet experience shows this claim is false. It seems that there are several problems with the system in Fig. 11-2B.

First, coax has a property called *velocity factor* ( $V_F$ ), which is the fraction of the speed of light at which signals in the cable propagate. The  $V_F$  is a decimal fraction on the order of 0.66 to 0.90, depending upon the type of coax used. Unfortunately, the physical spacing between *A* and *B* is a real half-wavelength ( $L_3 = 492/F$ ), but the cable length is shorter by the velocity factor [ $L_3' = (V_F \times 492) / F$ ].

Consider an example. A 15-m phased vertical antenna system will have two 11-ft radiators, spaced 22 ft apart (approximately, depending upon exact frequency). If we use foam coax, with  $V_F = 0.80$ , the cable length is  $0.8 \times 22$  ft, or 17.6 ft. In other words, despite lots of publicity, the cable won't fit between the towers!

Second, the patterns shown in Fig. 11-1 are dependent upon one condition: the *antenna currents are equal*. If both of them are the same impedance, and are fed from the same transmitter, then it is reasonable to assume that the currents are equal—right? No, wrong! What about coax loss? Because of normal coax loss, which increases at higher frequencies, the power available to antenna *B* in Fig. 11-1B is less than the power available to antenna *A*. Thus, the pattern will be somewhat distorted, because the current produced in *B* is less than the current in *A*, when they should be equal.

The first problem is sometimes fixed by using unequal lengths for cables  $L_1$  and  $L_2$  (Fig. 11-2A), and using it for the out-of-phase case. For example, if we make  $L_1$  one-quarter wavelength and  $L_2$  three-quarter wavelength (Fig. 11-2C), antenna *A* is fed with a  $90^\circ$  phase lag (relative to the tee connector signal), while antenna *B* is fed with a  $270^\circ$  phase shift. The result is still a  $180^\circ$  phase difference. Unfortunately, we



11-2C Corrected feed for a phased array antenna.

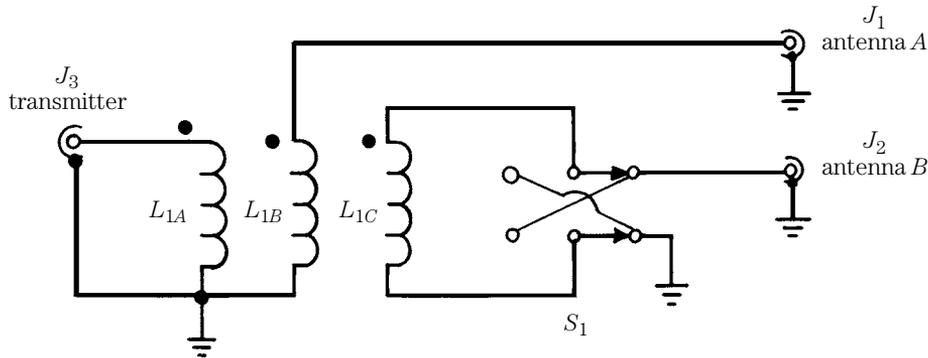
have not solved the current level problem, and may have actually made it worse by adding still more lossy cable to the system.

There is still another problem that is generic to the whole class of phased verticals. Once installed, the pattern is fixed. This problem doesn't bother most point-to-point commercial stations, or broadcasters, because they tend to transmit in only one direction. But amateurs are likely to need a rotatable pattern. Neither the antennas in Fig. 11-1A nor that in Fig. 11-1B is rotatable without a lot of effort—like changing the coax feeds, or physically digging up the verticals and repositioning them.

Fortunately, there is a single solution to all three problems. Figure 11-3 shows a two-port phasing transformer made from a toroidal balun kit. Use the kind of kit that makes a 1:1 balun transformer. Although we are not making a balun, we will need enough wire to make three windings, and that is the normal case for 1:1 baluns. Amidon Associates and others make toroidal balun kits.

Wind the three coils in trifilar style, according to the kit instructions. The dots in Fig. 11-3 show the “sense” of the coils, and they are important for correct phasing; call one end the “dot end” and the other end the “plain end” to keep them separate. If the dot end of the first coil is connected to  $J_3$  (and the transmitter), then connect the dot end of the second coil to the  $0^\circ$  output ( $J_1$ , which goes to antenna A). The third coil is connected to a DPDT RF relay or switch. In the position shown,  $S_1$  causes the antennas to be  $180^\circ$  out of phase. In the other position, the “sense” of the third coil is reversed, so the antennas are in phase.

Another phasing method is shown in Fig. 11-4. In this scheme, two convenient, but equal, lengths of coaxial cable ( $L_1$  and  $L_2$ ) are used to carry RF power to the antennas. One segment ( $L_1$ ) is fed directly from the transmitter's coaxial cable ( $L_3$ ), while the other is fed from a phasing switch. The phasing switch is used to either by-



11-3A Phasing transformer circuit.

pass or insert a phase-shifting length of coaxial cable ( $L_4$ ). For  $180^\circ$  phasing use the following equation to find the length ( $L_4$ ):

$$L = \frac{492 V_F}{F_{\text{MHz}}} \text{ ft} \quad [11.1]$$

where

$L$  is the length of  $L_4$ , in feet

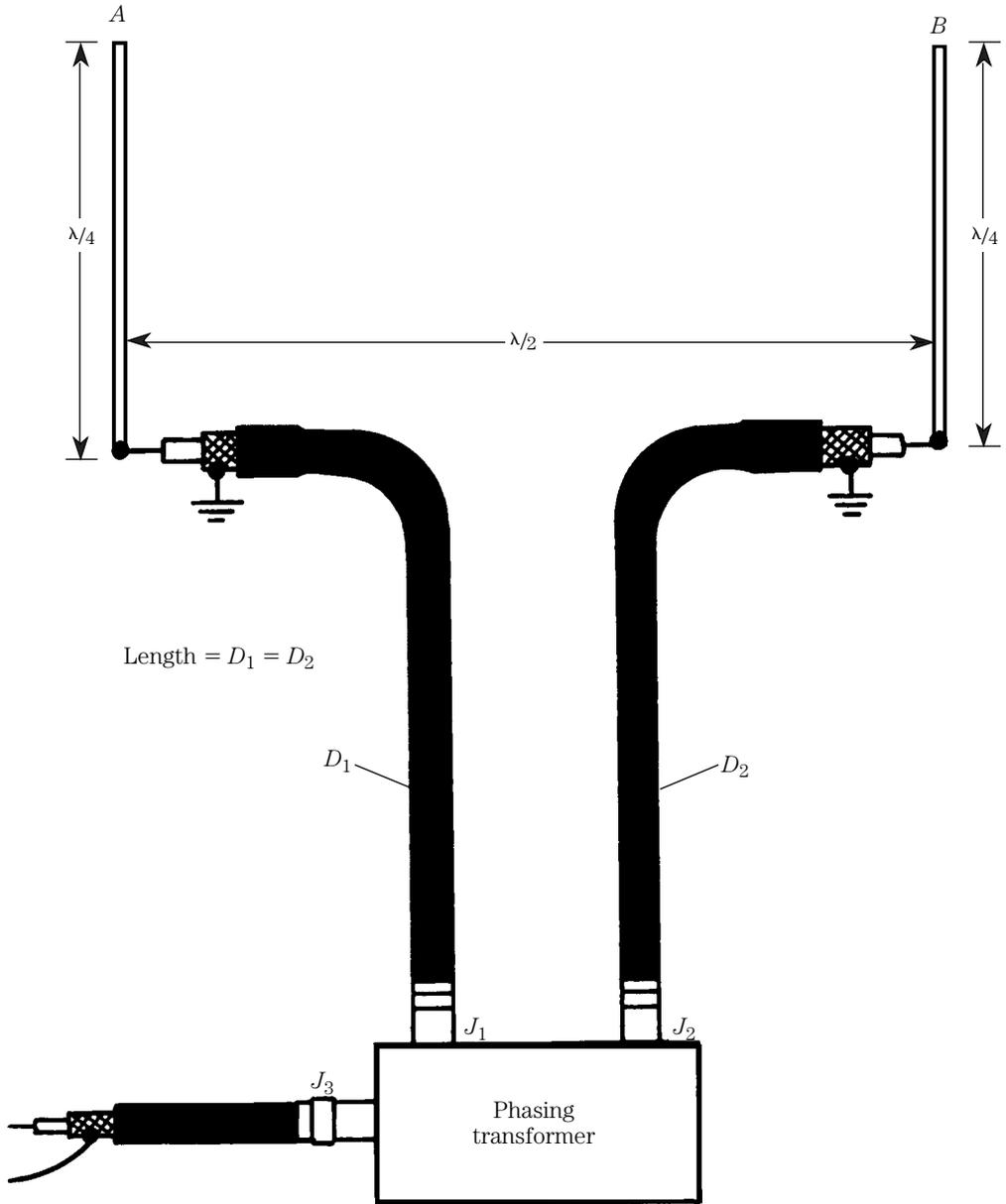
$V_F$  is the velocity factor (a decimal fraction)

$F_{\text{MHz}}$  is the operating frequency, in megahertz

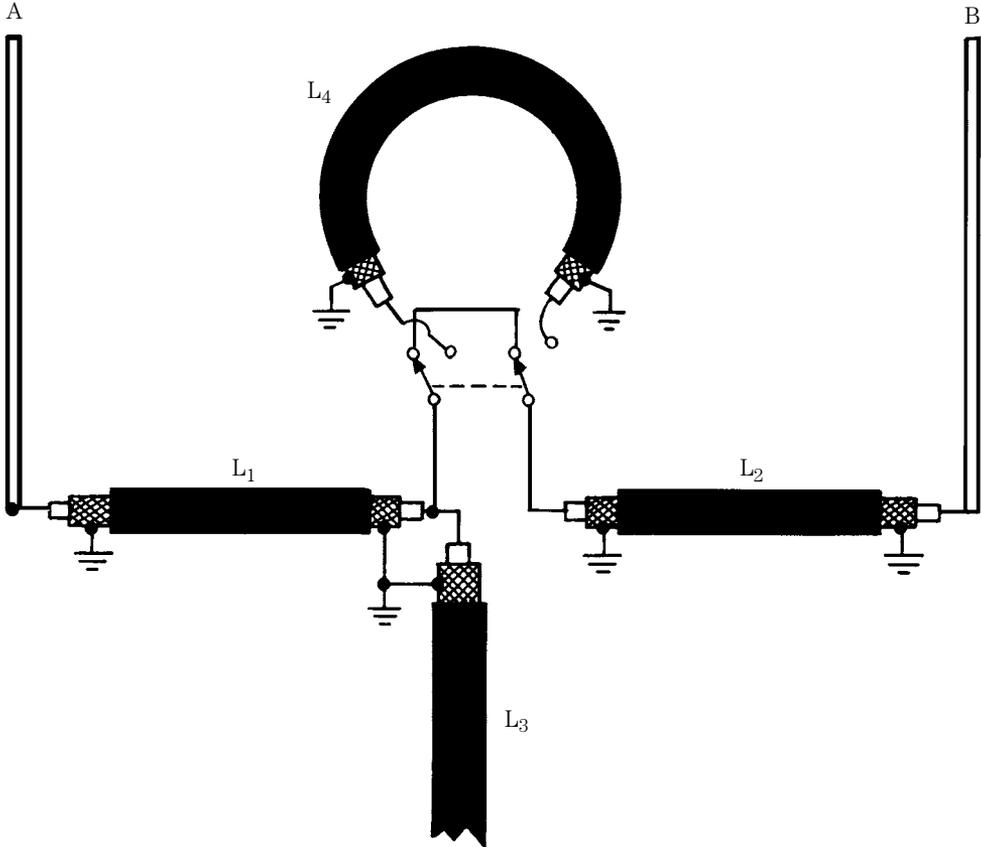
Some people use a series of switches to select varying amounts of phasing shift from  $45^\circ$  to  $270^\circ$ . Such a switch allows them to select any number of other patterns for special situations.

## 360° directional array

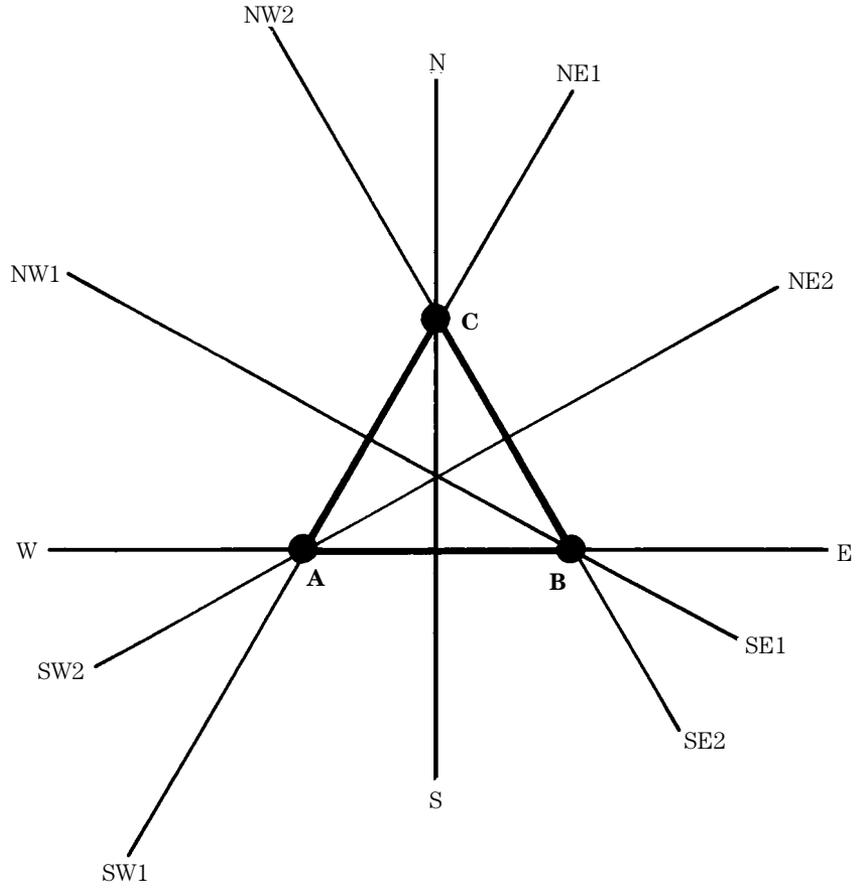
The phased vertical antenna concept can be used to provide round-the-compass control of the antenna pattern. Figure 11-5A shows how three quarter-wavelength verticals (arranged in a triangle that is a half-wavelength on each side) can be used to provide either end-fire or broadside patterns from any pair ( $A-B$ ,  $A-C$ , or  $B-C$ ). Any given antenna ( $A$ ,  $B$ , or  $C$ ) will be grounded, fed at  $0^\circ$ , or fed with  $180^\circ$ . The table in Fig. 11-5B shows the relative phasing for each direction that was labelled in Fig. 11-5A. Either manual phase changing or switch-operated phase changing can be used, although the latter is preferred for convenience. Some international showcase broadcasters use antenna arrays formed into two or more concentric circles of vertical elements, with one element at the center. Selection of elements and phasing determines directivity and gain.



11-3B Connection to antennas.



11-4 Phase-shifting antenna circuit.



11-5A Three-element phased array.

Direction	Ant A	Ant B	Ant C
N-S	0°	0°	GND
NE1-SW1	0°	GND	180°
NE2-SW2	GND	0°	0°
E-W	0°	180°	GND
SE1-NW1	0°	GND	0°
SE2-NW2	GND	0°	180°

**11-5B** Table of feed phasing for the three-element array. Ant = antenna. GND = ground

## 12 CHAPTER

# Directional beam antennas

THE DIRECTIONAL BEAM ANTENNA DOES SEVERAL JOBS. FIRST, IT PROVIDES AN APPARENT increase in radiated power, because it focuses available transmitter power into a single (or at worst limited) direction. For this reason, a bidirectional dipole has a gain of approximately 2 dB over an isotropic radiator. Add one or more additional elements, and the focusing becomes nearly unidirectional, which increases the *effective radiated power* (ERP) even more. Second, the beam increases the received signal available at the inputs of the receiver. Antennas are generally reciprocal, so they will work for receiving as they do for transmitting. Finally, the directivity of the beam antenna allows the operator to null interfering stations. In fact, it is the last attribute of the beam that is most useful on today's crowded bands. All in all, if your funds are too little to provide both increased RF power and a good antenna system, then spend what is available on the antenna—not on the power.

In this chapter we will focus on directional antennas that can be built relatively easily. It is assumed that most readers who want a triband multielement Yagi will prefer to buy a commercial product, rather than build a homebrew model. The material herein concentrates on homebrew projects that are within the reach and capabilities of most readers. The first of these is not a beam antenna at all, but rather a *rotatable dipole*.

## Rotatable dipole

The *dipole* is a bidirectional antenna with a figure-8 pattern (when viewed from above). The dipole is a half-wavelength and is usually installed horizontally, although vertical half-wavelength dipoles are known. Although the length of the dipole is too great for rotatability at the lower bands, it is within reason for the higher band. For example, the size of the halfwave dipole is approximately 16 ft on 10 m and 22 ft on 15 m. Even the 33-ft length on 20 m is not unreasonable for amateur constructors. The length of the dipole is found from

$$L = \frac{468}{F_{\text{MHz}}} \text{ ft} \quad [12.1]$$

This length is approximate because of end effects and other phenomena, so some “cut and try” is required.

**Example 12-1** Find the length of a dipole antenna for a frequency of 24.930 MHz in the 12-m amateur radio band.

**Solution:**

$$\begin{aligned} L &= \frac{468}{F_{\text{MHz}}} \text{ ft} \\ &= \frac{468}{24.930 \text{ MHz}} = 18.77 \text{ ft} \end{aligned}$$

The half-wave dipole is fed in the center by coaxial cable. Each element of the dipole is one-half of the overall length (or, in the example given, about 9.4 ft).

Figure 12-1 shows a rotatable dipole that can be designed for use on 15, 12, and 10 m. The radiator elements are made from 10-ft lengths of  $\frac{3}{4}$ -in aluminum tubing. The tubing is mounted on “beehive” standoff insulators, which in turn are mounted on a 4-ft length of  $2 \times 2$  lumber. The lumber should be varnished against weathering. In a real pinch, the elements can be mounted directly to the lumber without the insulators, but this is not the recommended practice.

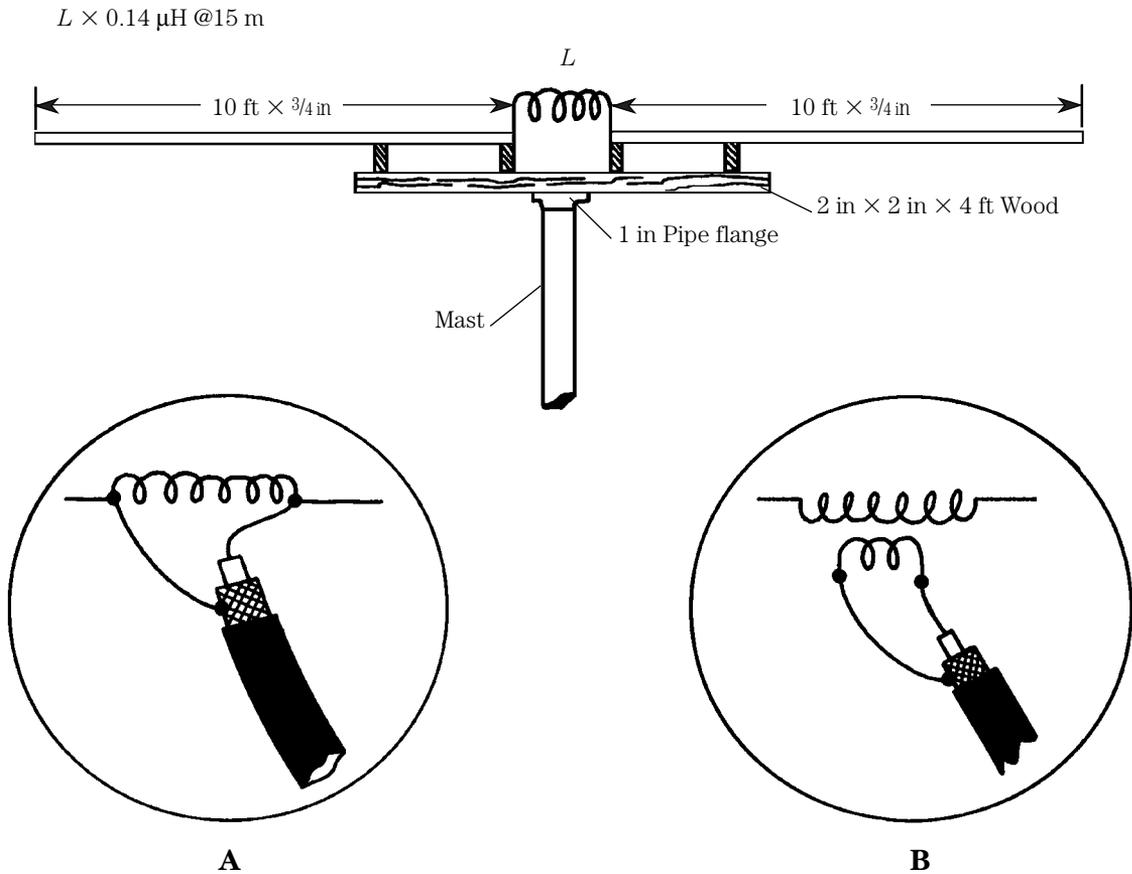
The mast is attached to the  $2 \times 2$  lumber through any of several means. The preferred method is the use of a 1-in pipe flange. These devices are available at hardware stores under the names *floor flange* and *right-angle flange*.

The 10-ft lengths of pipe are the standard lengths available in hardware stores, so it was selected as being closest to the required 22 ft for 15 m. A 0.14- $\mu\text{H}$  loading coil is used at the center, between the elements, in order to make up for the short length. The dimensions of the coil are 4 to 5 turns, 0.5-in diameter, 4-in length. For low power levels, the coil can be made of no. 10 (or no. 12) solid wire—and, for higher levels,  $\frac{1}{8}$ -in copper tubing.

There are two basic ways to feed the antenna, and these are shown in details A and B in Fig. 12-1. The traditional method is to connect the coaxial cable (in parallel) across the inductor. This method is shown in Fig. 12-1, detail A. A second method is to link couple the coil to the line through a one- to three-turn loop (as needed for impedance matching). This is the method that would be used for a toroidal inductor.

Lower frequencies can be accommodated by changing the dimensions of the coil. The coil cannot be scaled, simply because the relative length of the antenna changes as the frequency changes. But it is possible to cut and try by adding turns to the coil, one turn at a time, and remeasuring the resonant frequency. Adding inductance to the coil will make the antenna usable on 17 m and 20 m, as well as on 15 m.

Another method for building a rotatable dipole for lower frequencies is to increase the element lengths. On 17 m, the overall length is approximately 27.4 ft, so each element length is 13.7 ft long. This length can be achieved by either of two methods. First, adjacent sizes of aluminum tubing are designed so that the smaller will be a slip-fit inside of the larger. What constitutes “adjacent sizes” depends on the wall thickness, but for one common brand, the  $\frac{7}{8}$ -in is adjacent to the  $\frac{3}{4}$ -in size. You



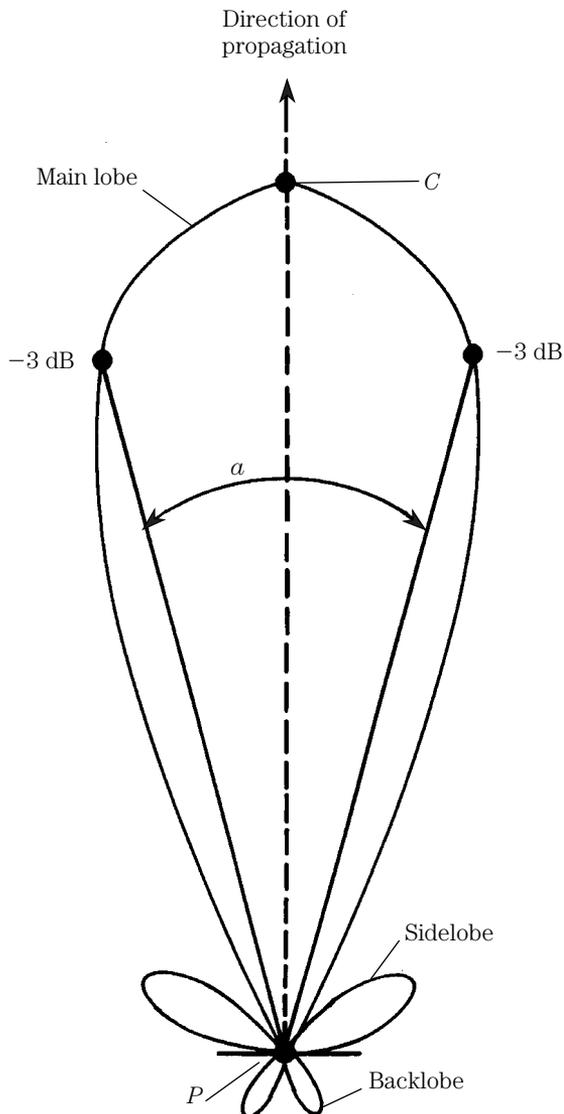
12-1 Rotatable dipole antenna. Inset A shows conventional feed; inset B shows transformer feed.

can use two smaller lengths to make the larger lengths of pipe, and cut it to size. This method is only available to those readers who have a commercial or industrial metals distributor nearby, because the 16-ft lengths are not generally available from hardware stores.

Bands higher than 15 m (i.e., 12 and 10 m) can be accommodated by using the 10-ft lengths of tubing, but without the inductor. The tubing is cut to the desired half-wavelength size and used directly.

## Yagi beam antennas

A *Yagi* antenna is one member of a class of directional beam antennas that are popular in the HF bands. Figure 12-2 shows the pattern (viewed from above) typical of the beam antenna. The antenna is located at point  $P$  and fires signals in the direction shown by the arrow. The beamwidth of the antenna is the angle  $a$  between the points on the main lobe that are  $-3$  dB down from the center point  $C$ .



12-2 Pattern of beam antenna.

A perfect beam antenna will have only the main lobe, but that situation occurs only in dreams. All real antennas have both sidelobes and backlobes, also shown in Fig. 12-2. These lobes represent wasted power transmitted in the wrong direction during transmission and interference opportunities while receiving. The goal of the antenna designer is to increase the main lobe while decreasing the sidelobes and backlobes.

Figure 12-3 shows, schematically, the basic *Yagi-Uda* antenna (usually called simply *Yagi*). The driven element is a simple half-wavelength dipole fed in the center. There are two additional elements: *reflectors* and *directors*. These are called *parasitic elements* because they are not directly excited by RF, but rather,

they receive energy radiated from the driven element and then reradiate it. The reflector is placed behind the driven element, and is typically about 4 percent longer than the driven element. The director is placed in front of the driven element (relative to the direction of propagation). The director is typically about 4 percent shorter than the driven element. Although there is no fixed rule regarding the number of either reflectors or directors, it is common practice to use a single director and a driven element for two-element beams, and a single reflector and a single director, in addition to the driven element, for three-element beams. Again, additional reflectors can be added for four- and more element beams, but standard practice calls for addition of directors instead.

The length of the elements is given by

$$L = \frac{K}{F_{\text{MHz}}} \text{ ft} \tag{12.2}$$

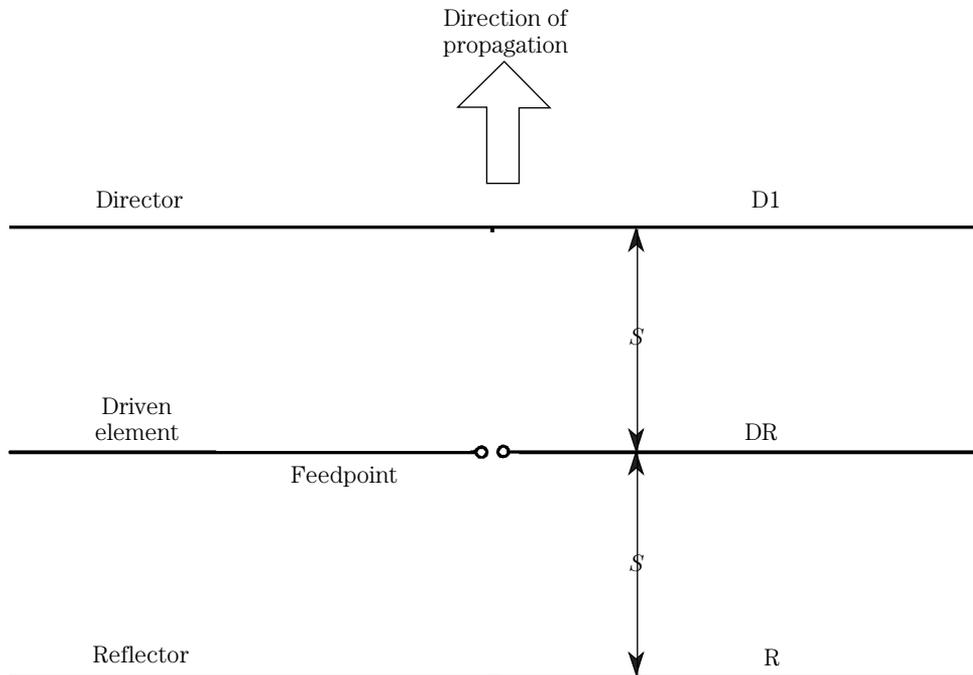
where

$L$  is the length, in feet

$F_{\text{MHz}}$  is the frequency, in megahertz

$K$  is a constant

The spacing of the elements is typically from 0.15 to 0.308 wavelengths, although 0.2 and 0.25 are the most common values.



12-3 Basic Yagi-Uda antenna.

**Example** Calculate the approximate element lengths for a three-element 15-m beam designed to operate on a frequency of 21.39 MHz.

**Solution:**

$$\begin{aligned}
 1. \text{ Driven element } (K = 478): \quad L &= \frac{K}{F_{\text{MHz}}} \text{ ft} \\
 &= \frac{478}{21.39 \text{ MHz}} \text{ ft} \\
 &= 22.34 \text{ ft}
 \end{aligned}$$

$$\begin{aligned}
 2. \text{ Reflector } (K = 492): \quad L &= \frac{K}{F_{\text{MHz}}} \text{ ft} \\
 &= \frac{492}{21.39 \text{ MHz}} \text{ ft} \\
 &= 23 \text{ ft}
 \end{aligned}$$

$$\begin{aligned}
 3. \text{ Director } (K = 461.5): \quad L &= \frac{K}{F_{\text{MHz}}} \text{ ft} \\
 &= \frac{461.5}{21.39 \text{ MHz}} \text{ ft} \\
 &= 21.58 \text{ ft}
 \end{aligned}$$

$$\begin{aligned}
 4. \text{ Element Spacing } (K = 142): \quad L &= \frac{K}{F_{\text{MHz}}} \text{ ft} \\
 &= \frac{142}{21.39 \text{ MHz}} \text{ ft} \\
 &= 6.64 \text{ ft}
 \end{aligned}$$

The elements of a rotatable beam antenna can be built in a manner similar to the rotatable dipole described earlier. In the beam antenna, however, a boom is needed between the elements to support them. The boom can be made of metal or wood. In the case of a metal boom, the driven element must be insulated from the boom, even though the parasitic elements can be mounted directly to it. In general, it is usually better to use wood as a matter of convenience. Metal boom antennas can be obtained from commercial sources. The wood boom is easy to build and maintain, even though a little less durable than a metal boom.

The feedpoint impedance of a dipole is on the order of  $72 \Omega$  in free space, although the actual impedance will vary above and below that figure for antennas close to the earth's surface. In addition, adding parasitic elements reduces the impedance even more. The feedpoint impedance of the antenna is too low to be directly fed with coaxial cable, so some means of impedance matching is needed. Some people feed the antenna through an impedance matching balun transformer. Figure 12-4 shows the *gamma match* system. The driven element of the Yagi is not

broken in the center, as in the case of the simple dipole. The outer conductor, or shield, of the coaxial cable is connected to the center point of the driven element. The center conductor is connected to the gamma match element. The dimensions of the gamma match are

1. Gamma match length:  $L/10$
2. Gamma match director:  $D/3$
3. Spacing of gamma match from driven element:  $L/70$

where

$L$  is the length of the driven element

$D$  is the diameter of the driven element

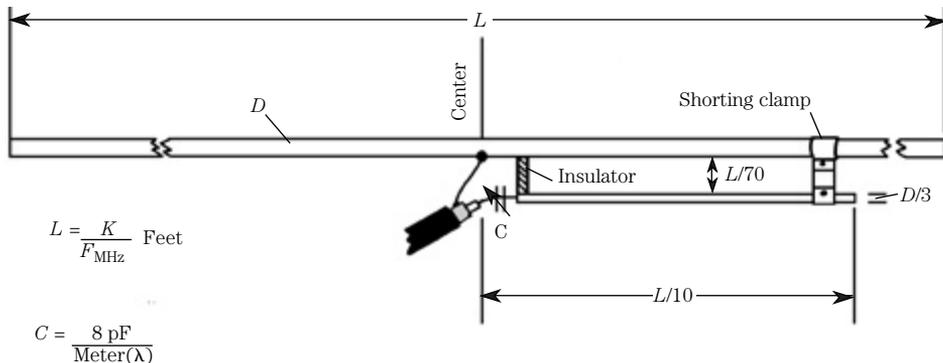
The capacitor in series with the center conductor of the coaxial cable has a value of approximately 8 pF per meter of wavelength at the lowest frequency in the band of operation, or approximately,

$$C = \frac{2400}{F_{\text{MHz}}} \text{ pF} \quad [12.3]$$

The capacitor must be a high-voltage transmitting variable type. In general, the gamma match capacitors are either air or vacuum variables.

There are three aspects to the adjustment of the Yagi antenna. Resonance is determined by the length of the element. The length is increased or decreased in order to find the resonant point. This point can be determined by the use of a noise bridge, VSWR meter or other means. The capacitor, and the shorting bar/clamp, are adjusted to match the impedance of the antenna to the transmission line impedance. For the dimensions shown, the coaxial cable should be 52  $\Omega$  (RG-58 or RG-8).

It is not necessary to use tubing or pipes for the antenna elements in order to obtain the benefits of the Yagi beam antenna. An example of a wire beam is shown in Fig. 12-5. The wire beam is made as if it were two half-wavelength dipoles, installed parallel to and about 0.2 to 0.25 wavelengths apart from each other. Although multielement wire beams are possible, the two-element version is the most common. Perhaps the most frequent



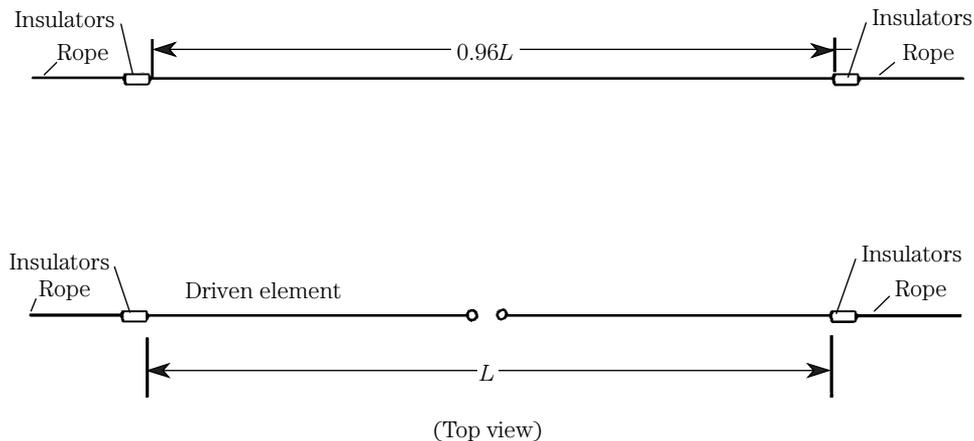
12-4 Gamma feed provides impedance matching.

use of the wire beam antenna is on the lower bands (e.g., 40 m and 75/80 m) where rotatable beams are more difficult to build.

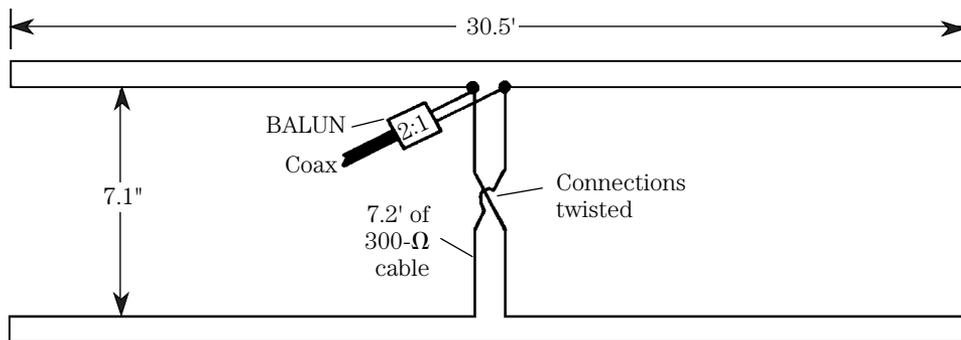
## 20-m ZL-special beam

The antenna shown in Fig. 12-6 is a close relative of the Yagi beam. It consists of a pair of folded dipoles, mounted approximately 0.12 wavelengths apart. The elements are 30.5 ft in length, and the spacing is 7.1 ft. The elements can be built from aluminum tubing if the antenna is to be rotatable. Alternatively, for a fixed antenna, the elements can be made of 300-Ω television-type twin lead. If the tubing type of construction is selected, then make the size of the tubing, and its spacing, sufficient for 300-Ω parallel transmission lines, according to

$$Z_o = 276 \log \frac{2S}{d} \quad [12.4]$$



12-5 Wire beam useful for low frequencies.



12-6 ZL-special beam antenna.

where

$Z_o$  is the impedance of the line (300  $\Omega$ )

$S$  is the center-to-center spacing of the parallel conductors on the elements

$d$  is the diameter of the conductors

The two half-wavelength elements of the ZL-special are fed  $135^\circ$  out of phase with each other. The feedline is connected to one of the dipoles directly; and then to the other through a length of 300- $\Omega$  twin lead that has an electrical length of about  $45^\circ$  ( $\lambda/8$ ).

The feedpoint impedance is on the order of 100 to 150  $\Omega$ , so it will make a good match to either 52- or 75- $\Omega$  coaxial cable if a 2:1 impedance matching transformer is used.

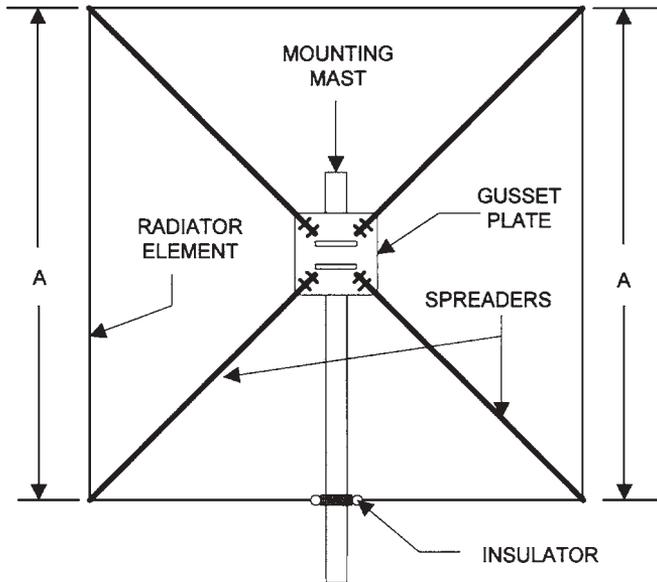
## Cubical quad beam antenna

The *cubical quad* antenna is a one-wavelength square wire loop. It was designed in the mid-1940s at radio station HCJB in Quito, Ecuador. HCJB is a Protestant missionary shortwave radio station with worldwide coverage. The location of the station is at a high altitude. This fact makes the Yagi antenna less useful than it is at lower altitudes. According to the story, HCJB originally used Yagi antennas. These antennas are fed in the center at a current loop, so the ends are high-voltage loops. In the thin air of Quito, the high voltage at the ends caused corona arcing, and that arcing periodically destroyed the tips of the Yagi elements. Station engineer Clarence Moore designed the cubical quad antenna (Fig. 12-7) to solve this problem. Because it is a full-wavelength antenna, each side being a quarter wavelength, and fed at a current loop in the center of one side, the voltage loops occur in the middle of the adjacent sides—and that reduces or eliminates the arcing. The elements can be fed in the center of a horizontal side (Figs. 12-7A and 12-8A), in the center of a vertical side (Fig. 12-8B), or at the corner (Fig. 12-8C).

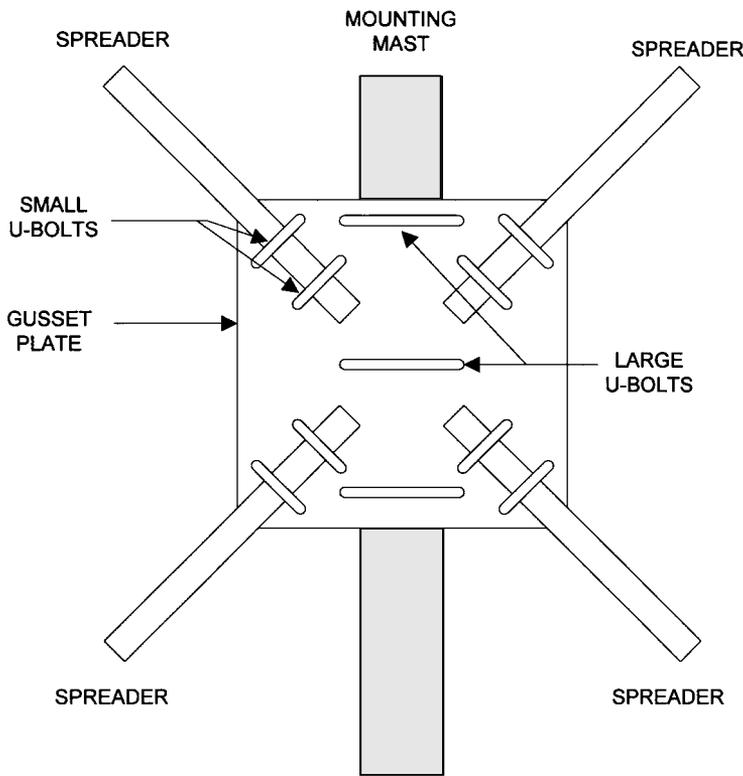
The antenna shown in Fig. 12-7A is actually a quad loop rather than a cubical quad. Two or more quad loops, only one of which needs to be fed by the coax, are used to make a cubical quad antenna. If only this one element is used, then the antenna will have a figure-8 azimuthal radiation pattern (similar to a dipole). The quad loop antenna is preferred by many people over a dipole for two reasons. First, the quad loop has a smaller “footprint” because it is only a quarter-wavelength on each side ( $A$  in Fig. 12-7A). Second, the loop form makes it somewhat less susceptible to local electromagnetic interference (EMI).

The quad loop antenna (and the elements of a cubical quad beam) is mounted to spreaders connected to a square gusset plate. At one time, carpets were wrapped around bamboo stalks, and those could be used for quad antennas. Those days are gone, however, and today it is necessary to buy fiberglass quad spreaders. A number of kits are advertised in ham radio magazines.

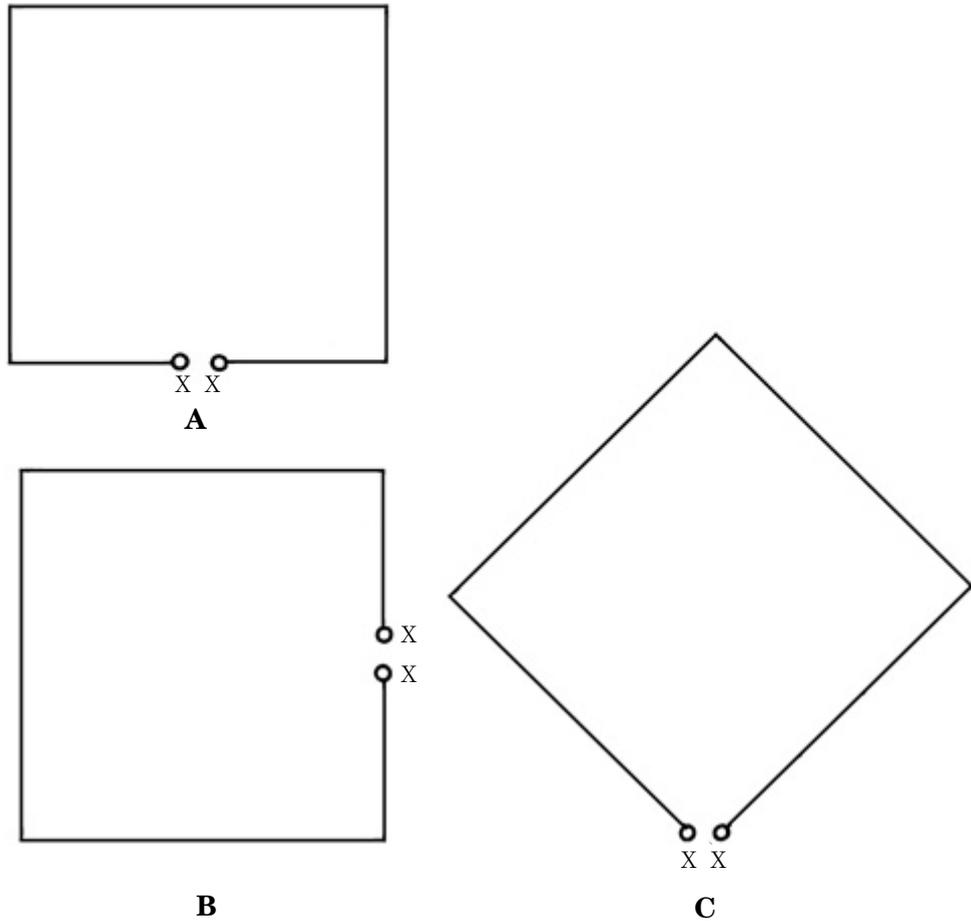
The details for the gusset plate are shown in Fig. 12-7B. The gusset plate is made of a strong insulating material such as fiberglass or  $\frac{3}{4}$ -in marine-grade plywood. It is mounted to a support mast using two or three large U bolts (stainless steel to prevent corrosion). The spreaders are mounted to the gusset plate using somewhat smaller U bolts (again, use stainless steel U bolts to prevent corrosion damage).



12-7A Quad loop antenna.



12-7B Quad loop antenna.

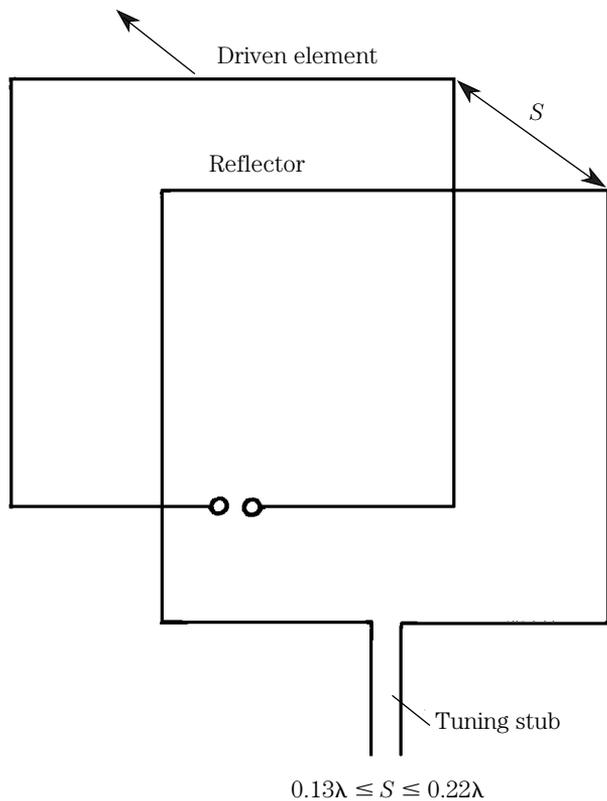


12-8 Feed options for the quad.

There is a running controversy regarding how the antenna compares with other beam antennas, particularly the Yagi. Some experts claim that the cubical quad has a gain of about 1.5 to 2 dB higher than a Yagi (with a comparable boom length between the two elements). In addition, some experts claim that the quad has a lower angle of radiation. Most experts agree that the quad seems to work better at low heights above the earth's surface, but the difference disappears at heights greater than a half-wavelength.

The quad can be used as either a single-element antenna or in the form of a beam. Figure 12-9 shows a pair of elements spaced 0.13 to 0.22 wavelengths apart. One element is the driven element, and it is connected to the coaxial-cable feedline directly. The other element is a reflector, so it is a bit longer than the driven element. A tuning stub is used to adjust the reflector loop to resonance.

Because the wire is arranged into a square loop, one wavelength long, the actual length varies from the naturally resonant length by about 3 percent. The driven



12-9 Quad beam antenna.

element is about 3 percent longer than the natural resonant point. The overall lengths of the wire elements are

$$1. \text{ Driven element:} \quad L = \frac{1005}{F_{\text{MHz}}} \text{ ft} \quad [12.5]$$

$$2. \text{ Reflector:} \quad L = \frac{1030}{F_{\text{MHz}}} \text{ ft} \quad [12.6]$$

$$3. \text{ Director:} \quad L = \frac{975}{F_{\text{MHz}}} \text{ ft} \quad [12.7]$$

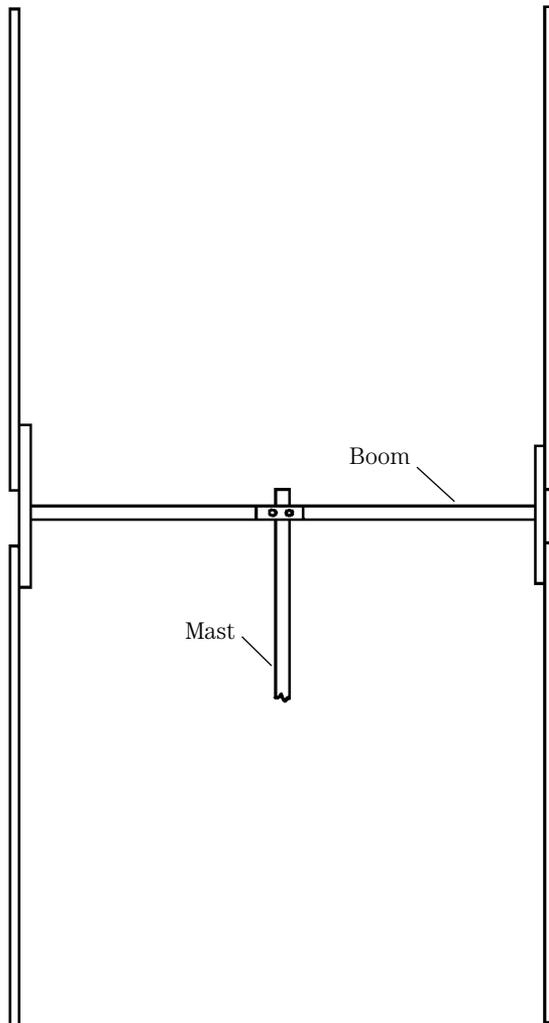
One method for the construction of the quad beam antenna is shown in Fig. 12-10. This particular scheme uses a  $12 \times 12$ -in wooden plate at the center, bamboo (or fiberglass) spreaders, and a wooden (or metal) boom. The construction must be heavy-duty in order to survive wind loads. For this reason, it is probably a better solution to buy a quad kit consisting of the spreaders and the center structural element.

More than one band can be installed on a single set of spreaders. The size of the spreaders is set by the lowest band of operation, so higher frequency bands can be accommodated with shorter loops on the same set of spreaders.

This quad antenna is an example of a multielement, large loop antenna. Additional information on large loops, but not in a beam antenna array, is found in Chap. 14.

## **Inverted bobtail curtain (Thorne array)**

The *bobtail curtain* antenna is a fixed array consisting of three individual quarter-wavelength elements spaced a half-wavelength apart, and fed from the top by a



**12-10** Quad construction.

shorting element or wire. The *inverted bobtail curtain*, or *Thorne array*, consists of an *upside down* bobtail curtain as shown in Fig. 12-11. The radiator elements are each a quarter-wavelength long. Their lengths are found from

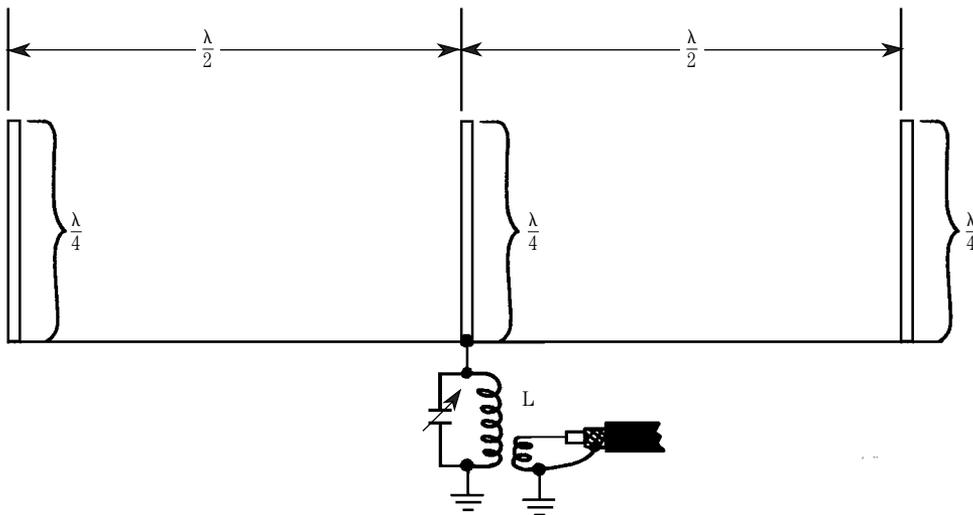
$$L = \frac{246}{F_{\text{MHz}}} \text{ ft} \quad [12.8]$$

The lengths of spacing between the elements are exactly twice above the value, or

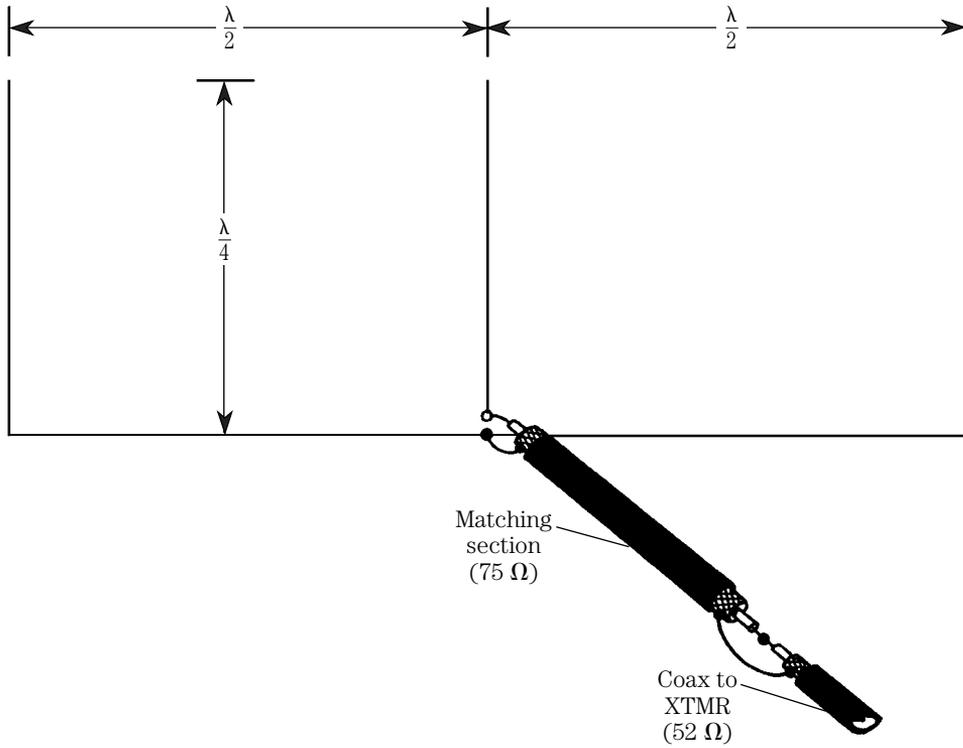
$$L = \frac{492}{F_{\text{MHz}}} \text{ ft} \quad [12.9]$$

The antenna is fed at the base of the center element, through a parallel resonant tuner. The capacitor is a 100- to 200-pF transmitting variable, while the inductor is set to resonate at the band desired (with the capacitor at half to three-quarters full capacitance). A loop or link coupling scheme connects the tuner to the transmission line.

An alternate feed method (Fig. 12-12) worked out by the late J. H. Thorne (K4NFU/5), feeds the end elements from the shield of the coaxial cable, and the center element of the array is fed from the center conductor of the coaxial cable. A coaxial impedance-matching section is used between the cable transmitter and the antenna feedpoint.



12-11 Inverted bobtail curtain.



**12-12** Thorne array.

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## 13 CHAPTER

# Antennas for shortwave reception

ALTHOUGH MOST OF THIS BOOK ADDRESSES ANTENNAS FOR TRANSMITTING, THERE IS A certain body of material that pertains purely to receiving antennas. This material also needs to be addressed, and that is the function of this chapter. There are two readers in mind for this chapter. First, and foremost, is the shortwave listener (SWL). Second, however, is the amateur radio operator who wants to either use a separate receiving antenna on the main station receiver or use an ancillary receiver (common among DXers).

## The law of reciprocity

Antennas possess a property called *reciprocity*. That is a fancy way of saying that an antenna works on reception the same way it does for transmission. Although articles occasionally appear in the literature claiming an HF or VHF design which violates the *law of reciprocity*, to date all have depended on either variable definitions, false premises, or faulty measurements. There is even a school of thought that falsely argues against antenna reciprocity based on an ionospheric anomaly in which propagation depends on the *direction* traveled (see the end of Chap. 2). The bottom line of reciprocity, for the SWL, is that every antenna described in this book can be also used with equal results on receive. For example, a half-wavelength dipole works equally well as a receiver antenna, or as a transmitter antenna.

## Which properties are important?

Selecting a receiver antenna is a function of several factors. Assuming that you want more than a simple longwire (which we will deal with shortly), you will want to home in on the properties desired for your particular monitoring application. Is the antenna to be fixed or rotatable? Do you want omnidirectional or directional reception? In which plane? What about gain?

What about gain? What is gain, for that matter? The concept of antenna gain derives from the fact that directional antennas focus energy. Two kinds of gain figure are often quoted: gain referenced to a dipole antenna on the same frequency, and gain relative to a theoretical construct called an *isotropic radiator*. In the isotropic case, the reference radiator to which the antenna is compared is a spherical point source that radiates equally well in all directions. The dipole exhibits a figure-8 radiation and reception pattern.

The gain of an antenna is merely the ratio (usually expressed in decibels) of the power radiated in a given direction by two antennas (i.e., the reference antenna and the test antenna). If an antenna gain is listed as “8-dB gain over isotropic,” this means that, in the direction specified, the power radiated is 8 dB higher than the same total power applied to an isotropic radiator.

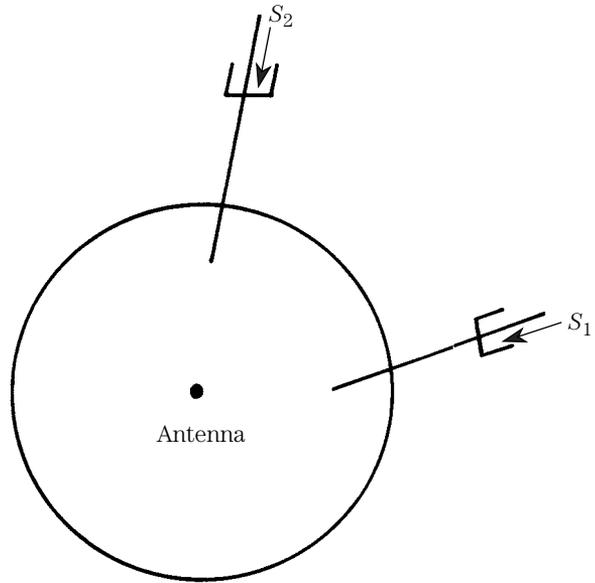
So, of what use is antenna gain? Two answers immediately present themselves. First, by accumulating more signal, the antenna essentially makes your receiver more sensitive. Note that the gain of the antenna does not create a higher powered signal, it merely increases the apparent signal power by focusing energy from a given direction. And, note well that *gain* implies *directivity*. Any antenna that claims to provide gain, but that is truly omnidirectional in all planes, is a fraud. The fundamental assumption is that *gain implies directivity*.

The concept of directivity (hence also of gain) is often taken to mean *horizontal directivity*, which is the case of a dipole antenna. But all forms of antenna radiate in three-dimensional space. Azimuth angle of radiation, and elevation angle of radiation, are both important. Certain 2-m vertical antennas are listed as “gain antennas,” yet the pattern in the horizontal direction is 360°—implying omnidirectional behavior. In the vertical plane, however, lost energy is compressed into a smaller range of elevation angles, so gain occurs by refocusing energy that would have been radiated at a higher than useful angle.

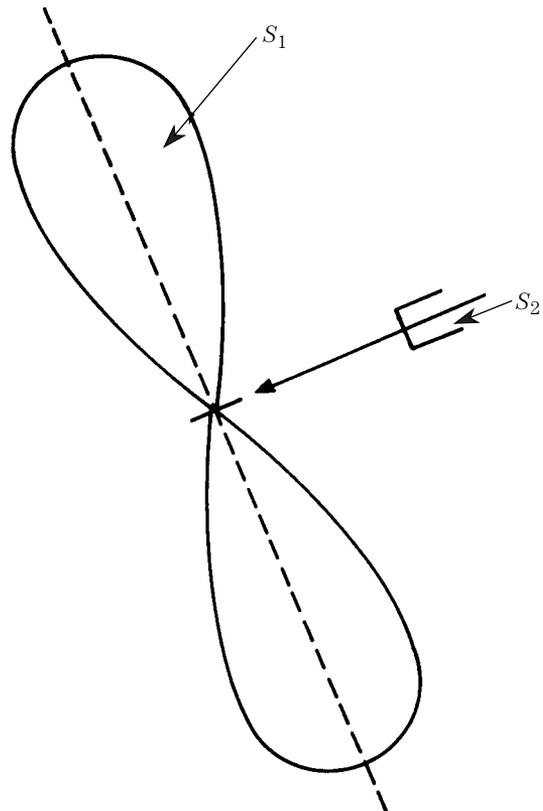
The second application of directivity is in suppressing interfering signals. On the regular AM and FM broadcast bands, channelization permits receiver selectivity to overcome adjacent channel interference in most cases. But in the HF amateur radio and international broadcast bands, channelization is either nonexistent, poorly defined, or ignored altogether. In these cases, interfering “adjacent” channel signals can wipe out a weaker, desired station. Similarly, with cochannel interference (i.e., when both stations are on the same frequency), two or more signals compete in a “dog fight” that neither will ever totally win. Consider Fig. 13-1A. Assume that two 9540-kHz signals,  $S_1$  and  $S_2$ , arrive at the same omnidirectional vertical antenna. Either both signals will be heard, or the stronger signal will drown out the weaker signal.

Now, consider Fig. 13-1B. Here, a dipole is used as the receiving antenna, so a little directivity is obtained. The main lobes of the dipole are wide enough to provide decent reception of signal  $S_1$ , even though the antenna is positioned such that  $S_1$  is not along the maximum line (dotted). But the positioning shown places the interfering cochannel signal ( $S_2$ ) in the null off the ends of the dipole, so it weakens it considerably. The result will be enhanced reception of  $S_1$ . In Fig. 13-1B, the idea is not to exploit the ability of the gain antenna to increase the level of  $S_1$ . Indeed, by placing the antenna as shown, we are not getting  $S_1$  levels as high as might otherwise be

**13-1A** Omnidirectional antenna picks up cochannel interfering signals equally well.



**13-1B** Directional antenna can discriminate against unwanted signal.



possible. The idea here is to place the unwanted signal ( $S_2$ ) into the “notch” in order to make it considerably weaker. Note that the notch is sharper than the peak of the main lobe. If the dipole is placed on a mast, with an antenna rotator, this ability is increased even more.

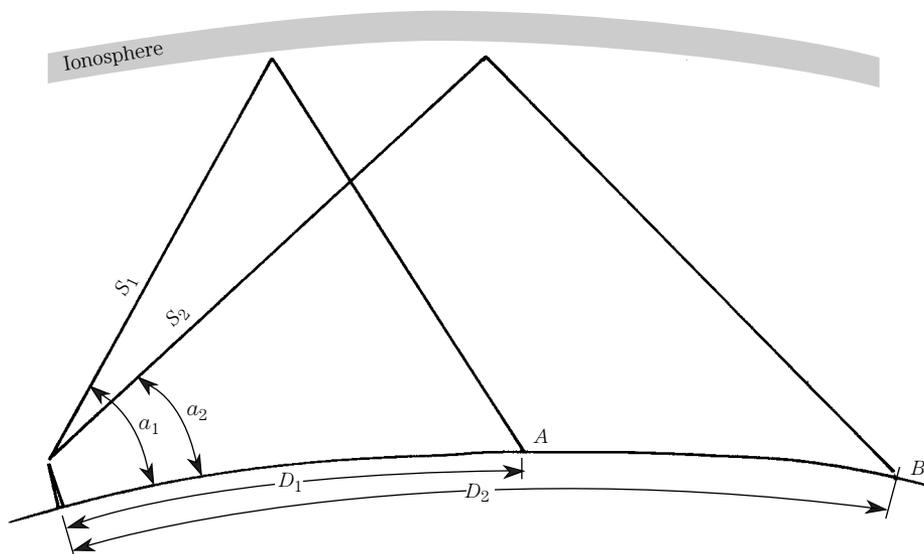
Another antenna parameter, of considerable interest, is *angle of radiation*  $a$ , which (by reciprocity) also means angle of reception. Because HF propagation over long distances is created by skip phenomena, the angle at which the signal hits the ionosphere becomes extremely important. Figure 13-2 shows two situations from the same station. Signal  $S_1$  has a high angle of radiation ( $a_1$ ), so its skip distance ( $D_1$ ) is relatively short. On signal  $S_2$ , however, the angle of radiation ( $a_2$ ) is low, so the skip distance ( $D_2$ ) is much longer than  $D_1$ .

So which situation do you want in your antenna? The impulsive answer would be the long distance angle of radiation ( $a_2$ ), but that is often wrong. The correct answer is: “It depends!” The desired angle of radiation is a function of whether you want to receive a station from point A or point B.

The angle of radiation of the antenna is fixed by its design, that is, by antenna physics. The desired angle is a function of the ionospheric properties at the time of interest, and the operating frequency. For this reason, some well-equipped radio hobbyists have several antennas, of differing properties, to enhance their listening.

## Connection to the receiver

It’s simply too naive to state, I suppose, but let’s do it anyway: An antenna must be properly connected to the receiver before it can be effective. If your antenna uses coaxial cable, and the receiver accepts coax, then no discussion is needed: Attach



13-2 Skip phenomena dependence on angle of radiation.

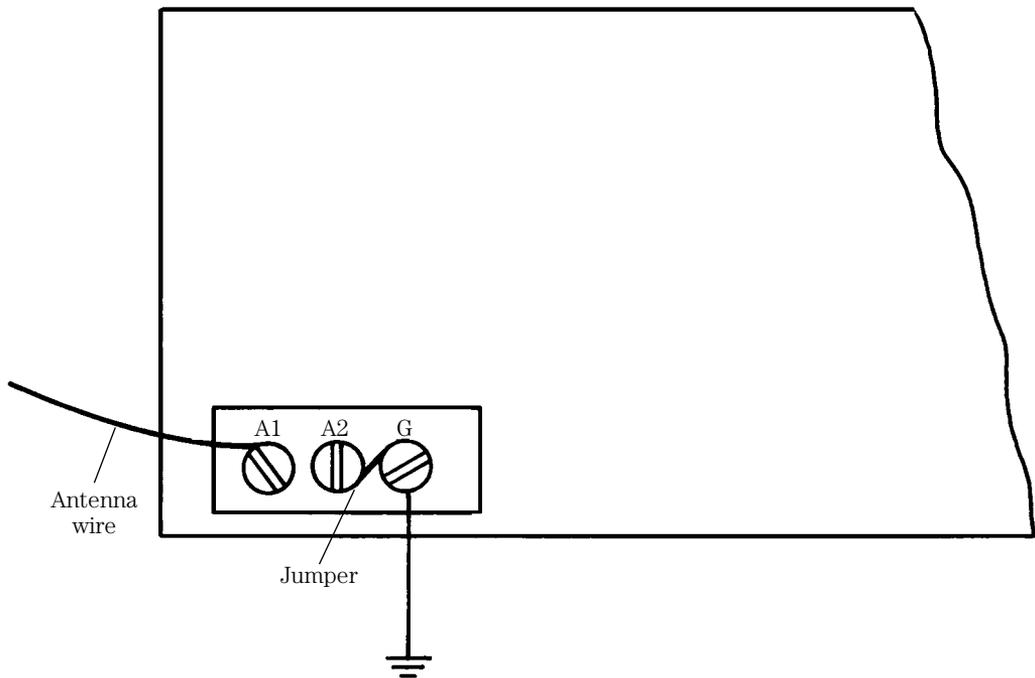
the proper coax connector and plug in. But in other cases, non-coaxial-cable antennas are used.

There are two major forms of antenna input connector used on shortwave receivers. One form uses two (or three) screws intended for either wrapped wire leads or spade lugs, while the other is one or more varieties of coaxial connector. This section covers how each type is connected to a single-wire antenna lead-in.

Consider first the screw-type connector (Fig. 13-3A). Depending upon the design, there will be either two or three screws. If only two screws are found, then one is for the antenna wire and the other is for the ground wire. These screws will be marked something like "A/G" or "ANT/GND," or with the schematic symbols for antenna and ground.

Three-screw designs are intended to accommodate balanced transmission lines such as twin lead, or parallel ladder line. Shortwave listeners can sometimes use ordinary ac line cord (called *zipcord*) as an antenna transmission line. Zipcord has an impedance that approximates the  $75\text{-}\Omega$  impedance of a dipole. When parallel lines of any type are used, connect one lead to A1 and the other to A2. Of course, the ground terminal (G) is connected to the earth ground.

For single-lead antenna lines connect a jumper wire or bar (i.e., a short piece of bare no. 22 solid hookup wire) between A2 and G. This jumper converts the balanced input line to unbalanced. The A2/G terminal is connected to earth ground, while A1 is connected to the single-lead antenna wire.



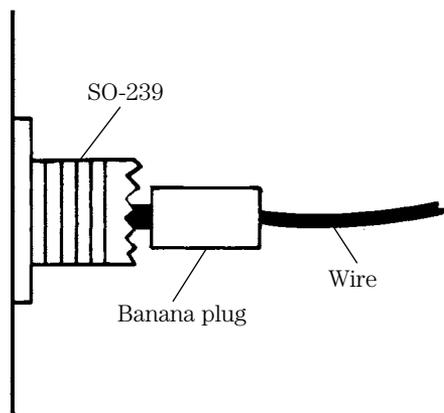
13-3A Connection of wire antenna to balanced antenna terminals on receiver.

On receivers that use an SO-239 coaxial connector, we can use either of two techniques to connect a single-lead wire. First, we can obtain the mating PL-259 plug, and solder the antenna lead to the center conductor pin. The PL-259 connector is then screwed into the mating SO-239 chassis connector. Regardless of the type of coaxial connector, however, the mate can be used for the antenna lead wire. But for SO-239 connectors another alternative is also available. Figure 13-3B shows a “banana plug” attached to the lead wire and inserted into the receptacle of the SO-239.

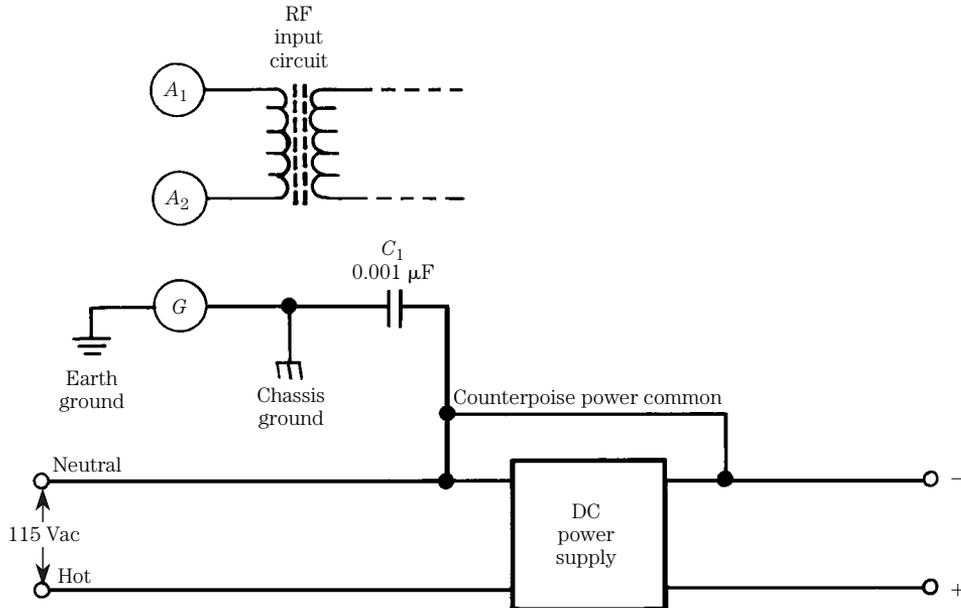
## DANGER!

Certain low-cost receivers, especially older vacuum-tube models, have a so-called ac/dc or transformerless internal dc power supply. On most receivers, the dc common is the chassis, which also serves as the RF signal common. But on ac/dc models the neutral wire of the ac power line serves as the dc common, and it is kept floating as a “counterpoise” ground above the chassis ground used by the RF signals. A capacitor ( $C_1$  in Fig. 13-4) sets the chassis and counterpoise ground at 0-V RF potential, while keeping the counterpoise isolated for dc and 60 Hz ac. A danger exists if either the ac plug is installed backwards or someone plugs the socket in the wall incorrectly (often happens!). Even if  $C_1$  is intact, *a nasty shock can be felt by touching the antenna ground (G or GND) terminal*. The capacitive reactance of  $C_1$  is about  $2.7\text{ M}\Omega$  for 60-Hz ac, so at least a “bite” is going to happen. But if that capacitor is shorted, which is likely on older receivers, then the bite is considerably worse, and might even prove *fatal*. The problem, in that case, is that reversed ac line polarity will set the hot line from the ac socket on the ground lead. The least to expect is massive fireworks and a possible fire hazard; the most to expect is a fire, and your possible electrocution.

The usual advice given to owners of such radios is to make sure that  $C_1$  is intact before using the radio. I prefer a better solution: buy, install, and use a 120:120 Vac isolation transformer to isolate your receiver from the ac power lines. Such a transformer is standard practice in repair shops, and it should also be standard practice in your house.



**13-3B** Connection of wire antenna to coaxial input receiver.



13-4 Dangerous form of antenna/receiver circuitry.

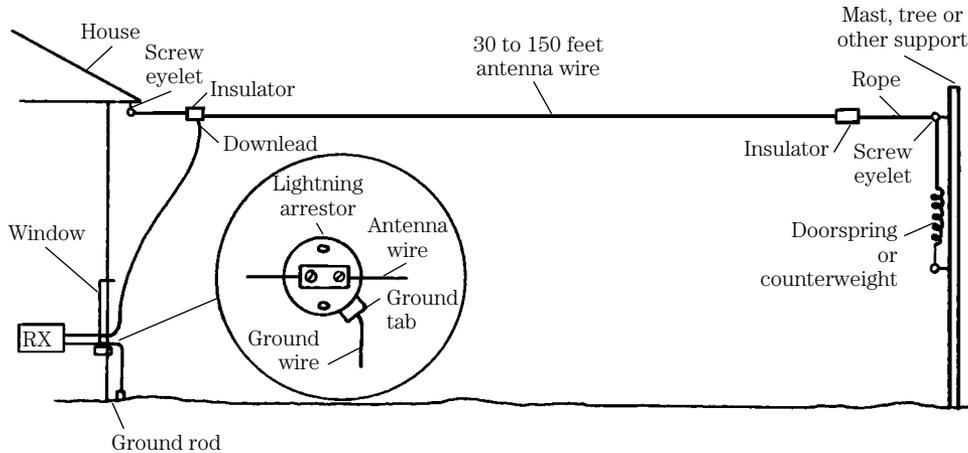
## Wire antennas

This section reviews simple wire antennas that are suitable for the reception of shortwave signals, although not necessarily for transmitting. Once again, you are reminded that the law of reciprocity permits you to use any transmitting antenna found in other chapters for receiving also.

Figure 13-5 shows the common receiving longwire. The antenna element should be 30 to 150 ft in length. Although most texts show it horizontal to the ground (and indeed, a case can be made that performance is better that way), it is not strictly necessary. If you must slope the wire, then it is doubtful that you will notice any reception problems.

The far end of the wire is attached to a supporting structure through an insulator and a rope. The support structure can be another building, a tree, or a mast installed especially for this purpose. Chapter 28 deals with antenna construction practices.

Wind will cause motion in the antenna wire and its supporting structure. Over time, the wind movement will fatigue the antenna wire and cause it to break. Also, if a big enough gust of wind (or a sustained storm) comes along, then even a new antenna can either sag badly or break altogether. You can do either of two things to reduce the problem. First, as shown in Fig. 13-5, a door spring can be used to provide a little variable slack in the wire. The spring tension is selected to be only partially expanded under normal conditions, so wind will increase the tension, and stretch the spring. Make sure that the spring is not too strong to be stretched by the action of wind on the antenna, or no good is accomplished.



13-5 Longwire SWL antenna.

Another tactic is to replace the spring with a counterweight that is heavy enough to keep the antenna nearly taut under normal conditions, but not so heavy that it fails to move under wind conditions. In other words, the antenna tension should exactly balance the counterweight under normal conditions, and not be too great that it stretches the antenna wire excessively.

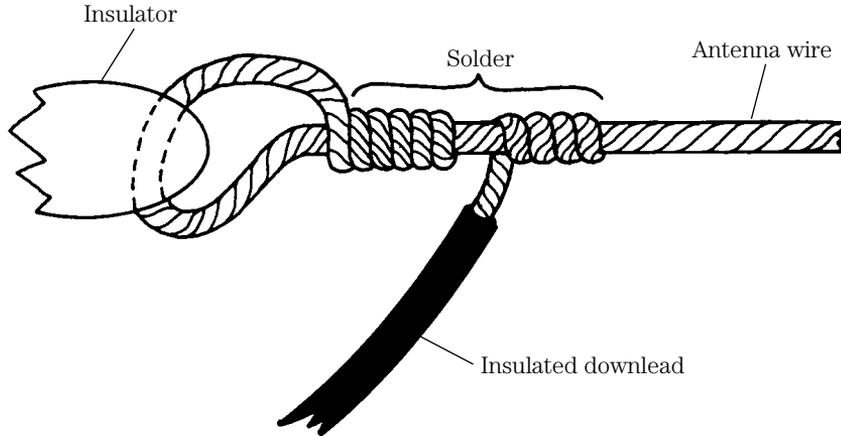
The antenna wire should be either no. 12 or no. 14 hard-drawn copper, or *Copperweld* stranded wire. The latter is actually steel-core wire, but has a copper coating on the outside. Because of skin effect, RF signals flow only in the outer copper coating. Soft-drawn copper wire will stretch and break prematurely, so it should be avoided.

The downlead of the antenna must be insulated, and it should also be stranded wire (which breaks less easily than solid wire). Again, no. 12 or no. 14 wire should be used, although no. 16 would be permissible. The point where the downlead and antenna are joined should be soldered to prevent corrosion of the joint. Mechanical strength is provided by proper splicing technique (see Fig. 13-6). Do not depend on the solder for mechanical strength, for it has none.

There are several ways to bring a downlead into the building. First, if you can tolerate a slight crack in the junction of the sash and sill, then run the wire underneath the sash and close the window. Alternatively, you can buy a flat strap connector to pass under the window. This method is electrically the same as running the lead, but is mechanically nicer. Chapter 28 deals with several methods, and should be consulted.

## Grounding

The ground lead should be a heavy conductor, such as heavy wire or braid. The shield stripped from RG-8 or RG-11 coaxial cable is suitable for most applications. For reception purposes only, the ground may be a cold-water pipe inside the house. Do not use the hot-water pipes (which are not well grounded) or gas pipes (which



13-6 Construction details.

are dangerous to use). Also, be aware that residential air conditioner liquid lines look like copper cold-water pipes in some cases. Don't use them. Chapter 29 deals with grounding.

The lightning arrester is a safety precaution, and it *must* be used! Its purpose is to supply an alternative path to ground in the event of a lightning strike. Although at least one text calls the arrester optional, it is not. Besides the obvious safety reasons (which are reason enough), there are also legal and economic reasons for using the arrester. Your local government building and/or fire codes might require a lightning arrester for outdoor antennas. Also, your insurance company might not honor your homeowner's policy if the lightning arrester (required by local code) is not used. The antenna lightning arrester is *not optional*, so use it.

**WARNING! DO NOT EVER ATTEMPT TO INSTALL AN ANTENNA BY CROSSING A POWER LINE! EVER! NO MATTER WHAT YOU BELIEVE OR WHAT YOUR FRIENDS TELL YOU, IT'S NEVER SAFE AND IT COULD KILL YOU.**

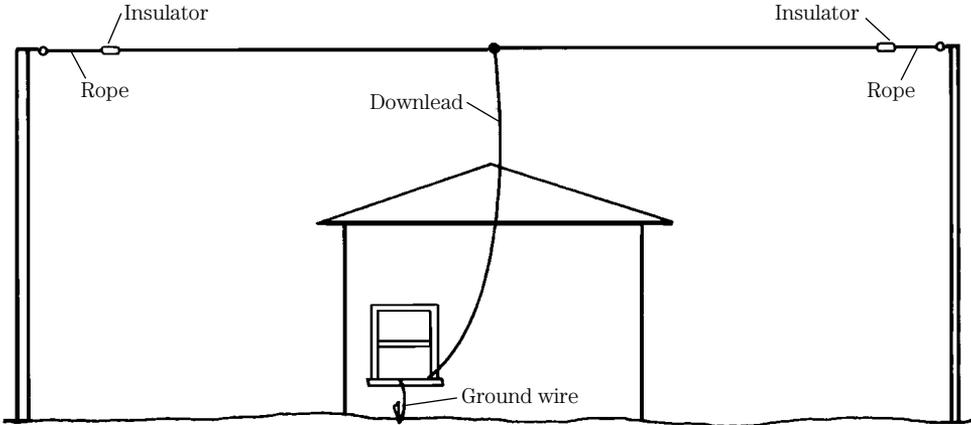
The rest of this discussion touches on antennas, other than the receiver long-wire. Because construction details are similar, we will not repeat them. You are, by the way, encouraged to also read Chap. 28 to glean more details on antenna construction methods.

### The flattop antenna

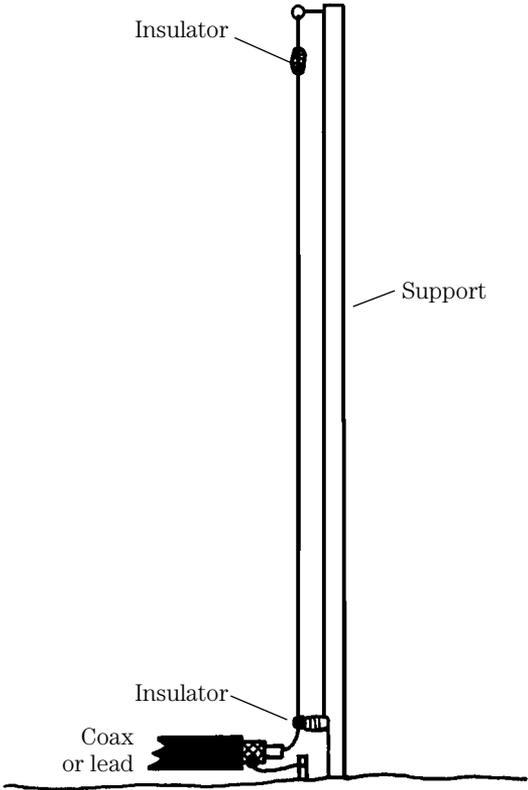
The "flattop" antenna is shown in Fig. 13-7. This antenna is a relative of the longwire, with the exception that the downline is at the approximate center of the antenna section. The flattop should be at least a half-wavelength ( $492/F_{\text{MHz}}$ ) at the lowest frequency of operation. The advantage of the flattop is that it allows the maximum use of space in the configuration shown.

## Vertical antennas

It is also possible to build shortwave listener antennas in vertical polarization. Figure 13-8 shows one version of this type of antenna. The support (which could



13-7 Flattop SWL antenna.



13-8 Wire vertical SWL antenna—works for hams too.

be a tree or building) should have enough height to be at least a quarter-wavelength on the lowest frequency of operation. The antenna is fed at the base with coaxial cable. The center conductor of the coax is connected to the antenna element, and the shield is connected to the ground rod at the base of the structure. You are encouraged to see Chap. 10, which deals with limited space and hidden antennas, for a different version of this antenna. It is possible to install the wire (or multiple wires of different lengths) inside of a length of PVC plumbing pipe. The pipe serves as the support structure, and the conductors are placed inside.

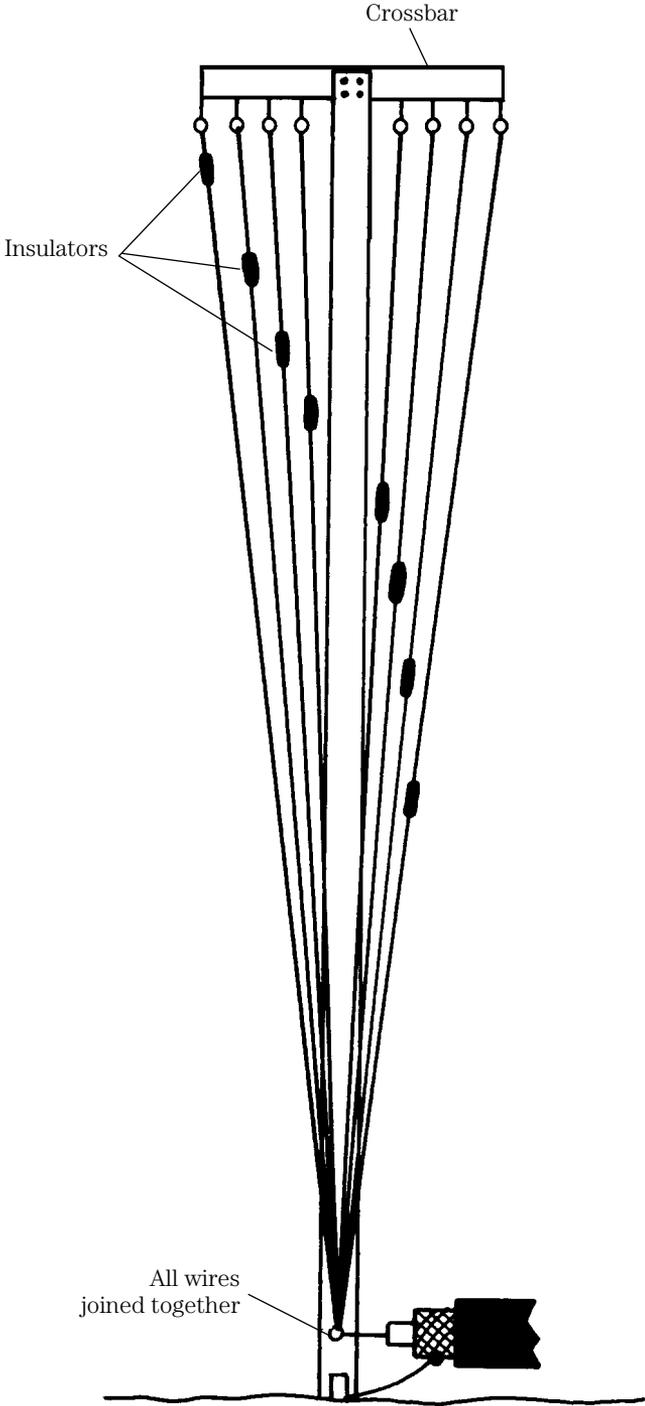
Different lengths of conductor are required for different bands of operation. You can calculate the length (in feet) required for quarter-wavelength vertical antennas from  $L_{ft} = 246/F_{MHz}$ . Figure 13-9A shows how several bands are accommodated from the same feedline on the same support structure. In this particular case eight different antenna elements are supported from the same tee bar. Be sure to insulate them from each other, and from the support structure. Again, PVC piping can be used for the support. Another method for accommodating several different bands is to tie the upper ends of the wires to a sloping rope (as in Fig. 13-9B).

## Wire directional antenna

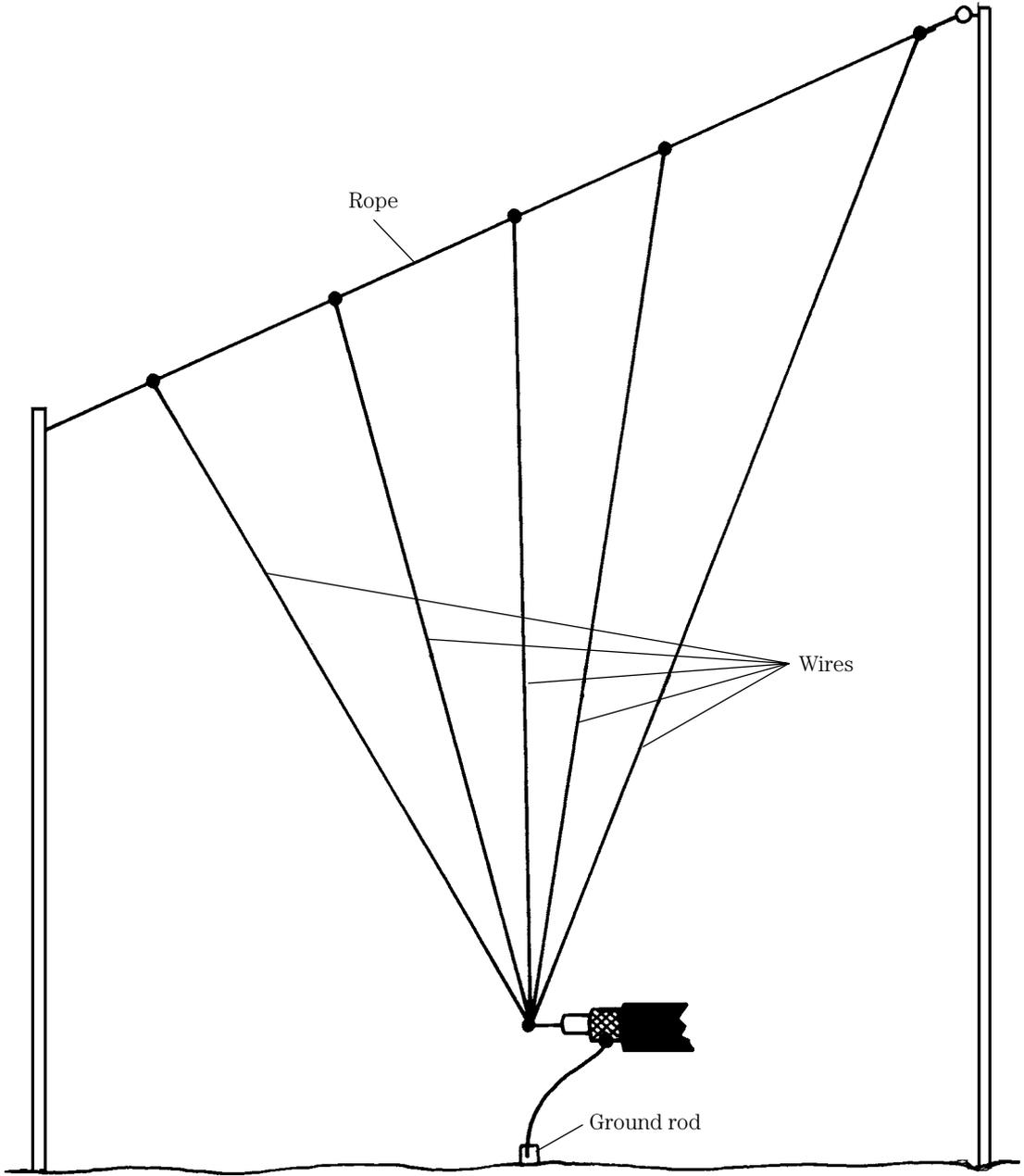
A directional antenna has the ability to enhance reception of desired signals, while rejecting undesired signals arriving from slightly different directions. Although directivity normally means a beam antenna, or at least a rotatable dipole, there are certain types of antenna that allow fixed antennas to be both directive and variable. See Chap. 7 for fixed but variable directional antennas and Chap. 11 for fixed and non-variable directive arrays. Those antennas are transmitting antennas, but they work equally well for reception. This section shows a crude, but often effective, directional antenna that allows one to select the direction of reception with pin plugs or switches.

Consider Fig. 13-10. In this case, a number of quarter-wavelength radiators are fanned out from a common feedpoint at various angles from the building. At the near end of each element is a female banana jack. A pair of balanced feedlines from the receiver (300- $\Omega$  twin lead, or similar) are brought to the area where the antenna elements terminate. Each wire in the twin lead has a banana plug attached. By selecting which banana jack is plugged into which banana plug, you can select the directional pattern of the antenna. If the receiver is equipped with a balanced antenna input, then simply connect the other end of the twin lead direction to the receiver. Otherwise, use one of the couplers shown in Fig. 13-11.

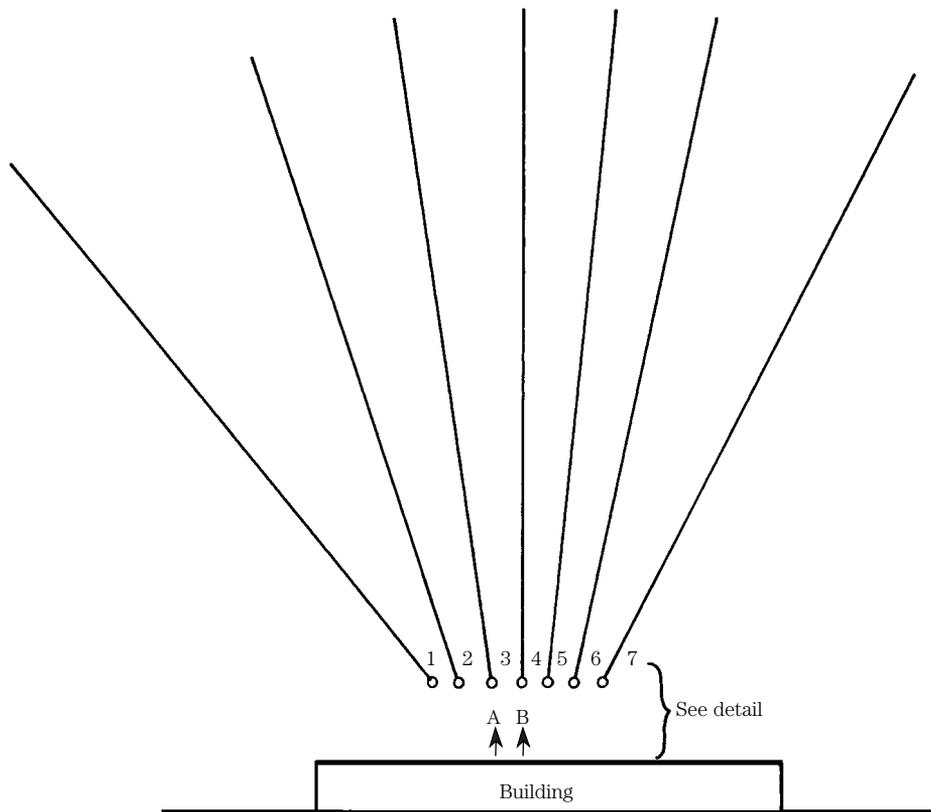
Figure 13-11A shows a balanced antenna coupler that is tuned to the frequency of reception. The coil is tuned to resonance by the interaction of the inductor and the capacitor. Antenna impedance is matched by selecting the taps on the inductor to which the feedline is attached. A simple RF broadband coupler is shown in Fig. 13-11B. This transformer is wound over a ferrite core, and consists of 12 to 24 turns of no. 26 enameled wire, with more turns being used for lower frequencies, and fewer for higher frequencies. Experiment with the number of turns in order to determine the correct value. Alternatively, use a 1:1 balun transformer instead of Fig. 13-11B; the type intended for amateur radio antennas is overkill powerwise, but it will work nicely.



**13-9A** Multiband wire vertical SWL antenna: tee bar style.

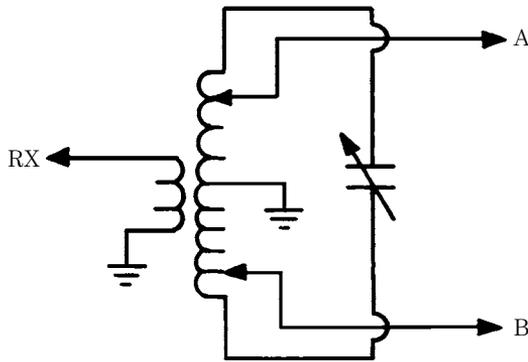


13-9B Multiband wire vertical SWL antenna: fan style.

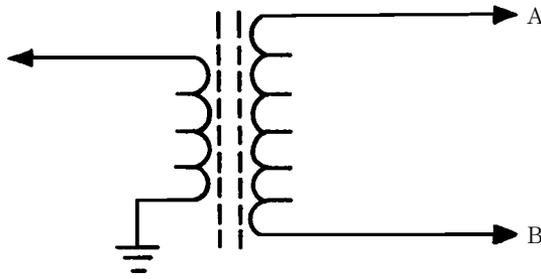


**13-10** Directional SWL antenna (view from above).

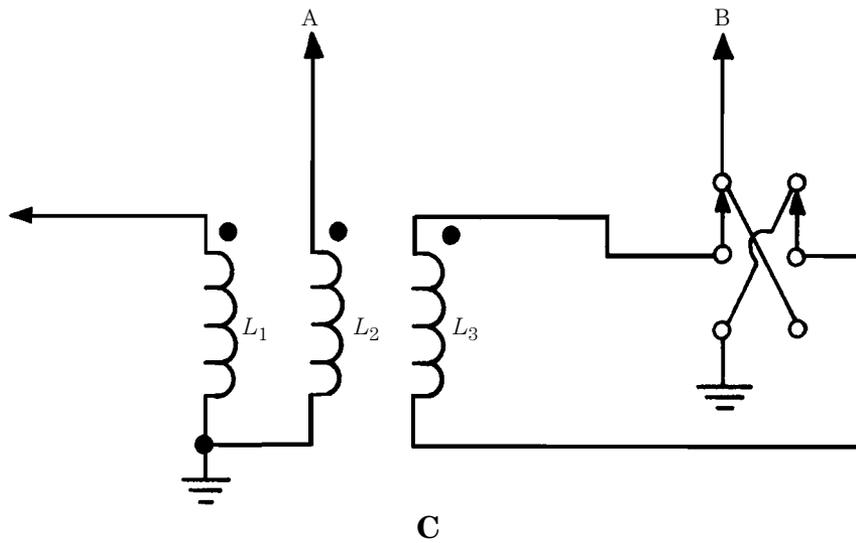
The antenna of Fig. 13-10 works by phasing the elements so as to null, or enhance (as needed), certain directions. This operation becomes a little more flexible if you build a phasing transformer, as shown in Fig. 13-11C and 13-11D. Windings  $L_1$ ,  $L_2$ , and  $L_3$  are wound “trifilar” style onto a ferrite core. Use 14 turns of no. 26 enameled wire for each winding. The idea in this circuit is to feed one element from coil  $L_2$  in the same way all of the time. This port becomes the  $0^\circ$  phase reference. The other port, B, is fed from a reversible winding, so it can either be in phase or  $180^\circ$  out of phase with port A. Adjust the DPDT switch and the banana plugs of Fig. 13-10 for the best reception.



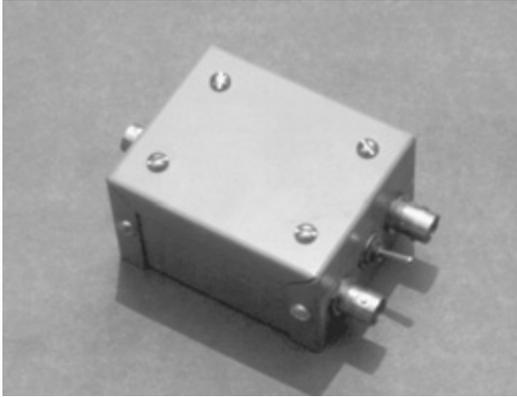
13-11A Antenna matching tuner.



13-11B Antenna matching transformer.



13-11C Phase-switching antenna matcher.



**13-11D** Phasing box permits phasing antennas  $0^\circ$  to  $180^\circ$ .

# 14

CHAPTER

## Large wire loop antennas

THERE ARE TWO BASIC FORMS OF LOOP ANTENNAS: SMALL AND LARGE. THESE TWO TYPES have different characteristics, work according to different principles, and have different purposes. Small loops are those in which the current flowing in the wire has the same phase and amplitude at every point in the loop (which fact implies a very short wire length, i.e., less than  $0.2\lambda$ ). Such loops respond to the magnetic field component of the electromagnetic radio wave. A large loop antenna has a wire length greater than  $0.2\lambda$ , with most being either  $\lambda/2$ ,  $1\lambda$ , or  $2\lambda$ . The current in a large loop varies along the length of the wire in a manner similar to other wire antennas.

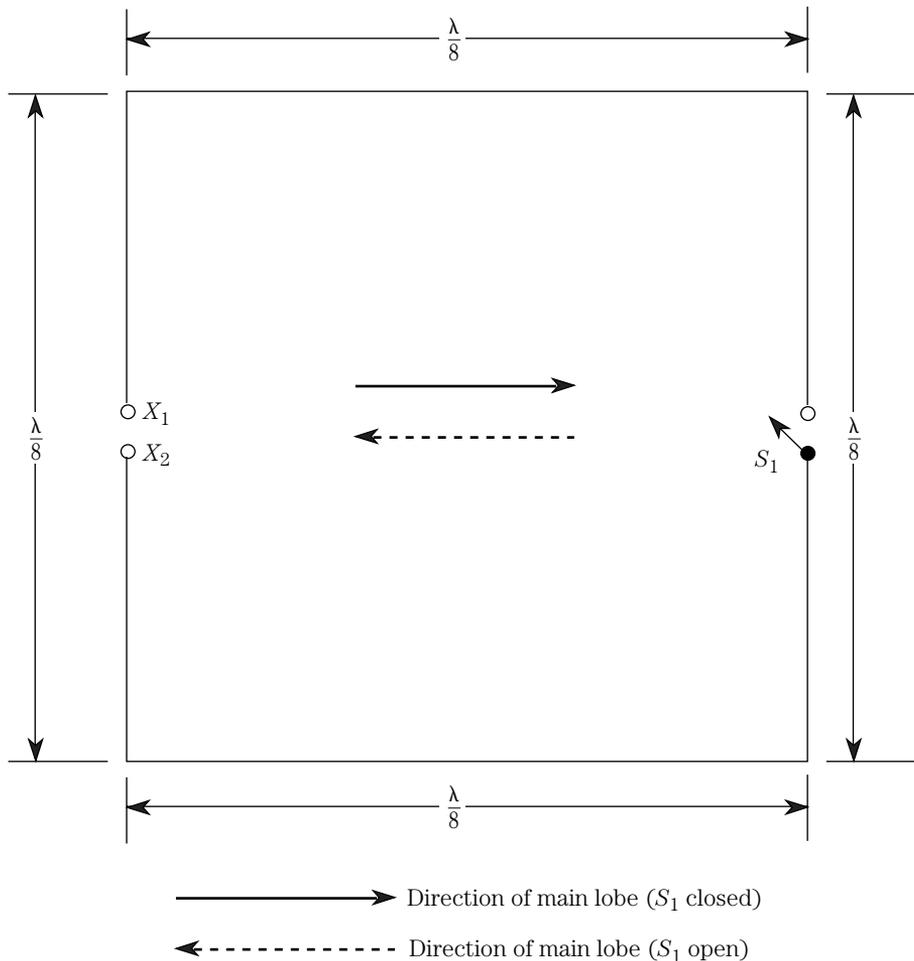
### $\lambda/2$ large loops

The performance of large wire loop antennas depends in part on their size. Figure 14-1 shows a half-wavelength loop (i.e., one in which the four sides are each  $\lambda/8$  long). There are two basic configurations for this antenna: *continuous* ( $S_1$  closed) and *open* ( $S_1$  open). In both cases, the feedpoint is at the midpoint of the side opposite the switch.

The direction of the main reception, or radiation lobe (i.e., the direction of maximum reception), depends on whether  $S_1$  is open or closed. With  $S_1$  closed, the main lobe is to the right (solid arrow); and with  $S_1$  open, it is to the left (broken arrow). Direction reversal can be achieved by using a switch (or relay) at  $S_1$ , although some people opt for unidirectional operation by eliminating  $S_1$ , and leaving the loop either open or closed.

The feedpoint impedance is considerably different in the two configurations. In the closed-loop situation (i.e.,  $S_1$  closed), the antenna can be modeled as if it were a half-wavelength dipole bent into a square and fed at the ends. The feedpoint ( $X_1-X_2$ ) impedance is on the order of  $3\text{ k}\Omega$  because it occurs at a voltage antinode (current node). The current antinode (i.e.,  $I_{\max}$ ) is at  $S_1$ , on the side opposite the feedpoint. An antenna tuning unit (ATU), or RF impedance transformer, must be used to match the lower impedance of the transmission lines needed to connect to receivers.

The feedpoint impedance of the open-loop configuration ( $S_1$  open) is low because the current antinode occurs at  $X_1-X_2$ . Some texts list the impedance as “about  $50\ \Omega$ ,” but my own measurements on several test loops were somewhat higher (about  $70\ \Omega$ ). In either case, the open loop is a reasonable match for either 52- or 75- $\Omega$  coaxial cable.



**14-1** Half-wave square loop antenna with reversal switch.

Neither  $\lambda/2$  loop configuration shows gain over a dipole. The figure usually quoted is  $-1$ -dB forward gain (i.e., a loss compared with a dipole), and about 6-dB front-to-back ratio (FBR). Such low values of FBR indicate that there is no deep notch (“null”) in the pattern.

A lossy antenna with a low FBR seems like a born loser, and in most cases it is. But the  $\lambda/2$  loop finds a niche where size must be constrained, for one reason or another. In those cases, the  $\lambda/2$  loop can be an alternative. These antennas can be considered limited-space designs, and can be mounted in an attic, or other limited-access place, as appropriate.

A simple trick will change the gain, as well as the direction of radiation, of the closed version of the  $\lambda/2$  loop. In Fig. 14-2, a pair of inductors ( $L_1$  and  $L_2$ ) are inserted into the circuit at the midpoints of the sides adjacent to the side containing the feedpoints. These inductors should have an inductive reactance  $X_L$  of about  $360 \Omega$  in the center of the band of operation. The inductance of the coil is

$$L_{\mu\text{H}} = \frac{3.6 \times 10^8}{2\pi F_{\text{Hz}}} \quad [14.1]$$

where

$L_{\mu\text{H}}$  is the coil inductance, in microhenrys ( $\mu\text{H}$ )

$F_{\text{Hz}}$  is the midband frequency, in hertz (Hz)

**Example** Find the inductance for the coils in a loaded half-wavelength closed-loop antenna that must operate in a band centered on 10.125 MHz.

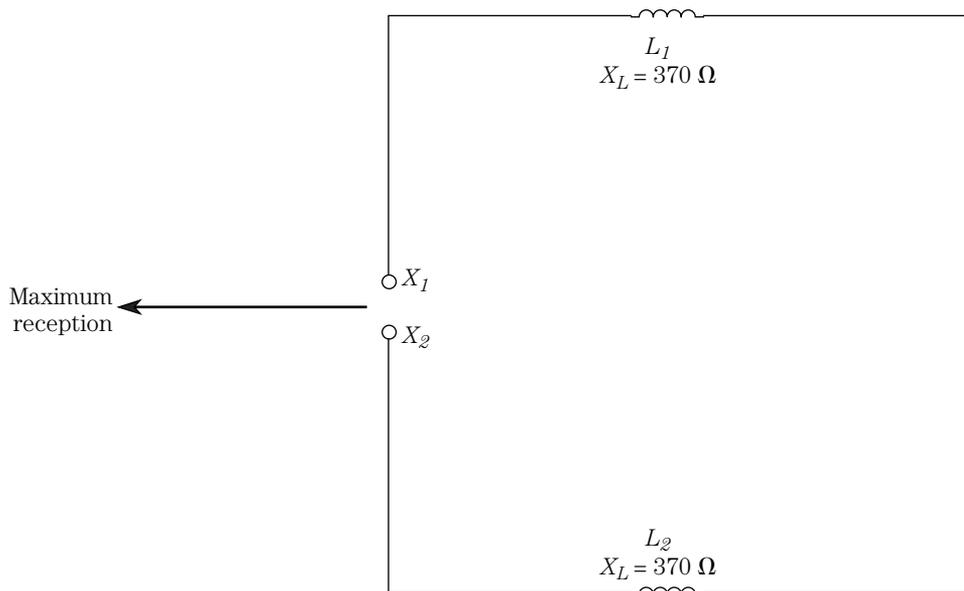
**Solution:**

[Note: 10.125 MHz = 10,125,000 Hz]

$$L_{\mu\text{H}} = \frac{3.6 \times 10^8}{2\pi (10,125,000 \text{ Hz})} = 5.7 \mu\text{H}$$

The coils force the current antinodes toward the feedpoint, reversing the direction of the main lobe, and creating a gain of about +1 dB over a half-wavelength dipole.

The currents flowing in the antenna can be quite high, so when making the coils, be sure to use a size that is sufficient for the power and current levels anticipated. The 2- to 3-in B&W Air-Dux style coils are sufficient for most amateur radio use. Smaller coils are available on the market, but their use is limited to low-power situations.



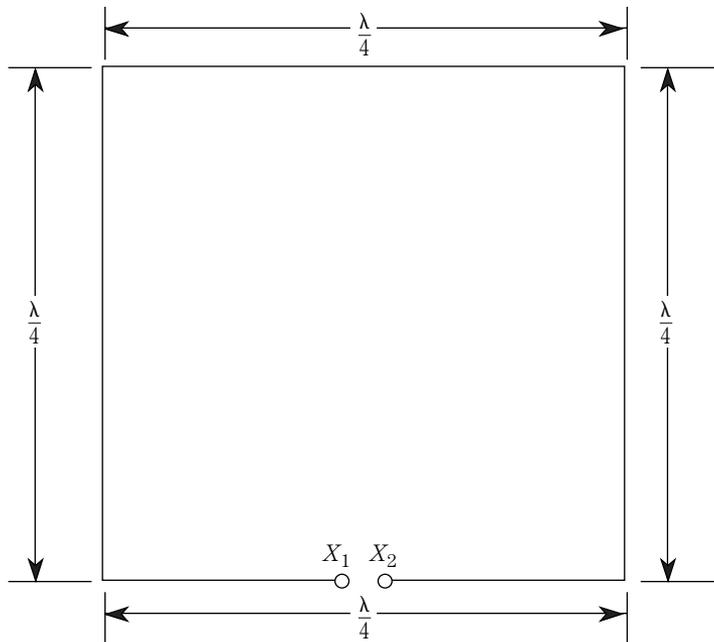
**14-2** Use of inductive loading to reduce the size of antenna, and make the pattern more uniform.

## 1λ large loops

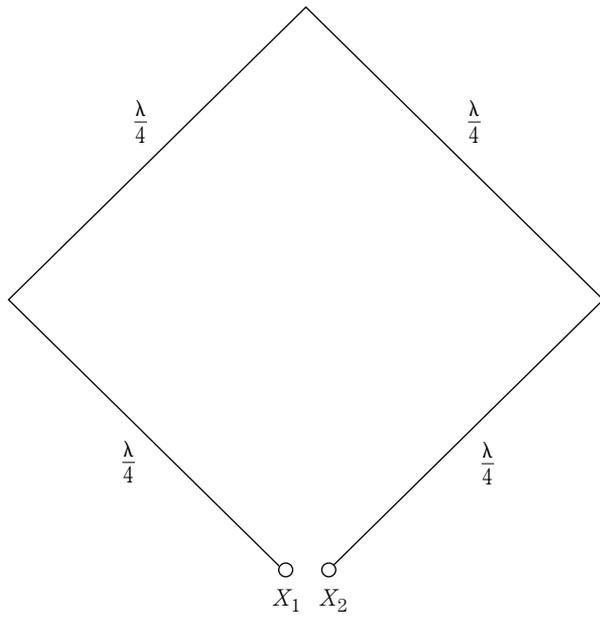
If size is not forcing you to a  $\lambda/2$  loop, then a  $1\lambda$  loop might be just the ticket. It produces a gain of about +2 dB over a dipole in the directions that are perpendicular to the plane of the loop. The azimuth patterns formed by these antennas are similar to the figure-8 pattern of the dipole. Three versions are shown: the square loop (Fig. 14-3), the diamond loop (Fig. 14-4), and the delta loop (a.k.a. D-loop and triangle—Fig. 14-5). The square and diamond loops are built with  $\lambda/4$  on each side, and the delta loop is  $\lambda/3$  on each side. The overall length of wire needed to build these antennas is

$$L_{\text{feet}} = \frac{1005}{F_{\text{MHz}}} \quad [14.2]$$

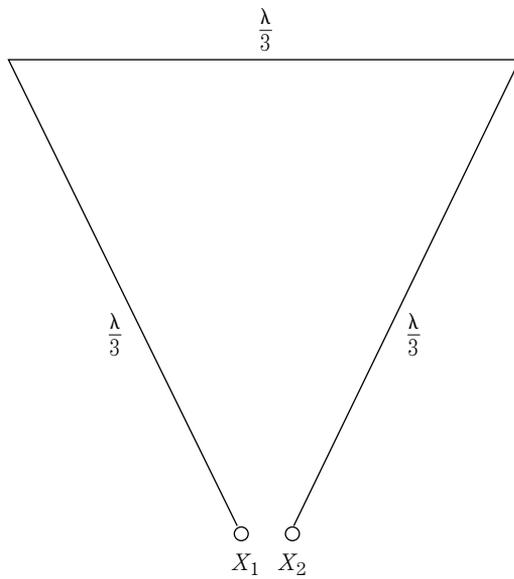
The polarization of the three loop antennas is horizontal, because of the location of the feedpoints. On the square loop, moving the feedpoint to the middle of either vertical side will provide vertical polarization. Similarly, on the diamond loop vertical polarization is realized by moving the feedpoint to either of the two adjacent apexes. On the delta loop, placing the feedpoint at either of the two other apexes produces a diagonal polarization that offers approximately equal vertical and horizontal polarization components.



**14-3** Quarter-wavelength square loop (single-element quad).



**14-4** Bottom-fed loop.



**14-5** Delta loop antenna.

The feedpoint impedance of the  $1\lambda$  loop is around  $100\ \Omega$ , so it provides a slight mismatch to  $75\text{-}\Omega$  coax and a 2:1 mismatch to  $52\text{-}\Omega$  coax. A very good match to  $52\text{-}\Omega$  coax can be produced using the scheme of Fig. 14-6. Here, a quarter-wavelength coaxial cable matching section is made of  $75\text{-}\Omega$  coaxial cable. The length of this cable should be

$$L_{\text{feet}} = \frac{246V}{F_{\text{MHz}}} \quad [14.3]$$

where

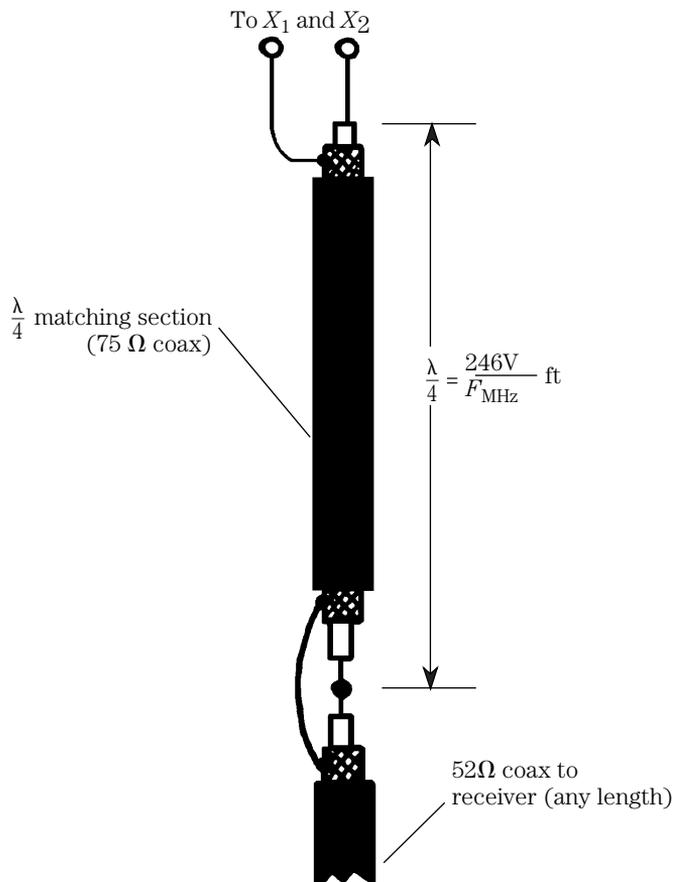
$L_{\text{feet}}$  is the length, in feet (ft)

$V$  is the velocity factor of the coax

$F_{\text{MHz}}$  is the frequency, in megahertz (MHz)

The impedance  $Z_o$  of the cable used for the matching section should be

$$Z_o = \sqrt{Z_L Z_s} \quad [14.4]$$



**14-6** Quarter-wavelength coaxial matching section.

where

$Z_o$  is the characteristic impedance of the coax used in the matching section, in ohms

$Z_L$  is the feedpoint impedance of the antenna, in ohms

$Z_s$  is the source impedance (i.e., the 52- $\Omega$  characteristic impedance of the line to the receiver in standard systems)

When Eq. 14.4 is applied to this system, where  $Z_s = 52 \Omega$ ,  $Z_L = 100 \Omega$ :

$$Z_o = \sqrt{(100 \Omega)(52 \Omega)} = 72 \Omega \quad [14.5]$$

This is a very good match to 75- $\Omega$  coaxial cable.

## Demiquad loop antenna

The *demiquad* is a single-element  $1\lambda$  quad antenna. The length of the antenna is, like the cubical quad beam antenna (see Chap. 12), one wavelength. Figure 14-7 shows a type of demi-quad based on the tee-cross type of mast.

The impedance-matching section is a quarter-wavelength piece of 75- $\Omega$  coaxial cable (RG-58/U or RF-11/U). The length of the matching section is determined from:

$$L = \frac{246V}{F_{\text{MHz}}} \quad [14.6]$$

where

$L$  is the overall length, in feet

$F_{\text{MHz}}$  is the frequency, in megahertz

$V$  is the velocity factor of the coaxial cable (typically 0.66, 0.70, or 0.80)

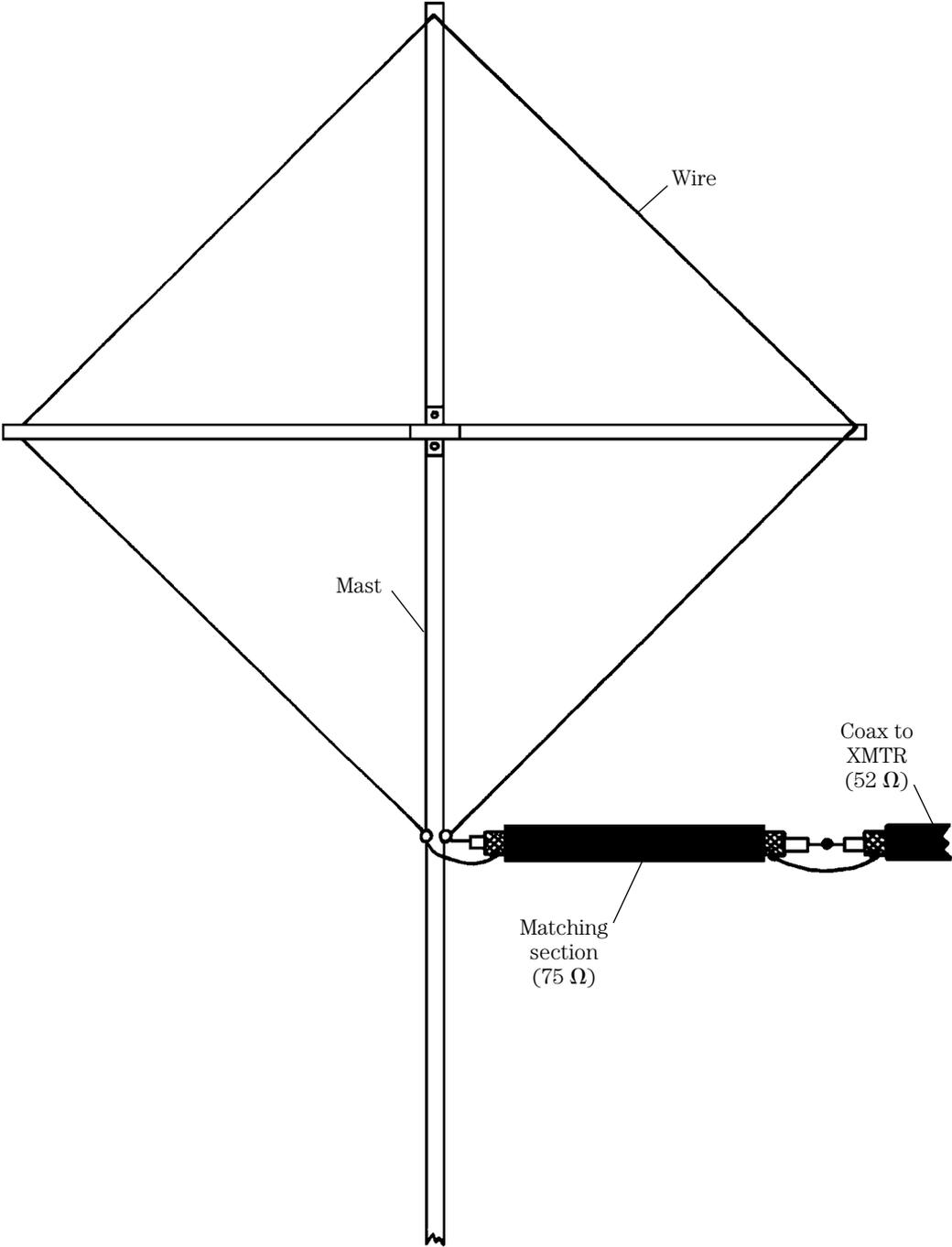
## Delta loop

The *delta loop antenna*, like the Greek uppercase letter “delta” ( $\Delta$ ) from which it draws its name, is triangle-shaped (Fig. 14-8). The delta loop is a full wavelength, with elements approximately 2 percent longer than the natural wavelength (like the quad). The actual length will be a function of the proximity and nature of the underlying ground, so some experimentation is necessary. The approximate preadjustment lengths of the sides are found from:

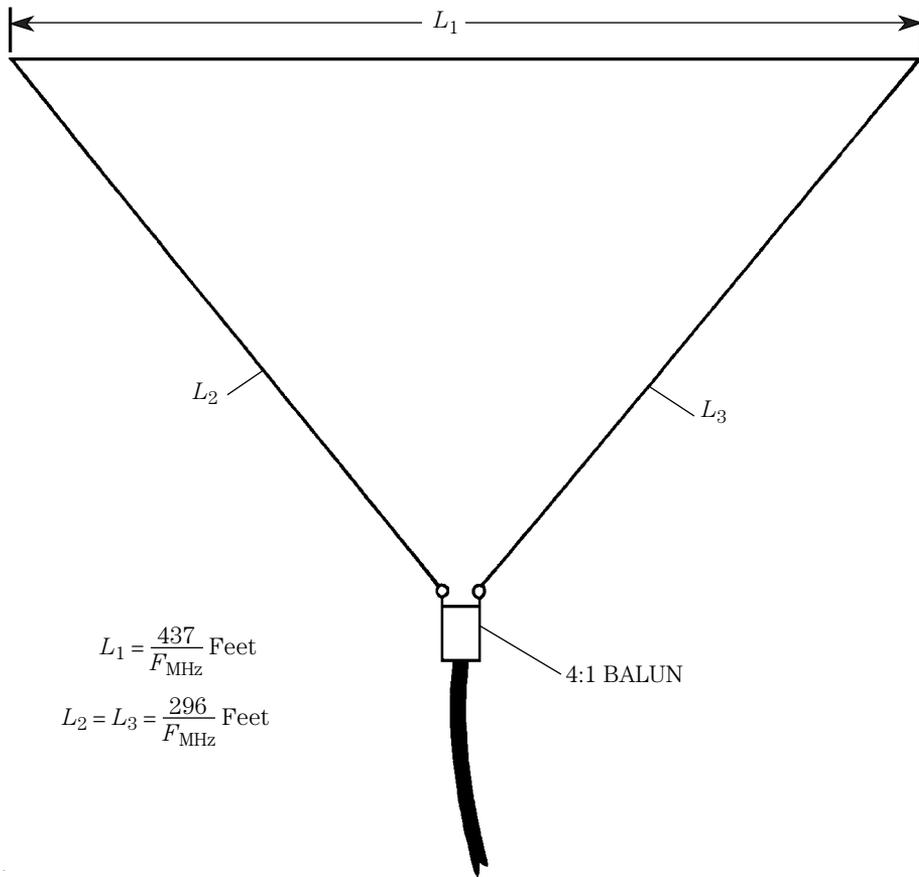
$$L_1 = \frac{437}{F_{\text{MHz}}} \text{ ft} \quad [14.7]$$

$$L_2 = L_3 = \frac{296}{F_{\text{MHz}}} \quad [14.8]$$

The delta loop antenna is fed from 52- $\Omega$  coaxial cable through a 4:1 balun transformer. The delta loop can be built in a fixed location, and will offer a bidirectional pattern.



14-7 Demiquad antenna.



14-8 Delta loop antenna.

## Half-delta sloper (HDS)

The *half-delta sloper* (HDS) antenna (Fig. 14-9) is similar to the full delta loop, except that (like the quarter-wavelength vertical) half of the antenna is in the form of an “image” in the ground. Gains of 1.5 to 2 dB are achievable. The HDS antenna consists of two elements: a  $\lambda/3$ -wavelength sloping wire and a  $\lambda/6$  vertical wire (on an insulated mast), or a  $\lambda/6$  metal mast. Because the ground currents are very important, much like the vertical antenna, either an extensive radial system at both ends is needed, or a base ground return wire (buried) must be provided.

The HDS will work on its design frequency, plus harmonics of the design frequency. For a fundamental frequency of 5 MHz, a vertical segment of 33 ft and a sloping section of 66 ft is needed. The lengths for any frequency are found from

$$\frac{\lambda}{3} = \frac{328}{F_{\text{MHz}}} \quad [14.9]$$

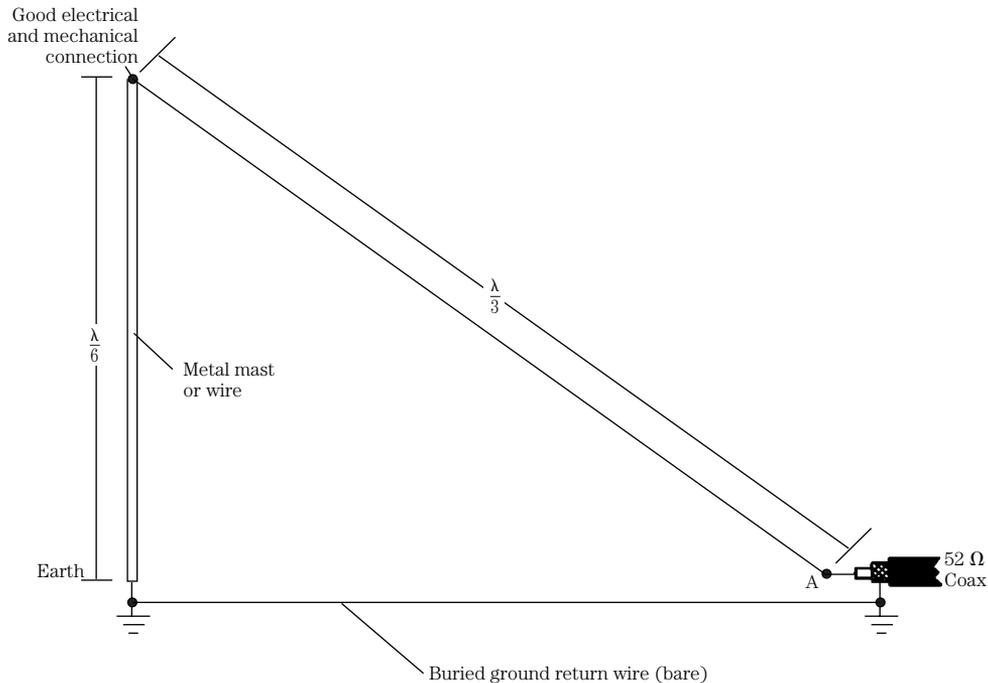
and

$$\frac{\lambda}{6} = \frac{164}{F_{\text{MHz}}} \quad [14.10]$$

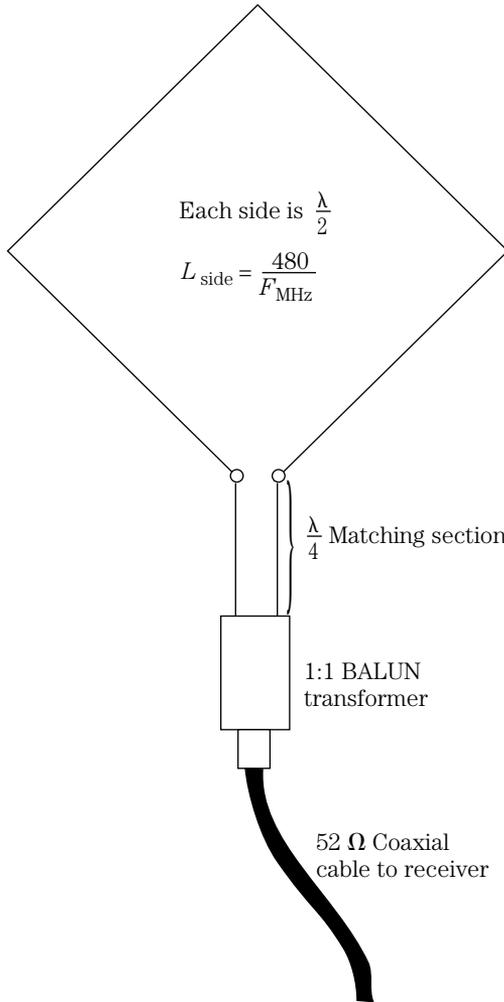
The HDS is fed at one corner, close to the ground. If only the fundamental frequency is desired, then you can feed it with 52- $\Omega$  coaxial cable. But at harmonics, the feedpoint impedance changes to as high as 1000  $\Omega$ . If harmonic operation is intended, then an antenna tuning unit (ATU) is needed at point *A* to match these impedances.

## Bisquare loop antenna

The *bisquare antenna*, shown in Fig. 14-10, is similar to the other large loops, except that it is  $\lambda/2$  on each side, making a total wire length of two wavelengths. This antenna is built like the diamond loop shown earlier (i.e., it is a large square loop fed at an apex that is set at the bottom of the assembly). In this case, the loop is fed either with an antenna tuning unit (to match a 1000- $\Omega$  impedance) or a quarter-wave-length matching section made of 300- $\Omega$  or 450- $\Omega$  twin-lead transmission line. A 1:1 balun transformer connects the 75- $\Omega$  coaxial cable to the matching section.



14-9 Half-delta loop antenna.



**14-10** Bisquare  $2\lambda$  square loop antenna.

The bisquare antenna offers as much as 4-dB gain broadside to the plane of the antenna (i.e., in and out of the book page), in a figure-8 pattern, on the design frequency. It is horizontally polarized. When the frequency drops to one-half of the design frequency, the gain drops to about 2 dB, and the antenna works like the diamond loop covered previously.

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# 15

Chapter

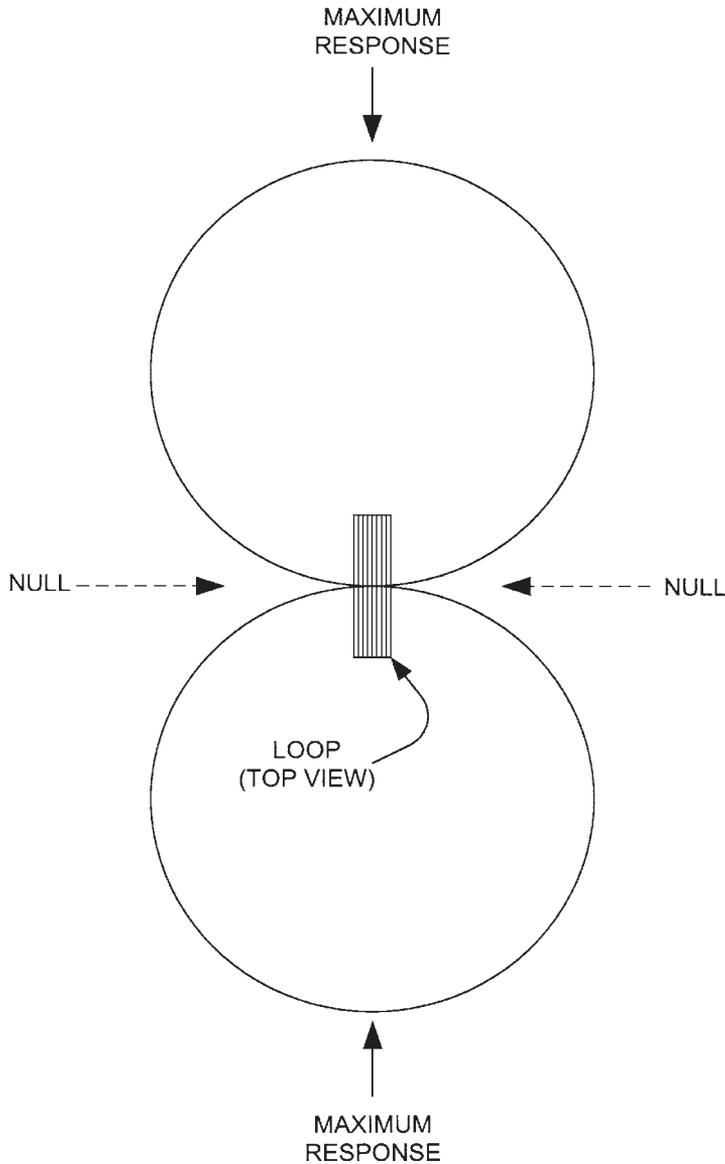
## Small loop receiving antennas

Radio direction finders and people who listen to the AM broadcasting bands, VLF, medium-wave, or the so-called low-frequency tropical bands are all candidates for a small loop antenna. These antennas are fundamentally different from large loops and other sorts of antennas used in these bands. Large loop antennas have a length of at least  $0.5\lambda$ , and most are quite a bit larger than  $0.5\lambda$ . Small loop antennas, on the other hand, have an overall length that is less than  $0.22\lambda$ , with most being less than  $0.10\lambda$ .

The small loop antenna responds to the magnetic field component of the electromagnetic wave instead of the electrical field component. One principal difference between the large loop and the small loop is found when examining the RF currents induced in a loop when a signal intercepts it. In a large loop, the current will vary from one point in the conductor to another, with voltage varying out of phase with the current. In the small loop antenna, the current is the same throughout the entire loop.

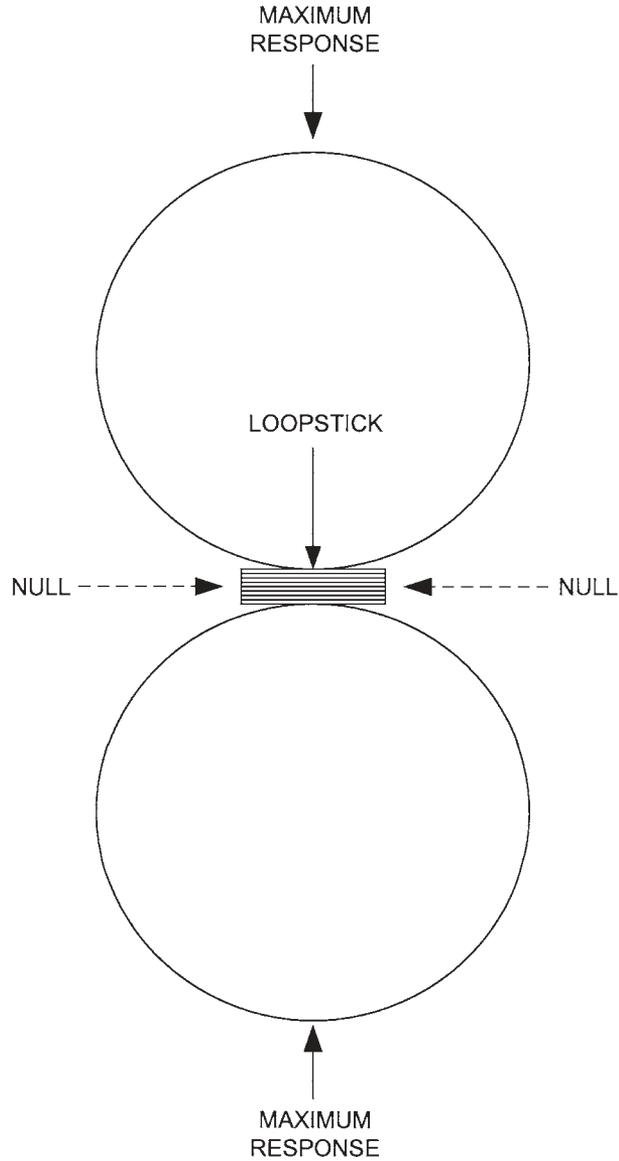
The differences between small loops and large loops show up in some interesting ways, but perhaps the most striking is the directions of maximum response—the main lobes—and the directions of the nulls. Both types of loops produce figure-8 patterns but in directions at right angles with respect to each other. The large loop antenna produces main lobes *orthogonal*, at right angles or “broadside,” to the plane of the loop. Nulls are off the sides of the loop. The small loop, however, is exactly the opposite: The main lobes are off the sides of the loop (in the direction of the loop plane), and the nulls are broadside to the loop plane (Fig. 15-1A). Do not confuse small loop behavior with the behavior of the loopstick antenna. Loopstick antennas are made of coils of wire wound on a ferrite or powdered-iron rod. The direction of maximum response for the loopstick antenna is broadside to the rod, with deep nulls off the ends (Fig. 15-1B). Both loopsticks and small wire loops are used for radio direction-finding and for shortwave, low-frequency medium-wave, AM broadcast band, and VLF listening.

The nulls of a loop antenna are very sharp and very deep. Small changes of pointing direction can make a profound difference in the response of the antenna. If you point a loop antenna so that its null is aimed at a strong station, the signal strength of



**15-1A** Large loop antenna.

the station appears to drop dramatically at the center of the notch. Turn the antenna only a few degrees one way or the other, however, and the signal strength increases sharply. The depth of the null can reach 10 to 15 dB on sloppy loops and 30 to 40 dB on well-built loops (30 dB is a very common value). I have seen claims of 60-dB nulls for some commercially available loop antennas. The construction and uniformity of the loop are primary factors in the sharpness and depth of the null.



**15-1B** Loopstick antenna.

At one time, the principal use of the small loop antenna was radio direction-finding, especially in the lower-frequency bands. The RDF loop is mounted with a compass rose to allow the operator to find the direction of minimum response. The null was used, rather than the peak response point, because it is far narrower than the peak. As a result, precise determination of direction is possible. Because the null is bidirectional, ambiguity exists as to which of the two directions is the correct

direction. What the direction-finder “finds” is a line along which the station exists. If the line is found from two reasonably separated locations and the lines of direction are plotted on a map, then the two lines will cross in the area of the station. Three or more lines of direction (a process called *triangulation*) yields a pretty precise knowledge of the station’s actual location.

Today, these small loops are still used for radio direction-finding, but their use has been extended into the general receiving arena, especially on the low frequencies. One of the characteristics of these bands is the possibility of strong local interference smothering weaker ground-wave and sky-wave stations. As a result, you cannot hear cochannel signals when one of them is very strong and the other is weak. Similarly, if a cochannel station has a signal strength that is an appreciable fraction of the desired signal and is slightly different in frequency, then the two signals will heterodyne together and form a whistling sound in the receiver output. The frequency of the whistle is an audio tone equal to the difference in frequency between the two signals. This is often the case when trying to hear foreign BCB signals on frequencies (called *split frequencies*) between the standard spacing. The directional characteristics of the loop can help if the loop null is placed in the direction of the undesired signal.

Loops are used mainly in the low-frequency bands even though such loops are either physically larger than high-frequency loops or require more turns of wire. Loops have been used as high as VHF and are commonly used in the 10-m ham band for such activities as hidden transmitter hunts. The reason why low frequencies are the general preserve of loops is that these frequencies are more likely to have substantial ground-wave signals. Sky-wave signals lose some of their apparent directivity because of multiple reflections. Similarly, VHF and UHF waves are likely to reflect from buildings and hillsides and so will arrive at angles other than the direction of the transmitter. As a result, the loop is less useful for the purpose of radio direction-finding. If your goal is not RDF but listening to the station, this is hardly a problem. A small loop can be used in the upper shortwave bands to null a strong local ground-wave station in order to hear a weaker sky-wave station. Finally, loops can be useful in rejecting noise from local sources, such as a “leaky” electric power line or a neighbor’s outdoor light dimmer.

Let’s examine the basic theory of small loop antennas and then take a look at some practical construction methods.

## Grover’s equation

Grover’s equation (Grover, 1946) seems closer to the actual inductance measured in empirical tests than certain other equations that are in use. This equation is

$$L_{\mu\text{H}} = (K_1 N^2 a) \times \left\{ \ln \left[ \frac{K_2 a N}{(N + 1) b} \right] + K_3 + \left[ \frac{K_4 (N + 1) b}{aN} \right] \right\}$$

where  $L_{\mu\text{H}}$  is the inductance, in microhenrys ( $\mu\text{H}$ )  
 $a$  is the length of a loop side, in centimeters (cm)  
 $b$  is the loop width, in centimeters (cm)

$n$  is the number of turns in the loop  
 $K_1$  through  $K_4$  are shape constants and are given in Table 15-1  
 $\ln$  is the natural log of this portion of the equation

## Air core frame loops (“box” loops)

A wire loop antenna is made by winding a large coil of wire, consisting of one or more turns, on some sort of frame. The shape of the loop can be circular, square, triangular, hexagonal, or octagonal. For practical reasons, the square loop seems to be most popular. With one exception, the loops considered in this section will be square, so you can easily duplicate them.

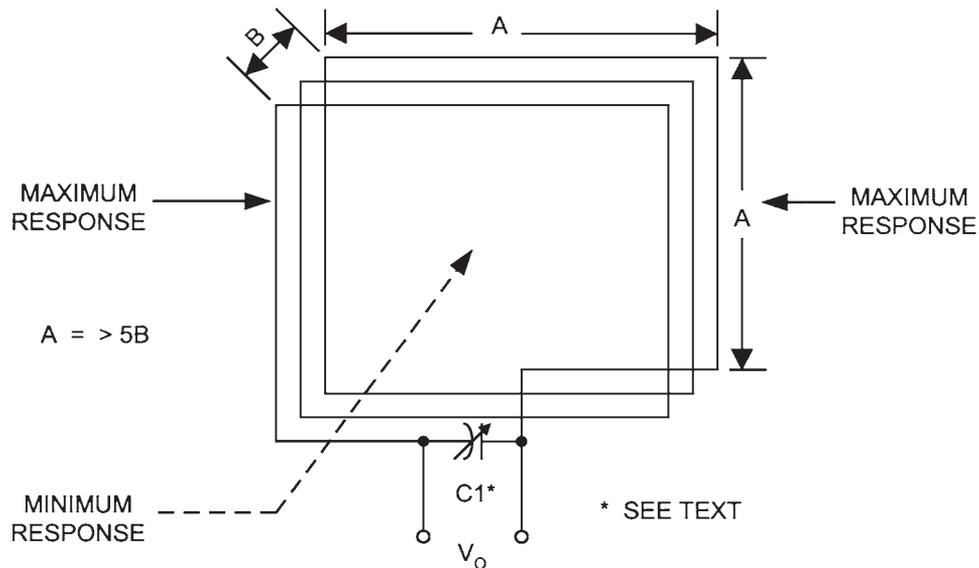
The basic form of the simplest loop is shown in Fig. 15-2. This loop is square, with sides the same length  $A$  all around. The width of the loop ( $B$ ) is the distance from the first turn to the last turn in the loop, or the diameter of the wire if only one turn is used. The turns of the loop in Fig. 15-2 are *depth wound*, meaning that each turn of the loop is spaced in a slightly different parallel plane. The turns are spaced evenly across distance  $B$ . Alternatively, the loop can be wound such that the turns are in the same plane (this is called *planar winding*). In either case, the sides of the loop ( $A$ ) should be not less than five times the width ( $B$ ). There seems to be little difference between depth- and planar-wound loops. The far-field patterns of the different shape loops are nearly the same if the respective cross-sectional areas ( $\pi r^2$  for circular loops and  $A^2$  for square loops) are less than  $\lambda^2/100$ .

The reason why a small loop has a null when its broadest aspect is facing the signal is simple, even though it seems counterintuitive at first blush. Take a look at Fig. 15-3. Here, we have two identical small loop antennas at right angles to each other. Antenna  $A$  is in line with the advancing radio wave, whereas antenna  $B$  is broadside to the wave. Each line in the wave represents a line where the signal strength is the same, i.e., an “isopotential line.” When the loop is in line with the signal (antenna  $A$ ), there is a difference of potential from one end of the loop to the other, so current can be induced in the wires. When the loop is turned broadside, however, all points on the loop are on the same potential line, so there is no difference of potential between segments of the conductor. Thus little signal is picked up (and the antenna therefore sees a null).

The actual voltage across the output terminals of an untuned loop is a function of the angle of arrival of the signal  $\alpha$  (Fig. 15-4), as well as the strength of the signal and the design of the loop. The voltage  $V_o$  is given by

**Table 15-1. Shape constants**

Shape	$K_1$	$K_2$	$K_3$	$K_4$
Triangle	0.006	1.155	0.655	0.135
Square	0.008	1.414	0.379	0.33
Hexagon	0.012	2.00	0.655	0.135
Octagon	0.016	2.613	0.7514	0.0715



**15-2A** A simple loop antenna.

$$V_o = \frac{2\pi ANE_f \cos(\alpha)}{\lambda}$$

where  $V_o$  is the output voltage of the loop

$A$  is the area of the loop, in square meters ( $\text{m}^2$ )

$N$  is the number of turns of wire in the loop

$E_f$  is the strength of the signal, in volts per meter (V/m)

$\alpha$  is the angle of arrival of the signal

$\lambda$  is the wavelength of the arriving signal

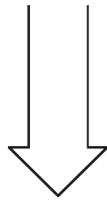
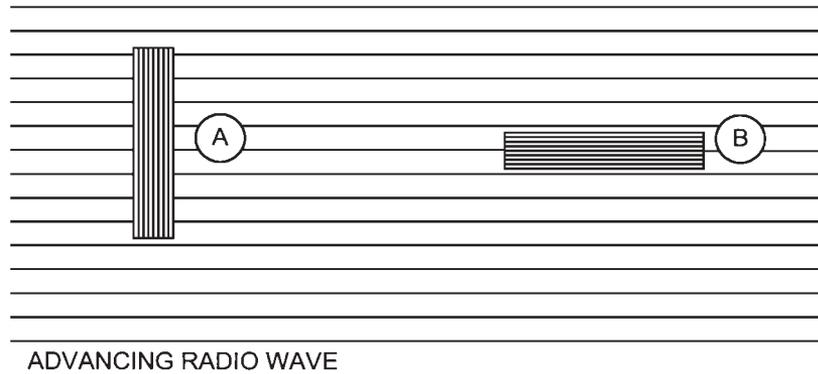
Loops are sometimes specified in terms of the *effective height* of the antenna. This number is a theoretical construct that compares the output voltage of a small loop with a vertical piece of the same kind of wire that has a height of

$$H_{\text{eff}} = \frac{2\pi NA}{\lambda}$$

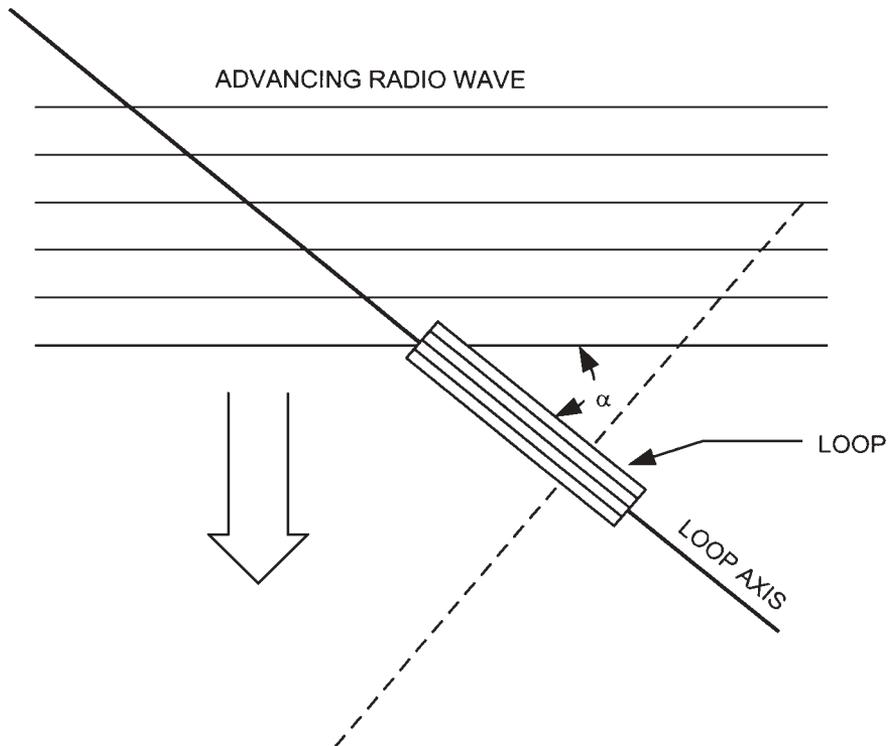
If a capacitor (such as  $C_1$  in Fig. 15-2) is used to tune the loop, then the output voltage  $V_o$  will rise substantially. The output voltage found using the first equation is multiplied by the loaded  $Q$  of the tuned circuit, which can be from 50 to 100:

$$V_o = \frac{2\pi ANE_f Q \cos(\alpha)}{\lambda}$$

Even though the output signal voltage of tuned loops is higher than that of untuned loops, it is nonetheless low compared with other forms of antenna. As a result, a loop preamplifier usually is needed for best performance.



15-3 Two small loop antennas at right angles to each other.

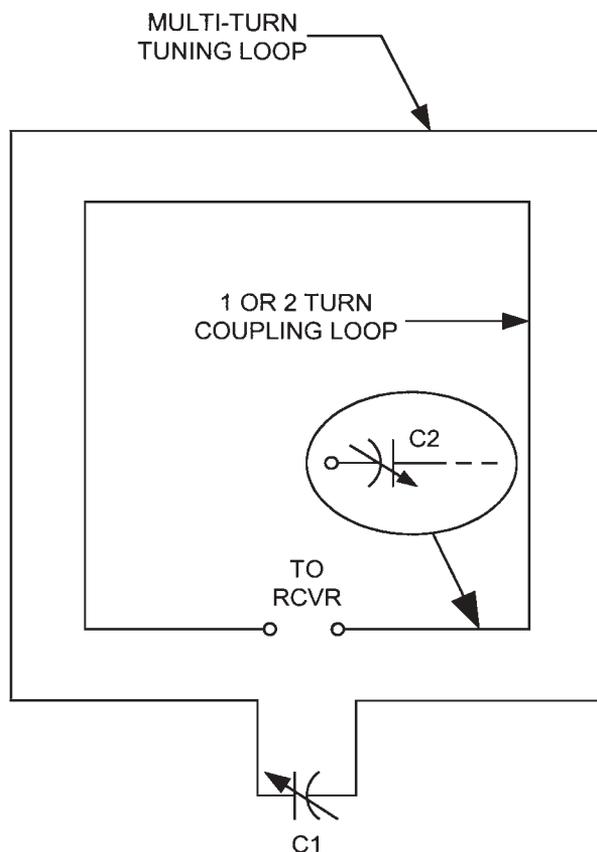


15-4 An untuned loop antenna.

## Transformer loops

It is common practice to make a small loop antenna with two loops rather than just one. Figure 15-5 shows such a transformer loop antenna. The main loop is built exactly as discussed above: several turns of wire on a large frame, with a tuning capacitor to resonate it to the frequency of choice. The other loop is a one- or two-turn *coupling loop*. This loop is installed in very close proximity to the main loop, usually (but not necessarily) on the inside edge not more than a couple of centimeters away. The purpose of this loop is to couple signal induced from the main loop to the receiver at a more reasonable impedance match.

The coupling loop is usually untuned, but in some designs a tuning capacitor ( $C_2$ ) is placed in series with the coupling loop. Because there are many fewer turns on the coupling loop than on the main loop, its inductance is considerably smaller. As a result, the capacitance to resonate is usually much larger. In several loop antennas constructed for purposes of researching this chapter, I found that a 15-turn main loop resonated in the AM BCB with a standard 365-pF capacitor, but the two-turn



15-5 A transformer loop antenna.

coupling loop required three sections of a ganged  $3 \times 365$ -pF capacitor connected in parallel to resonate at the same frequencies.

In several experiments, I used computer ribbon cable to make the loop turns. This type of cable consists of anywhere from 8 to 64 parallel insulated conductors arranged in a flat ribbon shape. Properly interconnected, the conductors of the ribbon cable form a continuous loop. It is no problem to take the outermost one or two conductors on one side of the wire array and use them for a coupling loop.

## Tuning schemes for loop antennas

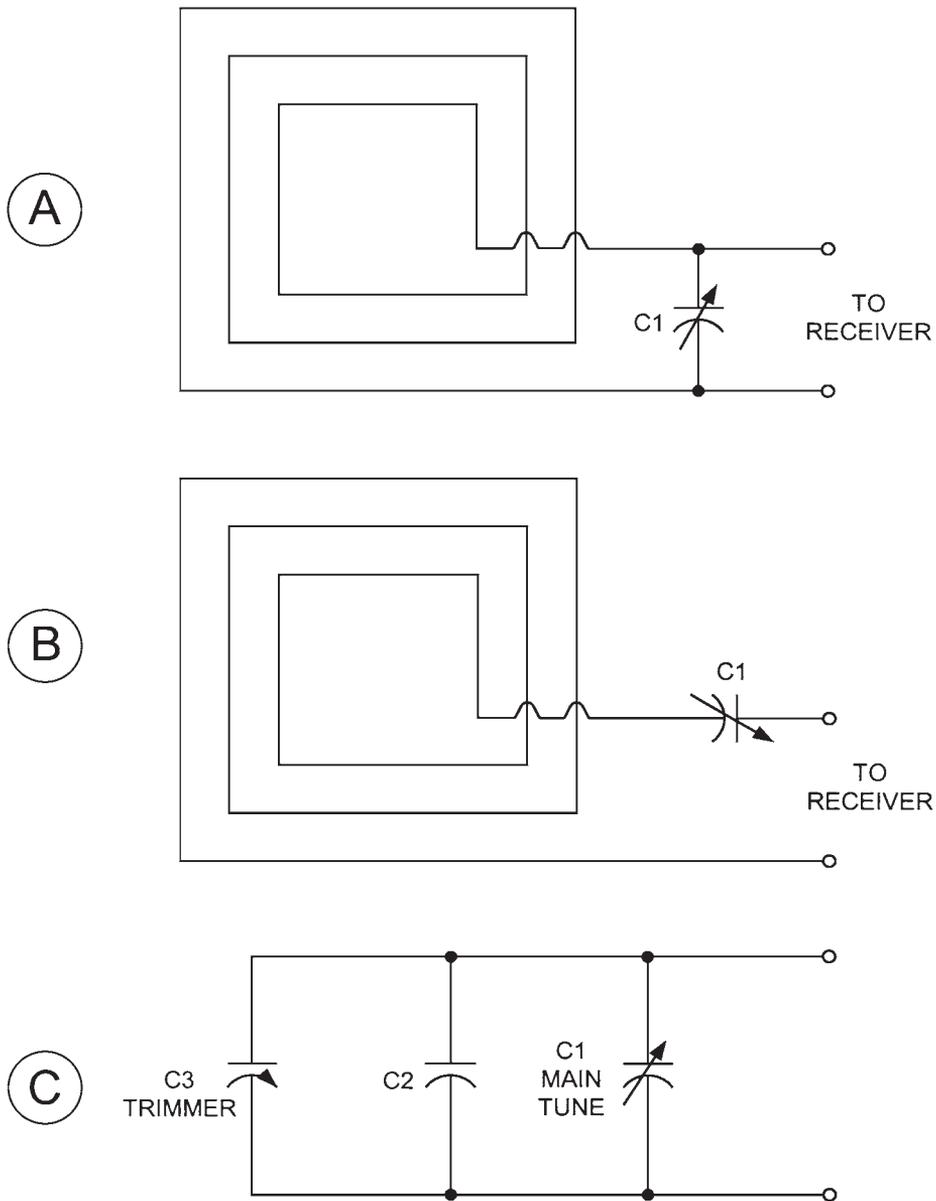
Loop performance is greatly enhanced by tuning the inductance of the loop to the desired frequency. The bandwidth of the loop is reduced, which reduces front-end overload. Tuning also increases the signal level available to the receiver by a factor of 20 to 100 times. Although tuning can be a bother if the loop is installed remotely from the receiver, the benefits are well worth it in most cases.

There are several different schemes available for tuning, and these are detailed in Fig. 15-6. The parallel tuning scheme, which is by far the most popular, is shown in Fig. 15-6A. In this type of circuit, the capacitor ( $C_1$ ) is connected in parallel with the inductor, which in this case is the loop. Parallel resonant circuits have a very high impedance to signals on their resonant frequency and a very low impedance to other frequencies. As a result, the voltage level of resonant signals is very much larger than the voltage level of off-frequency signals.

The series resonant scheme is shown in Fig. 15-6B. In this circuit, the loop is connected in series with the capacitor. A property of series resonant circuits is that they offer a high impedance to all frequencies except the resonant frequency (exactly the opposite of the case of parallel resonant circuits). As a result, current from the signal will pass through the series resonant circuit at the resonant frequency, but off-frequency signals are blocked by the high impedance.

There is a wide margin for error in the inductance of loop antennas, and even the precise-looking equations to determine the required values of capacitance and inductance for proper tuning are actually only estimations. The exact geometry of the loop “as built” determines the actual inductance in each particular case. As a result, it is often the case that the tuning provided by the capacitor is not as exact as desired, so some form of compensation is needed. In some cases, the capacitance required for resonance is not easily available in a standard variable capacitor, and some means must be provided for changing the capacitance. Figure 15-6C shows how this is done. The main tuning capacitor can be connected in either series or parallel with other capacitors to change the value. If the capacitors are connected in parallel, then the total capacitance is increased (all capacitances are added together). If the extra capacitor is connected in series, however, then the total capacitance is reduced. The extra capacitors can be switched in and out of a circuit to change frequency bands.

Tuning of a remote loop can be a bother if it is done by hand, so some means must be found to do it from the receiver location (unless you enjoy climbing into the attic or onto the roof). Traditional means of tuning called for using a low-rpm dc motor, or stepper motor, to turn the tuning capacitor. A very popular combination was the little 1- to 12-rpm motors used to drive rotating displays in retail store show



15-6 Various tuning schemes: (A) parallel, (B) series resonant, (C) series or parallel.

windows. But this approach is not really needed today. We can use varactor voltage-variable capacitance diodes to tune the circuit.

A varactor works because the junction capacitance of the diode is a function of the applied reverse-bias voltage. A high voltage (such as 30 V) drops the capacitance, whereas a low voltage increases it. Varactors are available with maximum capacitances of 22, 33, 60, 100, and 400 pF. The latter are of most interest to us because they have the same range as the tuning capacitors normally used with loops.

Figure 15-7 shows how a remote tuning scheme can work with loop antennas. The tuning capacitor is a combination of a varactor diode and two optional capacitors: a fixed capacitor ( $C_1$ ) and a trimmer ( $C_2$ ). The dc tuning voltage ( $V_t$ ) is provided from the receiver end from a fixed dc power supply ( $V+$ ). A potentiometer ( $R_1$ ) is used to set the voltage to the varactor, hence also to tune the loop. A dc blocking capacitor ( $C_3$ ) keeps the dc tuning voltage from being shorted out by the receiver input circuitry.

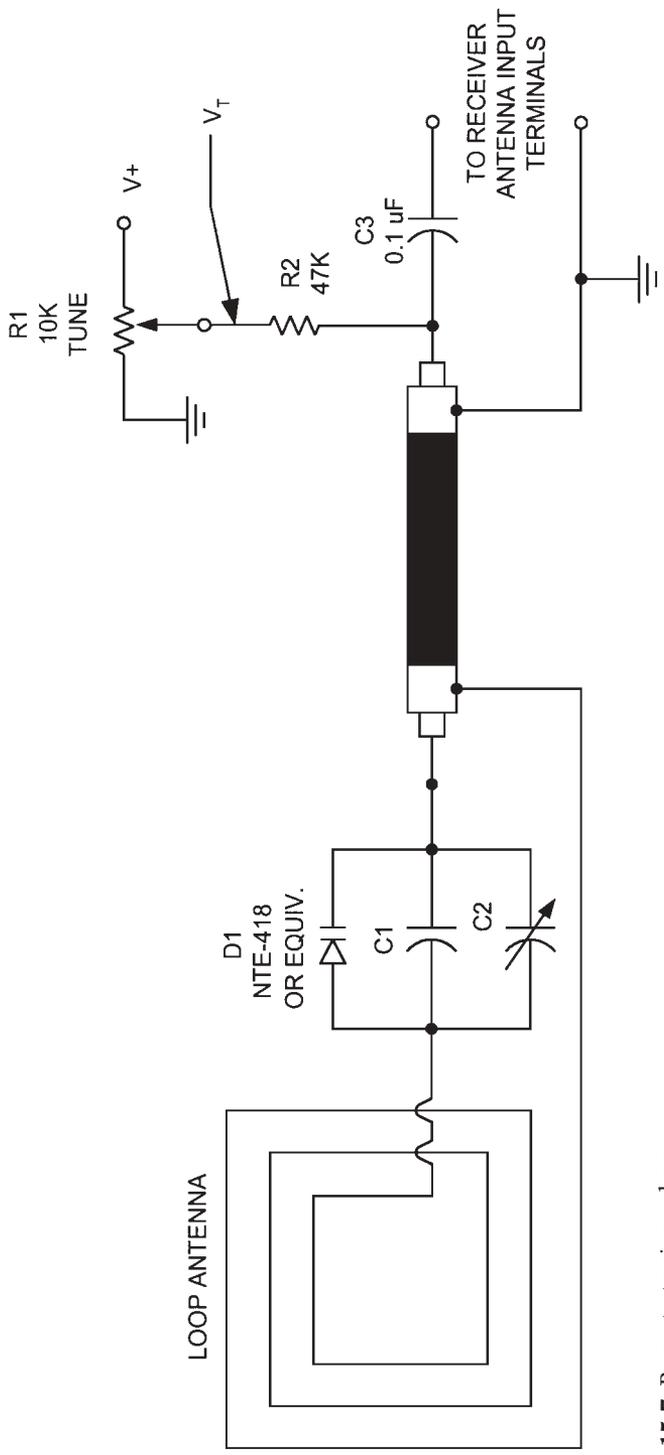
## The sports fan's loop

OK, sports fans, what do you do when the best game of the week is broadcast only on a low-powered AM station and you live at the outer edge of their service area where the signal strength leaves much to be desired? You use the sports fan's loop antenna, that's what! I first learned of this antenna from a friend of mine, a professional broadcast engineer, who worked at a religious radio station that had a pip-squeek signal but lots of fans. It really works—*one might say it's a miracle*.

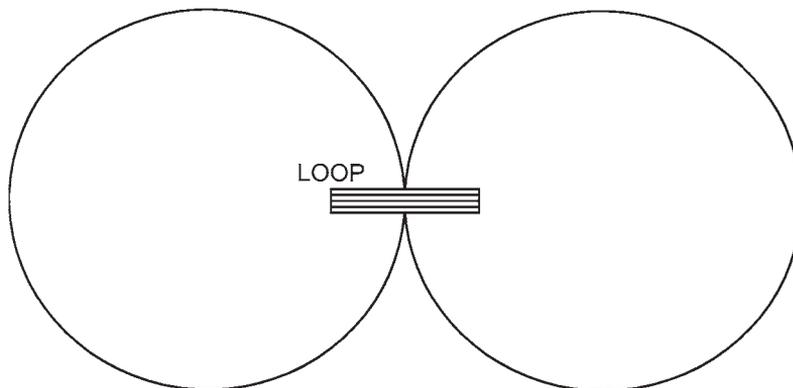
The basic idea is to build a 16-turn, 60-cm<sup>2</sup> tuned loop and then place the AM portable radio at the center so that its loopstick is aimed such that its null end is broadside of the loop. When you do so, the nulls of both the loop and the loopstick are in the same direction. The signal will be picked up by the loop and then coupled to the radio's loopstick antenna. Sixteen-conductor ribbon cable can be used for making the loop. For an extra touch of class, place the antenna and radio assembly on a dining room table lazy Susan to make rotation easier. A 365-pF tuning capacitor is used to resonate the loop. If you listen to only one station, then this capacitor can be a trimmer type.

## Shielded loop antennas

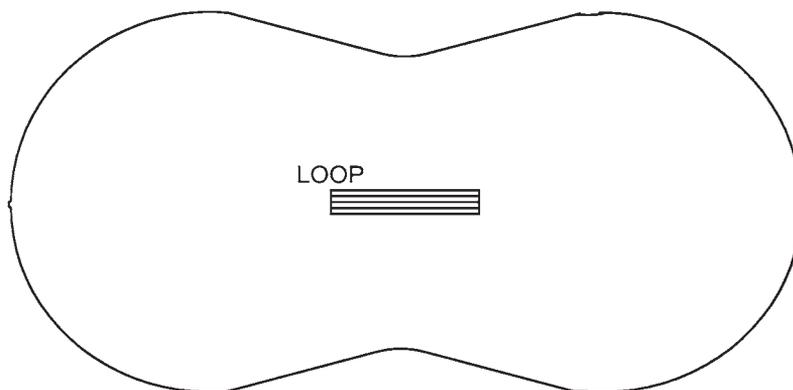
The loop antennas discussed thus far in this chapter have all been unshielded types. Unshielded loops work well under most circumstances, but in some cases their pattern is distorted by interaction with the ground and nearby structures (trees, buildings, etc.). In my own tests, trips to a nearby field proved necessary to measure the depth of the null because of interaction with the aluminum siding on my house. Figure 15-8 shows two situations. In Fig. 15-8A we see the pattern of the normal "free space" loop, i.e., a perfect figure-8 pattern. When the loop interacts with the nearby environment, however, the pattern distorts. In Fig. 15-8B we see some filling of the notch for a moderately distorted pattern. Some interactions are so severe that the pattern is distorted beyond all recognition.



15-7 Remote tuning scheme.



**15-8A** A normal "free space" loop.



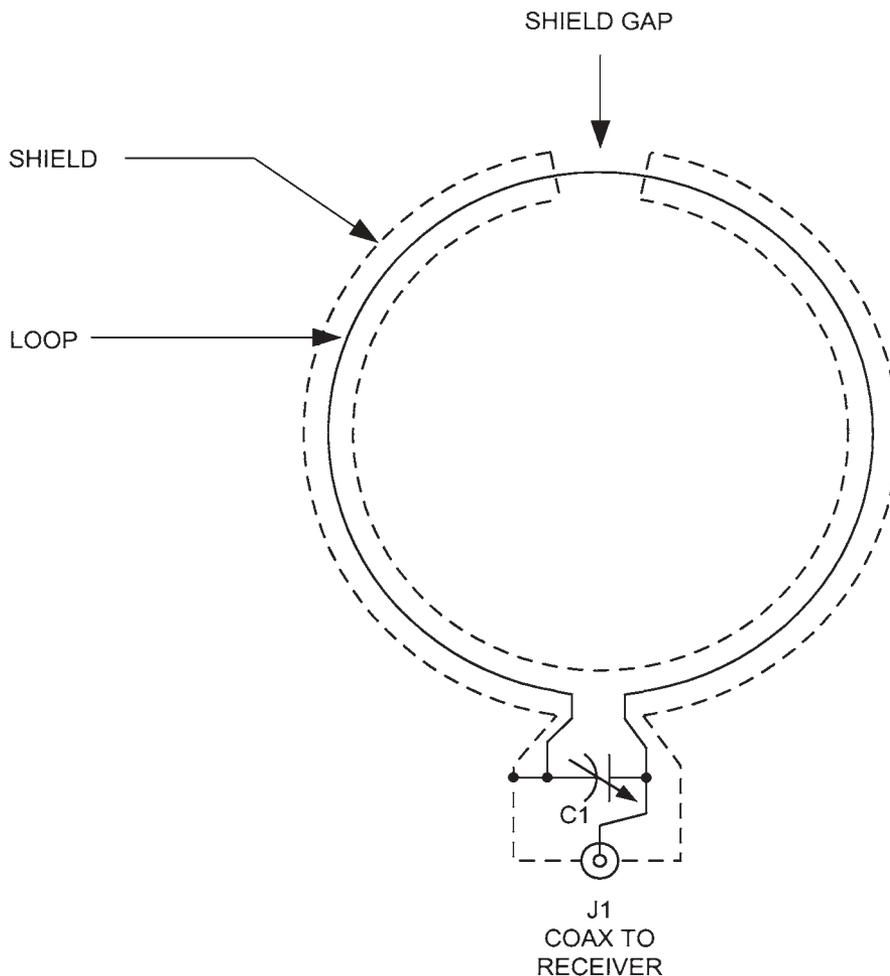
**15-8B** Filling of the notch.

The solution to the problem is to reduce interaction by shielding the loop, as in Fig. 15-9. Loop antennas operate on the magnetic component of the electromagnetic wave, so the loop can be shielded against voltage signals and electrostatic interactions. In order to prevent harming the ability to pick up the magnetic field, a gap is left in the shield at one point.

There are several ways to shield a loop. You can, for example, wrap the loop in adhesive-backed copper-foil tape. Alternatively, you can wrap the loop in aluminum foil and hold it together with tape. Another method is to insert the loop inside a copper or aluminum tubing frame. Or—the list seems endless.

## Using a loop antenna

Most readers will use a loop for DXing rather than hidden transmitter hunting, navigation, or other RDF purposes. For the DXer, there are actually two uses for the



**15-9** Shielding the loop.

loop. One is when you are a renter or live in a community that has routine covenants against outdoor antennas. In this situation, the loop will serve as an active antenna for receiving AM BCB and other low-frequency signals without the neighbors or landlord becoming PFJs (“purple-faced jerks”).

The other use is illustrated by the case of a friend of mine. He regularly tunes in clear channel WSM (650 kHz, Nashville) in the wee hours between Saturday evening (“Grand Ole Opry” time) and dawn. However, this “clear” channel of WSM is not really so clear, especially without a narrow filter in the receiver. He uses a loop antenna to null out a nearby 630-kHz signal that made listening a bit dicey and can now tape his 1940s–1950s vintage country music.

It is not necessary to place the desired station directly in the main lobes off the ends of the antenna but rather to place the nulls (broadside) in the direction of the

offending station that you want to eliminate. So what happens if the offending station and the desired station are in a direct line with each other and your receiving location is in the middle between them? Both nulls and lobes on a loop antenna are bidirectional, so a null on the offending station also will null the desired station in the opposite direction.

One method is to use a sense antenna to spoil the pattern of the loop to a cardioid shape. Another method is to use a spoiler loop to null the undesired signal. The spoiler loop is a large box loop placed 1 to 3 ft (found experimentally) behind the reception loop in the direction of the offending signal. This method was first described by Levintow and is detailed in Fig. 15-10. The small loopstick may be the antenna inside the receiver, whereas the large loop is a box loop such as the sports fan's loop. The large box loop is placed about 33 to 100 cm behind the loopstick and in the direction of the offending station. The angle with respect to the line of centers should be  $60^\circ$  to  $90^\circ$ , which also was found experimentally. It is also possible to use two air core loops to produce an asymmetrical receiving pattern.

## Sharpening the loop

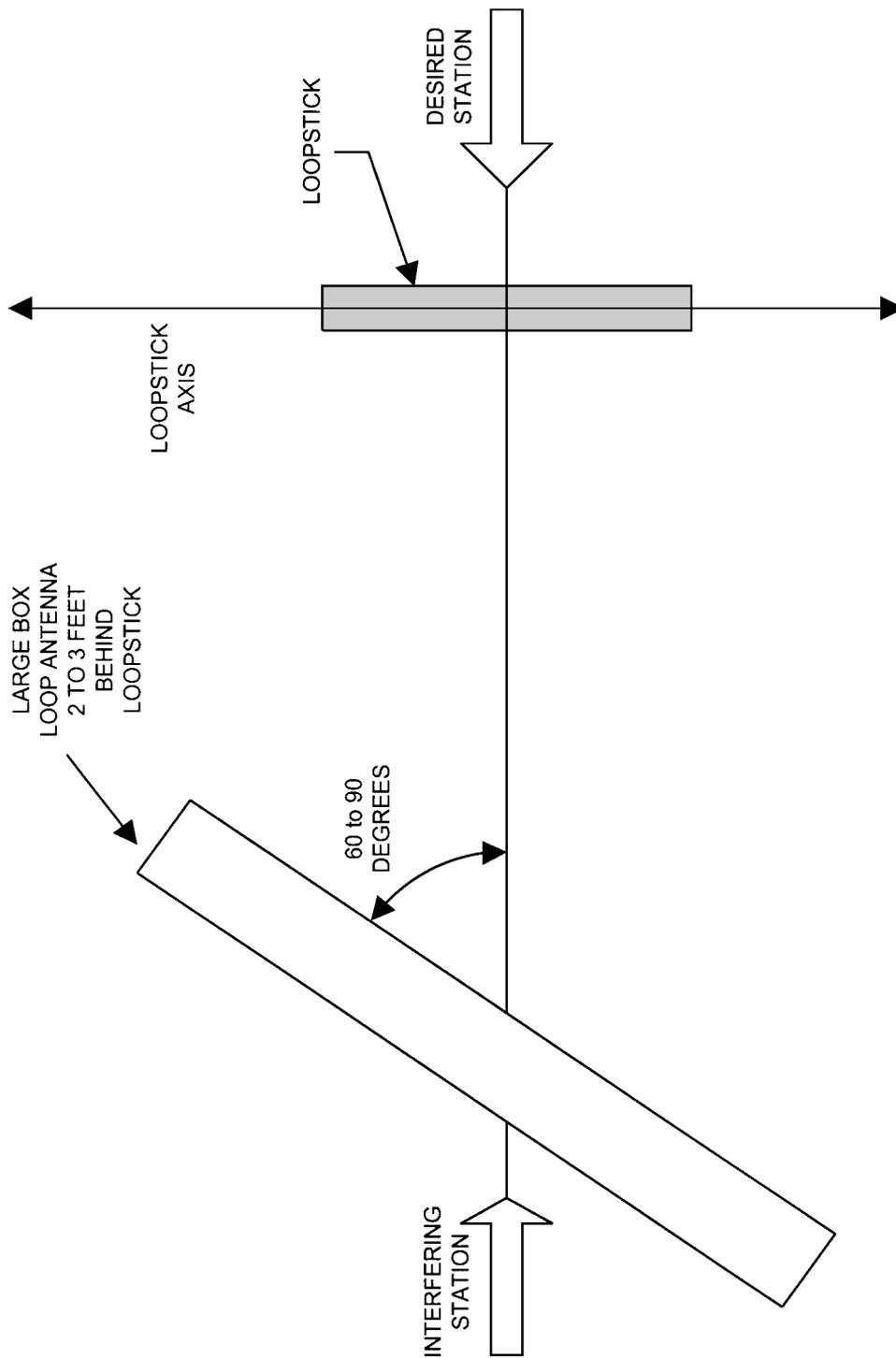
Many years ago, the *Q-multiplier* was a popular add-on accessory for a communications receiver. These devices were sold as Heathkits, and many construction projects could be found in magazines and amateur radio books. The Q-multiplier has the effect of seeming to greatly increase the sensitivity of a receiver, as well as greatly reducing the bandwidth of the front end. Thus it allows better reception of some stations because of increased sensitivity and narrowed bandwidth.

A Q-multiplier is an active electronic circuit placed at the antenna input of a receiver. It is essentially an Armstrong oscillator, as shown in Fig. 15-11, that does not quite oscillate. These circuits have a tuned circuit ( $L_1/C_1$ ) at the input of an amplifier stage and a feedback coupling loop ( $L_3$ ). The degree of feedback is controlled by the coupling between  $L_1$  and  $L_3$ . The coupling is varied by varying both how close the two coils are and their relative orientation with respect to each other. Certain other circuits use a series potentiometer in the  $L_3$  side that controls the amount of feedback.

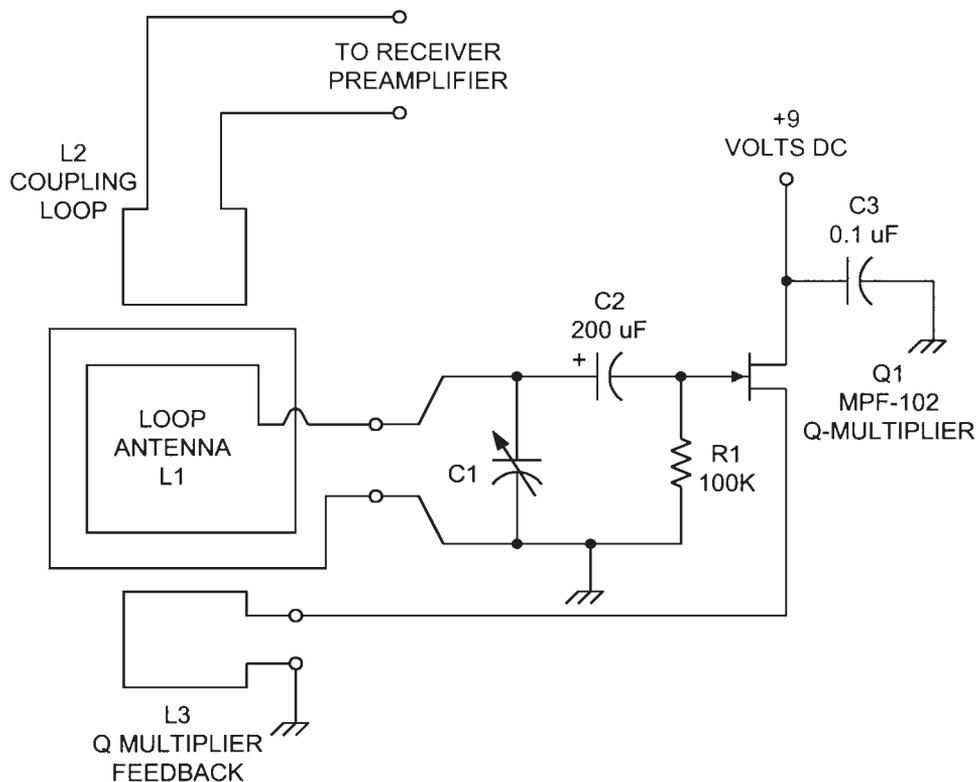
The Q-multiplier is adjusted to the point that the circuit is just on the verge of oscillating, but not quite. As the feedback is backed away from the threshold of oscillation, but not too far, the narrowing of bandwidth occurs, as does the increase in sensitivity. It takes some skill to operate a Q-multiplier, but it is easy to use once you get the hang of it and is a terrific accessory for any loop antenna.

## Loop amplifier

Figure 15-12 shows the circuit for a practical loop amplifier that can be used with either shielded or unshielded loop antennas. It is based on junction field effect transistors (JFET) connected in cascade. The standard common-drain configuration is used for each transistor, so the signals are taken from the source terminals. The drain terminals are connected together and powered from the +12-V dc power



**15-10** Use of a spoiler loop to null an undesired signal.



15-11 A Q-multiplier.

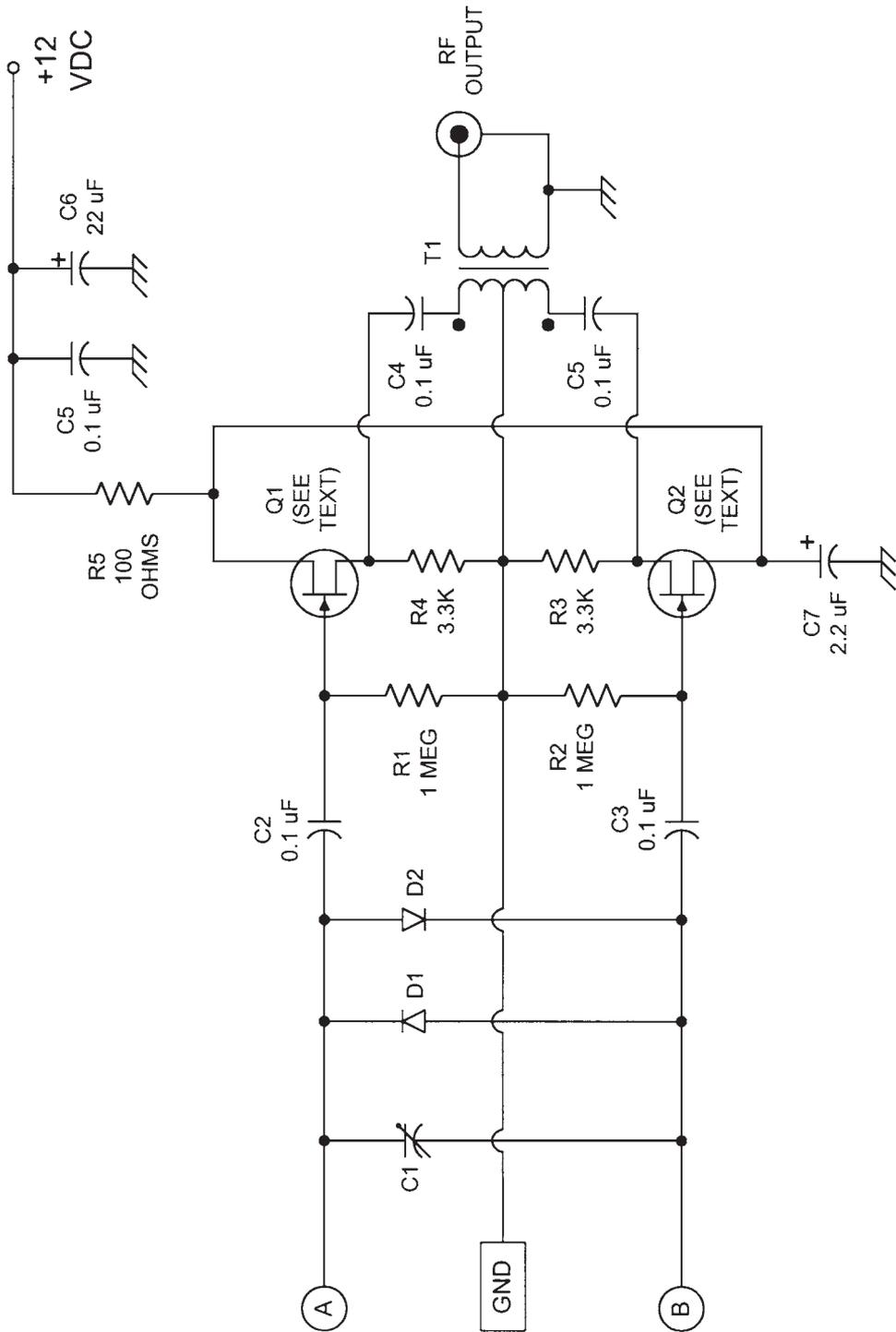
supply. A  $2.2\text{-}\mu\text{F}$  bypass capacitor is used to put the drain terminals of  $Q_1$  and  $Q_2$  at ground potential for ac signals while keeping the dc voltage from being shorted out.

The two output signals are applied to the primary of a center-tapped transformer, the center tap of which is grounded. To keep the dc on the source terminals from being shorted through the transformer winding, a pair of blocking capacitors ( $C_4, C_5$ ) is used.

The input signals are applied to the gate terminals of  $Q_1$  and  $Q_2$  through dc blocking capacitors  $C_2$  and  $C_3$ . A pair of diodes ( $D_1, D_2$ ) is used to keep high-amplitude noise transients from affecting the operation of the amplifier. These diodes are connected back to back in order to snub out both polarities of signal.

Tuning capacitor  $C_1$  is used in lieu of the capacitor in the loop and is used to resonate the loop to a specific frequency. Its value can be found from the equation given earlier.

The transistors used for the push-pull amplifier ( $Q_1, Q_2$ ) can be nearly any general-purpose JFET device (MPF-102, MPF-104, etc.). A practical approach for many people is to use transistors from service replacement lines, such as the NTE-312 and NTE-316 devices.

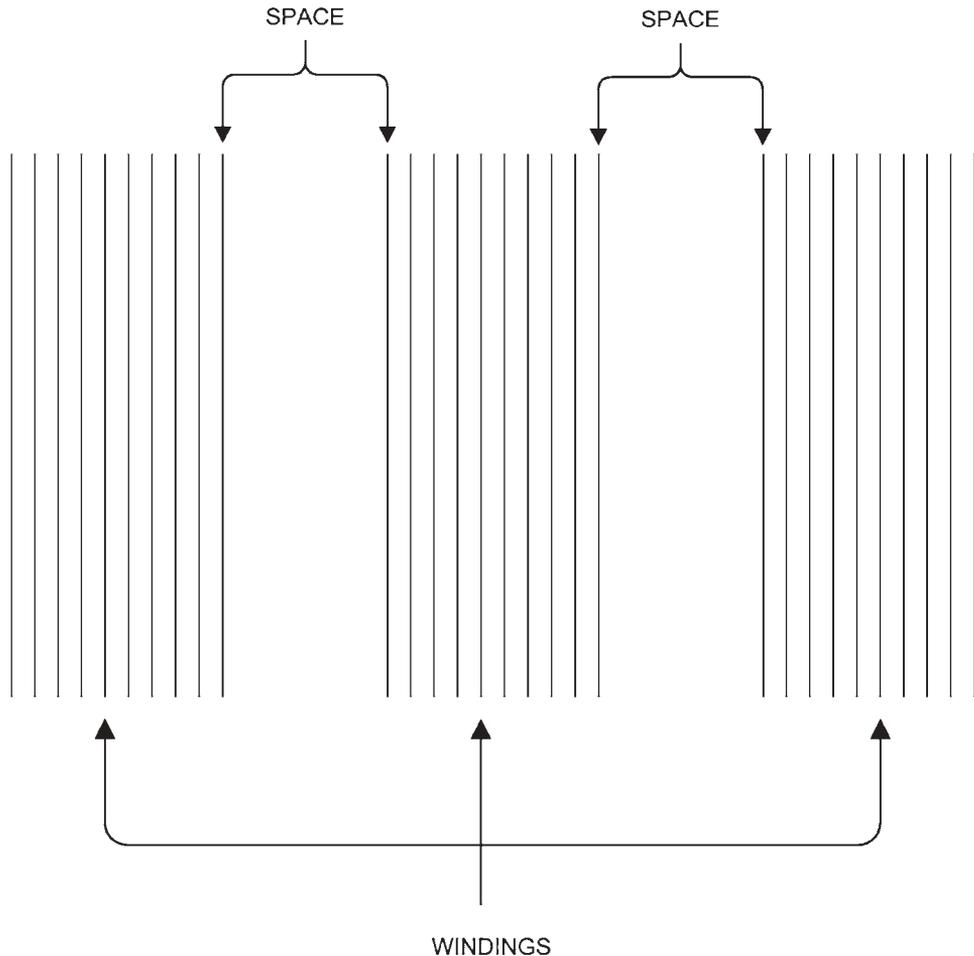


15-12 A practical loop amplifier.

## Special problem for VLF/LF loops

A capacitance is formed whenever two conductors are side by side. A coil produces capacitance as well as inductance because the turns are side by side. Unfortunately, with large multiturn loops, this capacitance can be quite large. The “distributed capacitance” of the loop causes a self-resonance with the inductance. The loop does not work well at frequencies above the self-resonant point, so it is sometimes important to raise the self-resonance to a point where it does not affect operation at the desired frequencies.

Figure 15-13 shows a solution that raises the self-resonant point. The turns are broken into two or more groups and separated by a space. This method reduces the effective capacitance by placing the capacitances of each group of wires in series with the others.

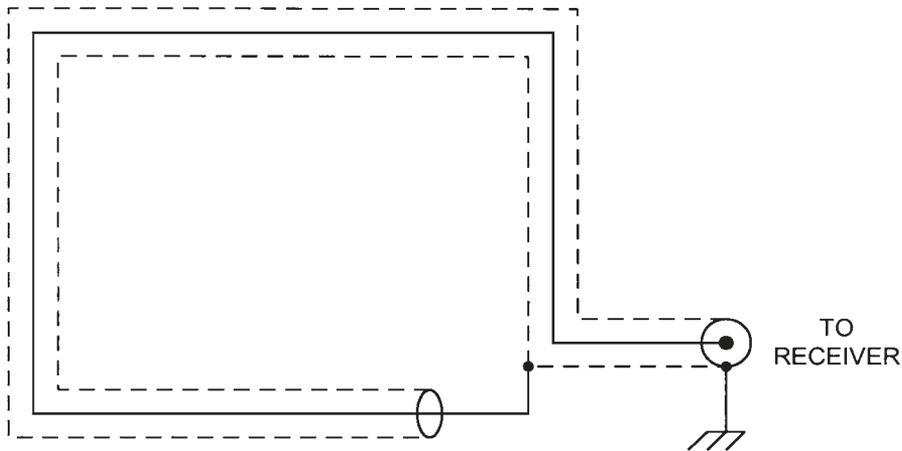


**15-13** Raising the self-resonant point.

## Coaxial-cable loop antennas

One of the more effective ways to make a shielded loop is to use coaxial cable. Figure 15-14 shows the circuit of such a loop. Although only a single-turn loop is shown, there can be any number of turns. One reader made a 100-kHz LORAN (a navigation system) loop using eight turns of RG-59/U coaxial cable on an 8-ft diameter.

Note the special way that the coaxial cable is connected. This method is called the *Faraday connection* after the fact that the shield of the coax forms a *Faraday shield*. At the output end, the center conductor of the coaxial cable is connected to the center conductor of the coaxial connector. The coax shield is connected to the connector ground/shield terminal. At the other end of the loop, the shield is left floating, but the center conductor is connected to the shield *at the connector*; not at just any convenient point.



**15-14** A coaxial-cable shielded loop.

# 16

Chapter

## Small transmitting loop antennas

Small transmitting loops are those which are less than one-third wavelength ( $\lambda/3$ ) according to Hart (1986) or between  $0.04\lambda$  and  $0.10\lambda$  according to Belrose (1993). Other sources claim that loops up to about  $0.20\lambda$  are small. The key characteristic is that the current is equal in all points of the loop rather than varying with loop length the way the current in a large loop (or other antenna) does.

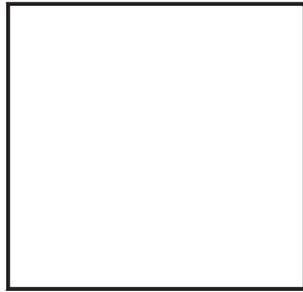
One of the parameters of the small loop antenna is a very low radiation resistance (it is a function of the area of the loop). The radiation resistance of the small loop antenna is less than  $1 \Omega$  and often considerably less than this figure (e.g.,  $0.05 \Omega$ ). This means that the copper or aluminum losses of the antenna must be kept very, very low. For such reasons, it is common to assemble the small transmitting loop antenna from 1- to 4-in copper or aluminum pipe. For this reason, the square (Fig. 16-1A) or octagon (Fig. 16-1B) shapes are preferred. The reason this is true is purely mechanical: The square loop can be built with  $90^\circ$  elbow joints, and the octagon can be built with  $45^\circ$  elbow joints. For antennas whose circumference is greater than  $\lambda/8$ , standard  $3/4$ -in copper pipe can be used in construction, but for smaller-sized loops, larger-sized copper piping is used.

The small transmitting loop tends to be highly inductive in its reactance. Values of inductive reactance to  $1000 \Omega$  are not unusual, although closer to  $100 \Omega$  is the norm. Because of this inductive reactance, the small transmitting loop antenna is tuned by one or more series capacitors.

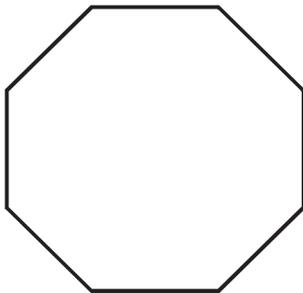
A sample loop was modeled by Belrose (1993). It was 1 m in diameter and was made of 2.54-cm (1-in) copper pipe. For this loop, the gains were

10 MHz	2.88 dBi
14 MHz	0.22 dBi
21 MHz	1.14 dBi
30 MHz	1.42 dBi

Note that the gains were all less than that of a dipole (12.15 dBi) but in the upper regions of the frequency spectrum were quite decent. Even at 10 MHz, the 1-m



(A)



(B)

**16-1** Small transmitting loop antennas: (A) square, (B) octagon.

small loop antenna was only  $-5.03$  dBi less than the dipole. The loop probably would show similar gains up to the point of self-resonance.

Calculations of the current and voltage within the loop are substantial. The loop impedance at 10 MHz is  $0.088 \ 1 \ j161 \ \Omega$ . This means that a 150-W transmitter, certainly modest for these frequencies, would produce 41 A circulating inside the loop and a voltage across the tuning capacitor of 6600 V. As Belrose (1993) warns: “Don’t touch your loop when transmitting!”

The tuning capacitor in Belrose’s loop should be rated for at least 10,000 V to guard against anyone using more than 150 W. Keep in mind that a power of 500 W will produce potentials across the capacitor of 30,000 V, so you will need a capacitor with at least  $\frac{1}{2}$ -in spacing. A 1-in spacing is good for about 75,000 V. It is probably best to use a vacuum-variable capacitor for the tuning capacitor of a small loop antenna if you can afford one. Regardless, the loop power should be kept under control!

The  $Q$  of the antenna at any given resonant frequency is very, very high. For the 1-m loop antenna described earlier, the  $Q$  at 10 MHz was 1824, which resulted in a bandwidth of 5.5 kHz (although in practice it is twice this amount before VSWR tol-

erances on the transmitter are exceeded). Regardless, the bandwidth of the resonated loop is tiny, so some means must be provided for retuning the loop as you shift frequency. If you do not shift frequency, then the loop can be tuned and forgotten, but for the rest of the radio fraternity, perhaps a remotely tuned capacitor is in order. Getting a large capacity variable in the voltages that are required for operation with a good power level is a bit tricky.

## Design equations

The design equations developed by Hart (1986) are as follows:

*Radiation resistance* ( $R_R$ ):

$$R_R = 3.38 \times 10^{-8} (F^2 A)^2 \quad [16.1]$$

*Loss resistance* ( $R_L$ ):

$$R_L = \frac{9.96 \times 10^{-4} S \sqrt{F}}{D} \quad [16.2]$$

*Efficiency* ( $\eta$ ):

$$\eta = \frac{R_R}{R_R + R_L} \quad [16.3]$$

*Inductance* ( $L$ ):

$$L = 1.9 \times 10^8 S (7.353 \log_{10} \frac{96S}{\pi D} - 6.386) \quad [16.4]$$

*Inductive reactance* ( $X_L$ ):

$$X_L = 2\pi FL \times 10^6 \quad [16.5]$$

*Tuning capacitance* ( $C$ ):

$$C = \frac{1}{2\pi FX_L \times 10^6} \quad [16.6]$$

*Quality factor* ( $Q$ ):

$$Q = \frac{F}{\Delta F} = \frac{X_L}{2(R_R + R_L)} \quad [16.7]$$

*Bandwidth* ( $\Delta F$ ):

$$\Delta F = \frac{F}{Q} \quad [16.8]$$

*Distributed capacitance* ( $C_D$ ):

$$C_D = 0.82S \quad [16.9]$$

*Capacitor voltage* ( $V_C$ ):

$$V_C = \sqrt{P X_L Q} \quad [16.10]$$

where

$A$  is the area of the loop, in square feet (ft<sup>2</sup>)

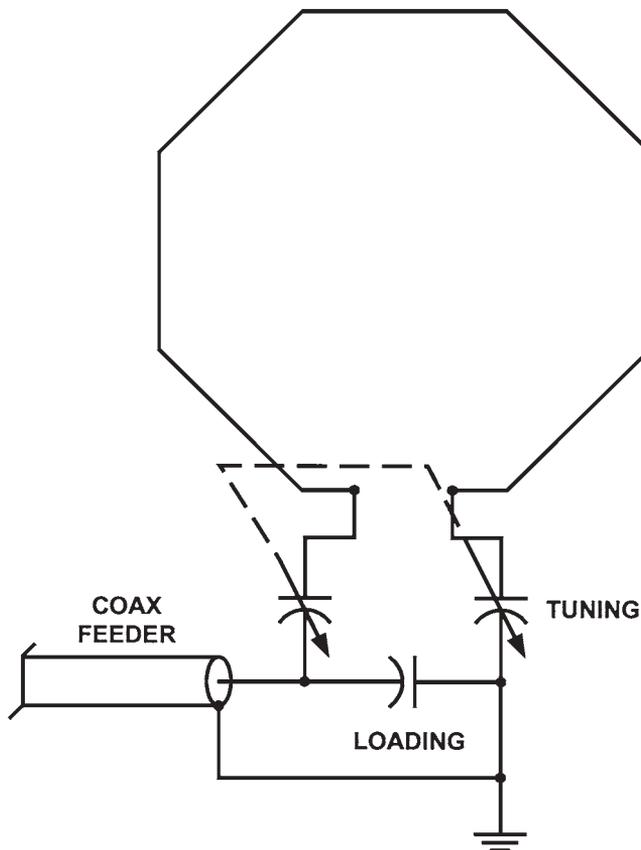
$S$  is the length of the loop conductor, in feet (ft)

$D$  is the diameter of the conductor, in inches (in)

$P$  is the applied transmitter output power, in watts (W)

## Some sample loop antennas

The Patterson (1967) loop antenna is shown in Fig. 16-2. It is made from 1.5-in copper tubing. Segments are cut and are joined together by eight 45° elbow joints, giving the octagon shape. Each segment is 0.5 ft long. The tuning is accomplished by three capacitors, two of which are a split-stator unit (two capacitors on the same shaft). The tuning control is the split-stator capacitor, whereas the loading control is a single capacitor.



**16-2** A Patterson loop antenna.

Another loop antenna is shown in Fig. 16-3. This antenna has a Faraday feed system rather than a capacitor-coupled feed system, as did the Patterson loop. The tuning is accomplished by  $C_A$ , the series tuning capacitor. The requirement must be met that  $D_1/D_2 = 5$ .

Still another loop is shown in Fig. 16-4. This loop relies on magnetic coupling to perform the coupling of the transmitter. In this loop antenna, the coupling is via a small coupling loop and 50- $\Omega$  coaxial cable to the transmitter. Capacitor  $C_1$  is used to resonate the loop, whereas capacitors  $C_2$  and  $C_3$  serve the purposes of loading and resonating the coupling loop. According to Mozzochi (1993), the voltages and currents with respect to the capacitors are

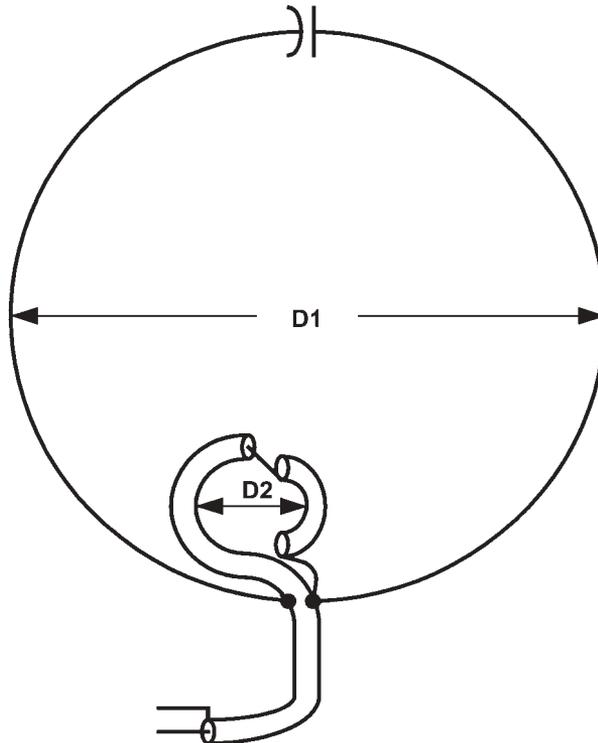
$$V_{C_1} = 1800 \sqrt{P} \quad [16.11]$$

$$V_{C_2} = 18 \sqrt{P} \quad [16.12]$$

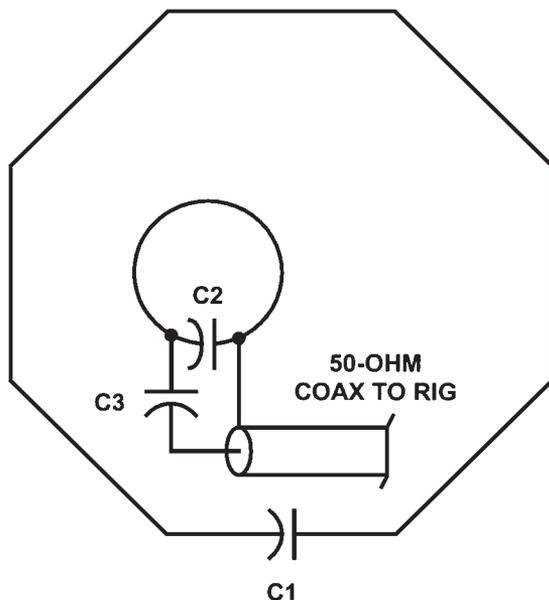
$$V_{C_3} = 17 \sqrt{P} \quad [16.13]$$

$$I_{C_1} = 2.2 \sqrt{P} \quad [16.14]$$

$$I_{C_2} = 0.21 \sqrt{P} \quad [16.15]$$



**16-3** A loop antenna with a Faraday feed system.



**16-4** A loop antenna with magnetic coupling.

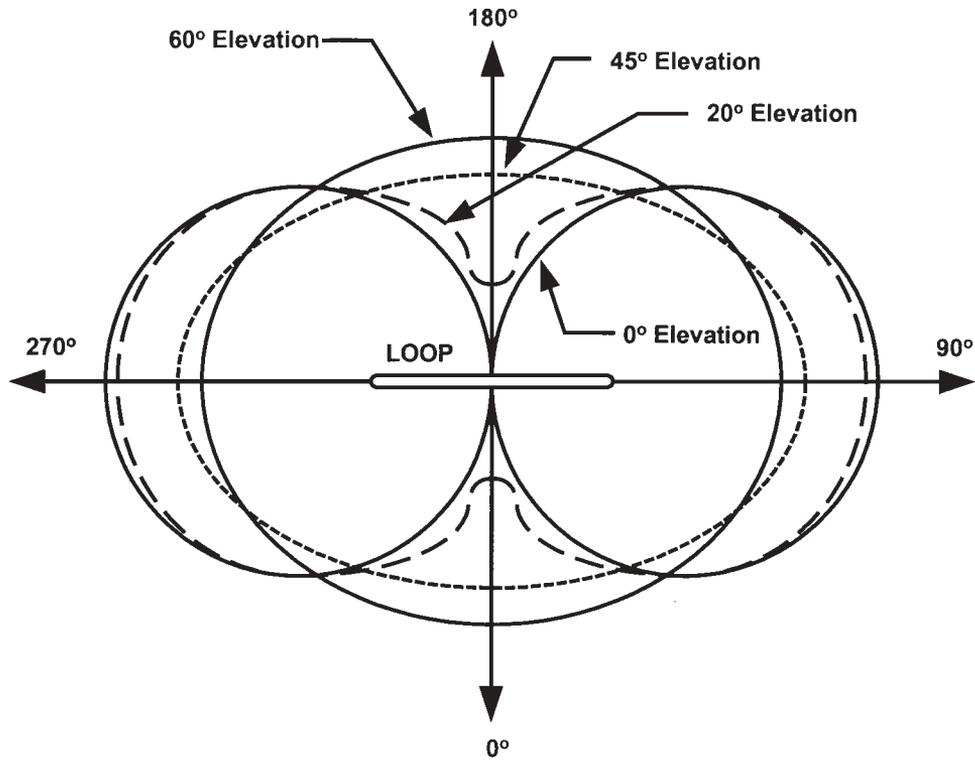
$$I_{C3} = 0.15 \sqrt{P} \quad [16.16]$$

The radiation patterns for the loop antenna are shown in Fig. 16-5. Note that four elevations are given ( $0^\circ$ ,  $20^\circ$ ,  $45^\circ$ , and  $60^\circ$ ).

The bottom line is that small transmitting loop antennas are not very good for those who can afford to put up a better antenna, but for those whose tight quarters permit only a small transmitting loop, they are quite viable.

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**16-5** Radiation patterns.

Mozzochi, Charles J. (1993), "A Small Loop Antenna for 160 Meters," *QST*, pp. 32-34.

Orr, William (1994), "Feedback on Hart's article," *QST*, July.

Patterson, K. (1967), "Down-to-Earth Army Antennas," *Electronics*, August 21, pp. 111-114.

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# 17

CHAPTER

# Antenna modeling software

One of the significant contributions of computer technology to antenna design is the improvement of modeling and simulation. Modeling and simulation are used in a wide variety of applications, including management, science, and engineering. We can now model just about any process, any device, or any circuit that can be reduced mathematically. The purpose of modeling, at least in engineering, is to do the design cheaply on the computer before “bending metal.” The old “cut and try” method works, to be sure, but it is costly in time and money (two things perpetually in short supply). If performance issues and problems can be solved on a computer, then we are time and money ahead of the game. Also, modeling and simulation make it possible to look at more alternatives and to gauge the effect of a change in an antenna design before the change is made.

Modeling is used not just in engineering but also in other processes. Organizations and management consultants use modeling and simulation to study their operations and find the most efficient way of doing things. I saw an example where an inefficient truck terminal and warehouse operation was improved significantly by modeling on a Windows desktop computer. The operators modeled the existing system and several alternative systems, and found that one particular rule (involving bulk deliver trucks such as tankers and dry bulk carriers) was jamming up the normal package deliveries at the warehouse dock. They could have tried different sets of rules over a couple of years, collected lots of data, presented scores of viewgraphs to management committees, and still have overlooked the actual cause of the problem. The model showed the problem in only a few minutes, however.

## What is a model?

Models are no big mystery. If you have fingers and can count, then you can make models. In fact, you probably do make models mentally, e.g., when you look at how much cash you have in your pocket when confronted with several different purchase choices but cannot afford all of them. The “model” is the “what if” considerations that lead to your selection.

Youngsters in the schools are learning that the equations and formulas they learn in algebra and trigonometry are “models.” This designation was not used when I was in school, but it is an apt description in most cases.

One of the earliest modeling methods for the masses was the computer-based spreadsheet program. Software such as Microsoft Excel can be used to test various scenarios by entering different values into the cells. I have seen spreadsheets used for the originally intended business applications, as well as engineering applications (including antenna design).

Antenna modeling software tends to use the “method of moments,” in which each wire in the system is broken into 1 to  $N$  segments, and the current in each segment is calculated.

## Some common programs

Some years ago an antenna modeling program came on the scene. Developed by the Navy, this program used numerical methods to perform the complex calculations needed to model an antenna. Called the *Numerical Electromagnetic Computation (NEC) program*, a smaller version soon became available known as miniNEC (which runs on PCs). You can download miniNEC-3 from several sources in the amateur radio newsgroups in the United States or United Kingdom, or from the Navy Postgraduate School Web site. A miniNEC-4 version is also available, but I understand that it requires a license to use and cannot be exported without a permit from the U.S. government.

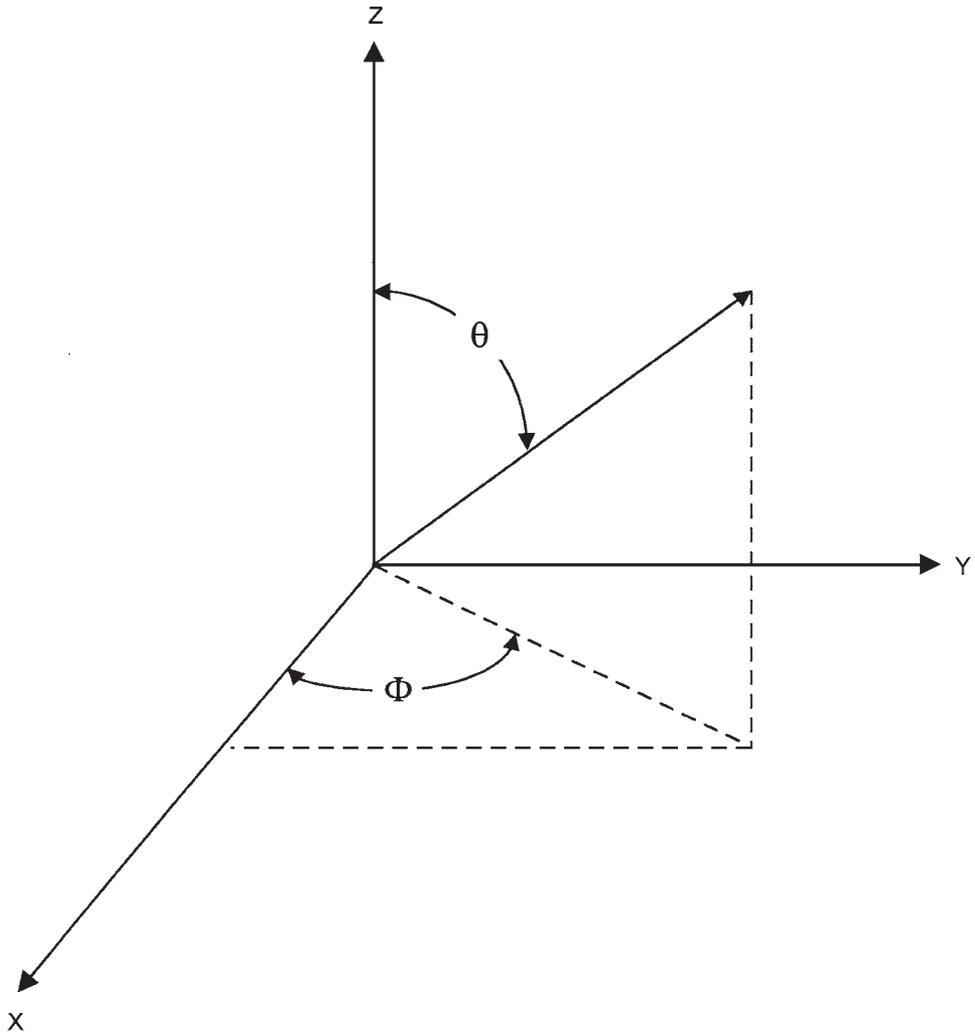
There are also a number of private vendors who offer the miniNEC at a price. Some of the better vendors offer their own versions, based on the miniNEC concept, but with some value added to make it worthwhile to buy theirs instead of using the “freebie” version.

The miniNEC programs model wire antennas using the coordinate system shown in Fig. 17-1. This is a standard three-dimensional Cartesian coordinate system in which the  $y$  axis is horizontal, the  $z$  axis is vertical, and the  $x$  axis is orthogonal to both the  $y$  and  $z$  axes. The idea is to specify the starting and ending points of the wires making up the antenna in terms of the  $xyz$  coordinate system. Figure 17-2 shows a single-wire antenna laid out in the  $x$  axis. This axis is in the horizontal plane but orthogonal to the  $y$  axis. The antenna designer will specify  $X_1$  and  $X_2$ , which are the starting and ending points of the wire.

The antenna can consist of a large number of wires, although the larger the number, the longer is the processing time. Each wire is broken into a number of different segments for calculation purposes.

### NEC-4

The NEC-4 software is the latest in the NEC series and is proprietary to the University of California, from whom a license must be obtained to use it. Because of national defense considerations, an export license is required to take NEC-4 overseas. The NEC-4 software will produce modeling of underground radials, elements of varying diameter, and carefully constructed close-spaced parallel wires.



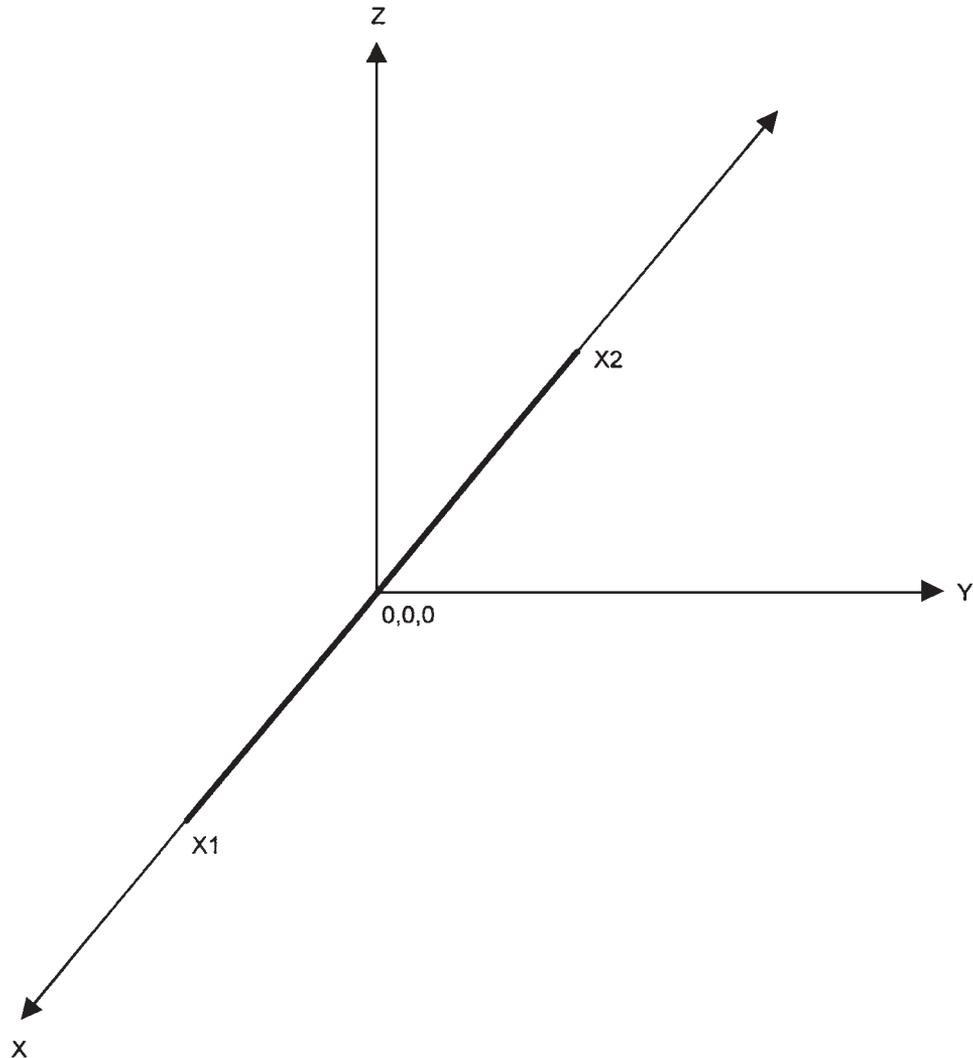
17-1 XYZ coordinate system for NecWin modeling software.

### **EZNEC Professional**

Roy Lewallen, W7EL, provides the EZNEC Pro software. This software has an option for NEC-4, provided that the license is obtained from the University of California (again, export restrictions apply).

### **GNEC**

Nittany-Scientific makes a version of NEC-4 known as GNEC. It permits the use of catenary wires, helices, networks, and other possibilities. Outputs include 3-D, polar plots, and rectangular plot graphs, as well as a large array of tabular reports.



**17-2** Example of a single wire laid out on *XYZ* coordinate system.

### **NEC-2**

This is a high-capability version of the NEC software, and it is in the public domain. It is limited to antenna elements of constant diameter, although some software providers who base their wares on NEC-2 provide corrections for multidiameters. It is fitted with the Sommerfield-Norton ground model for modeling horizontal wires close to the earth's surface.

### **NECWin Plus**

This software is provided by Nittany-Scientific. It features true spreadsheet geometry. The software also offers stepped-diameter corrections, a gain-averaging test, a

CAD file (\*.dxf), 2-D and 3-D plots, 2-D and 3-D antenna views, and certain graphic outputs. There are two versions, NECWin Pro and NECWin Basic (NECWin Basic is reviewed in this chapter).

### **EZNEC for Windows**

Roy Lewallen's EZNEC for Windows is available in both NEC-2 and NEC-4 versions, although the NEC-4 version requires a license from the University of California to operate. It offers 2-D and 3-D plots, 3-D plots with 2-D slicing, ground-wave output, stepped-diameter correction, and various shortcuts to antenna modeling. The standard EZNEC is limited to 500 segments (the Pro version handles a much larger number of segments).

### **Expert MiniNEC**

This new version of MiniNEC can handle sharp angles in antenna geometry directly and handles antennas close to the ground much more accurately than MiniNEC. Several versions of Expert MiniNEC are available from EM Scientific. These are MiniNEC for Windows and MiniNEC Broadcast Professional.

### **MiniNEC**

MiniNEC Version 3.13 is the latest of the public-domain MiniNEC code. It runs on a PC in the DOS mode. It handles element lengths of changing diameter. Horizontal antennas must be at least  $0.2\lambda$  above ground for accurate results to occur.

### **NEC4Win95**

Orion (Canada) offers a Windows 95/98 version of MiniNEC known as NEC4Win95. It offers spreadsheet antenna geometry input, pull-down boxes for all other antenna parameters, and a pattern-plotting routine that includes lobe identification and bandwidth.

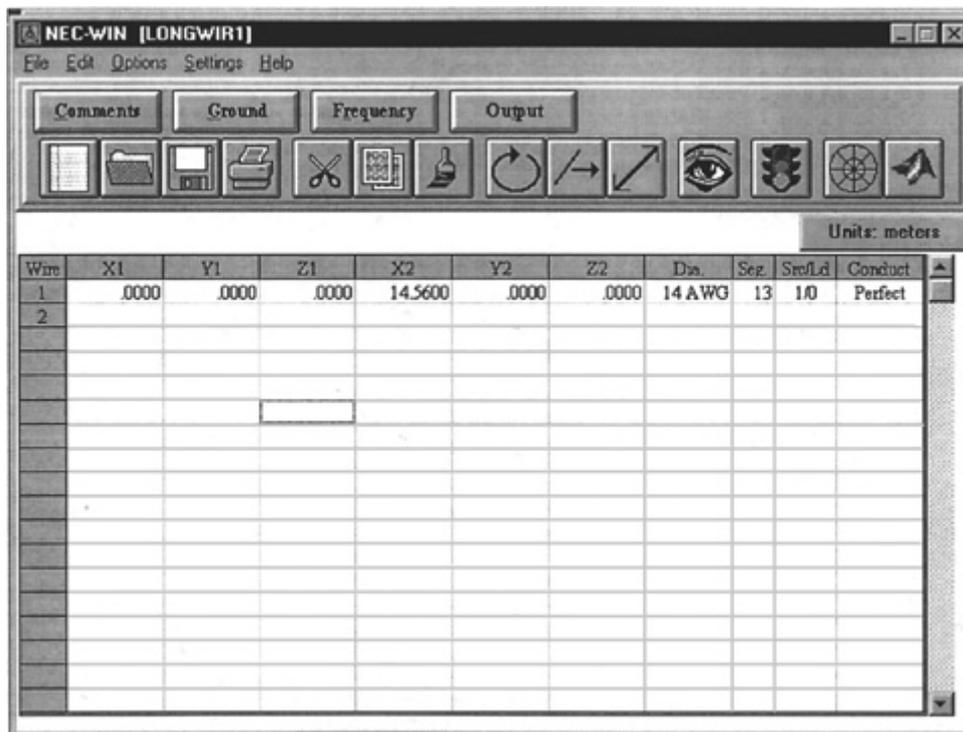
### **NECWin Basic**

Many amateur and professional users have been increasingly dissatisfied with the commercial miniNEC-based antenna modeling programs on the market (mostly because they only run in the DOS environment and are clumsy to use). A low-cost Windows-based modeling program is NecWin Basic for Windows. This program appears to be a DOS-based NEC engine run with a Windows graphic user interface written in Visual Basic 3.0. The vendor offering NecWin Basic for Windows is Nittany Scientific, Inc. [Airline Highway, Suite 361, Hollister, CA, 95023-5621; phone: (408) 634-0573; Web site: [www.nittany-scientific.com](http://www.nittany-scientific.com)]. The basic version (which I test drove and now use) costs less than \$100, and there is a professional version for a higher price.

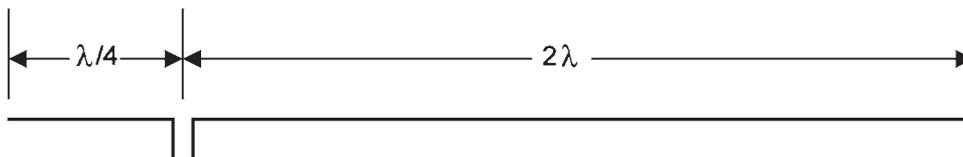
NecWin Basic for Windows allows you to specify the geometry of the antenna in the  $xyz$  coordinate system. This process is made a lot easier by the fact that it uses a "spreadsheet-like" grid (Fig. 17-3). Once the geometry is entered, you can select from several different ground situations (including "none" for free-space calculations), the operating frequency, and units of measure used for the antenna wires (meters is default). You also can customize the calculations using the "Output" button. Once the geometry and parameters are entered, you can visualize the geometry in "stick form" using the "Eye" button on the toolbar. The "Traffic Light" button

performs the NEC calculations, whereas the “Polar Coordinates” button creates the elevation and azimuth pattern graphics. When you “Save Screen” in the “File” menu, you will produce a bitmap (\*.bmp) file of the pattern.

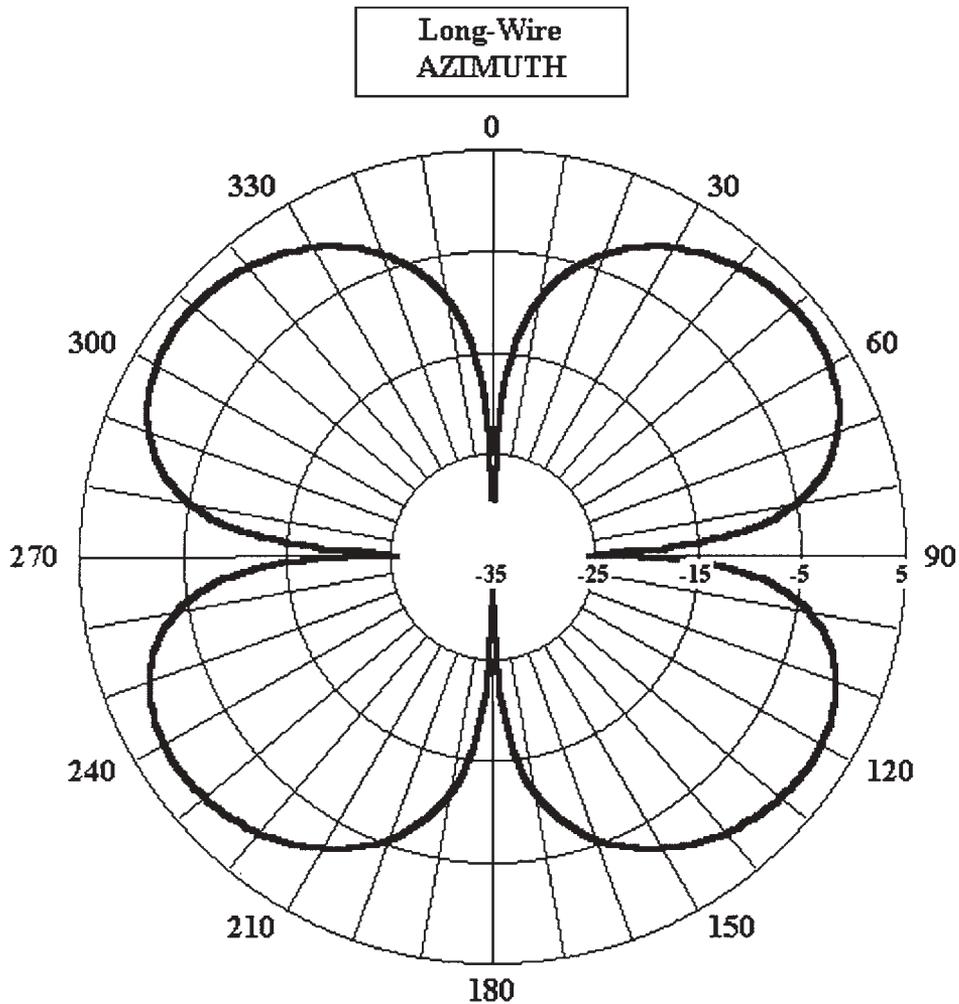
The particular geometry entered in Fig. 17-3 is for a long-wire antenna (Fig. 17-4) that is two wavelengths long and has a quarter-wavelength radial used as a counterpoise ground. There are no wires in the  $y$  and  $z$  planes, so the values of  $Y_1$ ,  $Y_2$ ,  $Z_1$ , and  $Z_2$  are zero. The antenna is described with the coordinates  $X_1 = 0.000$  and  $X_2 = 14.56$  m. The operating frequency is set to 25-MHz on another screen. When the NEC calculations are made and the patterns displayed, the azimuth (Fig. 17-5) and elevation (Fig. 17-6) patterns can be printed out.



17-3 NecWin screen.



17-4 Off-center fed doublet antenna.

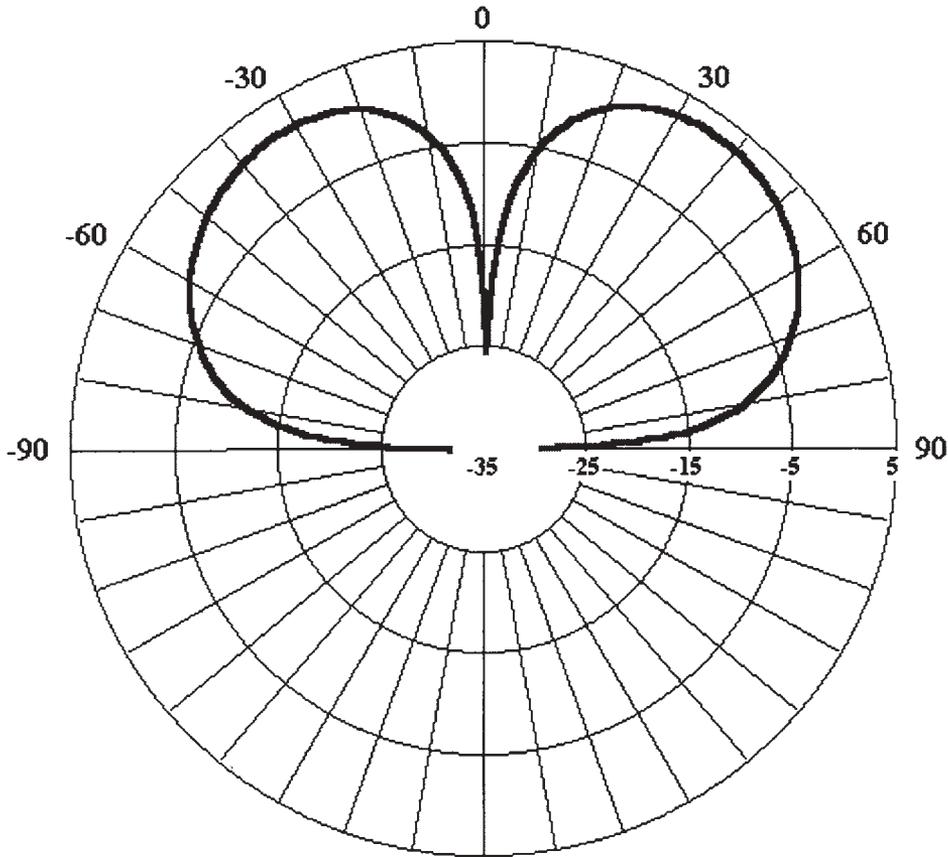


17-5 Modeled horizontal pattern.

NecWin Basic for Windows also will print out a data table of the frequency, wavelength, and VSWR. You can set the starting frequency, number of calculations, and step frequency.

If there is any criticism of the NecWin Basic for Windows program, it is that it does not graph the VSWR versus frequency curve. This is a little harder in Visual Basic 3.0 than in later versions, but users who also have Microsoft Excel on their computers can hand enter the numbers from the NecWin Basic for Windows output report and get a chart similar to Fig. 17-7.

The low price of NecWin Basic for Windows makes it useful to amateur radio operators, shortwave and scanner enthusiasts, and other amateur antenna builders. If



17-6 Modeled vertical pattern.

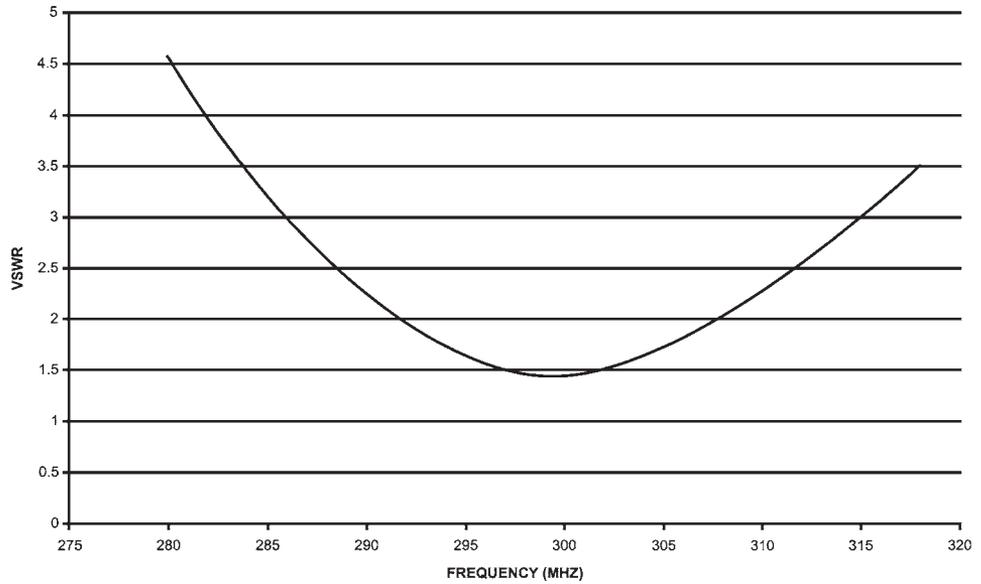
you are serious about antennas, then you need to obtain this software. It will perform well for many professional users, as well as student users in schools. However, professional users might want to consider buying the higher-priced NecWin Professional for Windows version.

### **EZNEC for Windows**

Roy Lewallen, W7EL, produces the EZNEC for Windows software. The software extends Roy's earlier work on DOS versions of EZNEC but is still based on the NEC-2 engine (a NEC-4 professional version is available for those with the required license from the University of California).

Figure 17-8 shows the basic window that opens when you open an antenna project (in this case, a 10-MHz dipole). Given in the active window are file name, frequency, wavelength (in meters), number of wires and segments, number of sources, number of loads, transmission lines, ground type, wire loss, the units being calculated

### VSWR Dipole



17-7 VSWR curve of the modeled antenna.



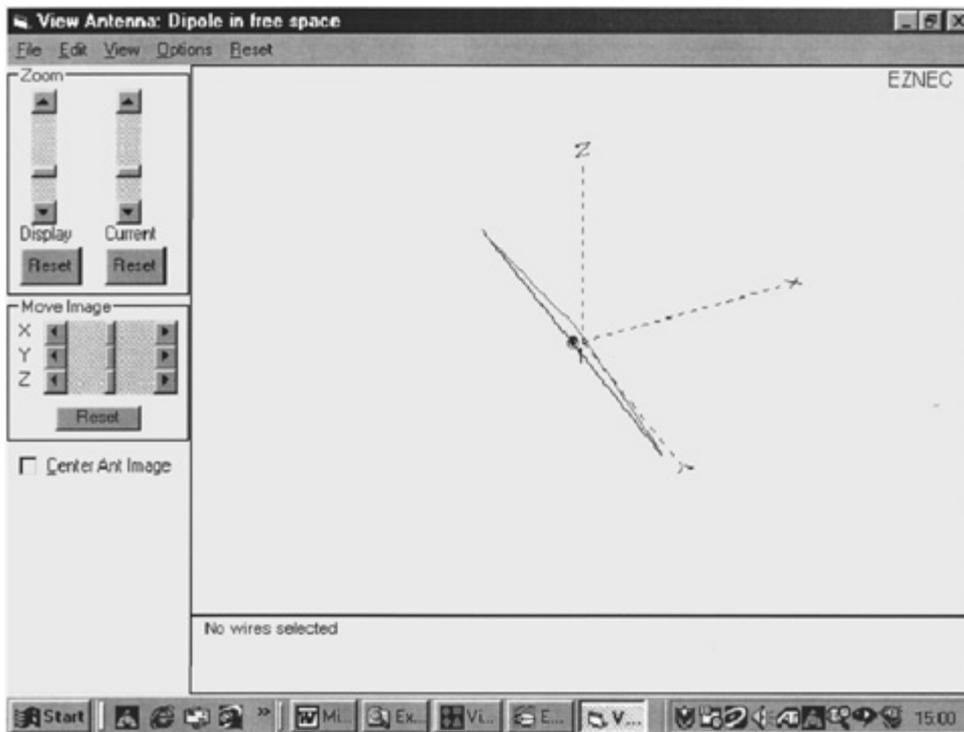
17-8 Basic window of EZNEC for Windows.

(meters, feet, etc.), plot type (azimuth, elevation), elevation angle, step size, a reference level, and an alternate SWR  $Z_o$ .

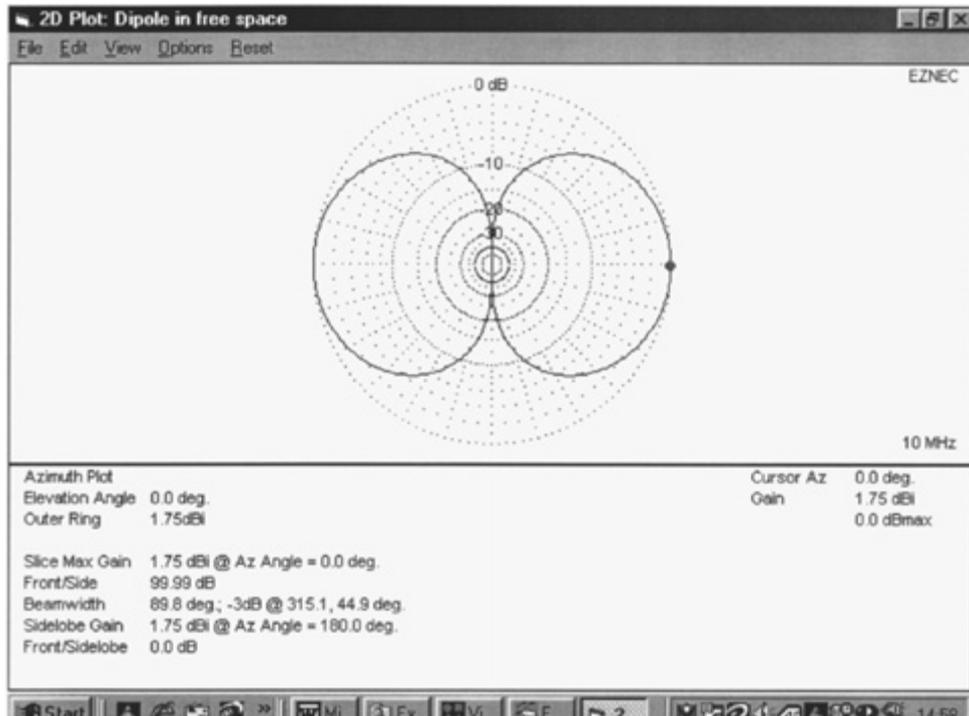
There are command buttons to the left of the main window. These include Open, Save As, Currents, Sto Dat, Load Dat, FF Tab, NF Tab, SWR, View Antenna, and FF Plot. The “View Antenna” button provides a view of the  $xyz$  coordinate system along with the wires (Fig. 17-9). The “FF Plot” button plots the far field of the antenna, and the “SWR” button plots the SWR curve based on parameters you give it (starting frequency, stopping frequency, frequency step).

The “View Antenna” window is shown in Fig. 17-9. This antenna view provides the  $xyz$  coordinate system, along with the wires. There are scroll bars to the left of the main window where the antenna appears that are used to vary the image to better see it. A series of three “Reset” buttons resets the image to the initial setting.

If you wish to plot the radiation pattern (azimuth or elevation) of the antenna, then press “FF Plot” in the main window. The result will be as shown in Fig. 17-10. Some data appears in the lower window on the plot. The data in this case are elevation angle, outer ring gain, slice gain maximum, front-to-side ratio, beamwidth, side-



17-9 View Antenna window of EZNEC for Windows.



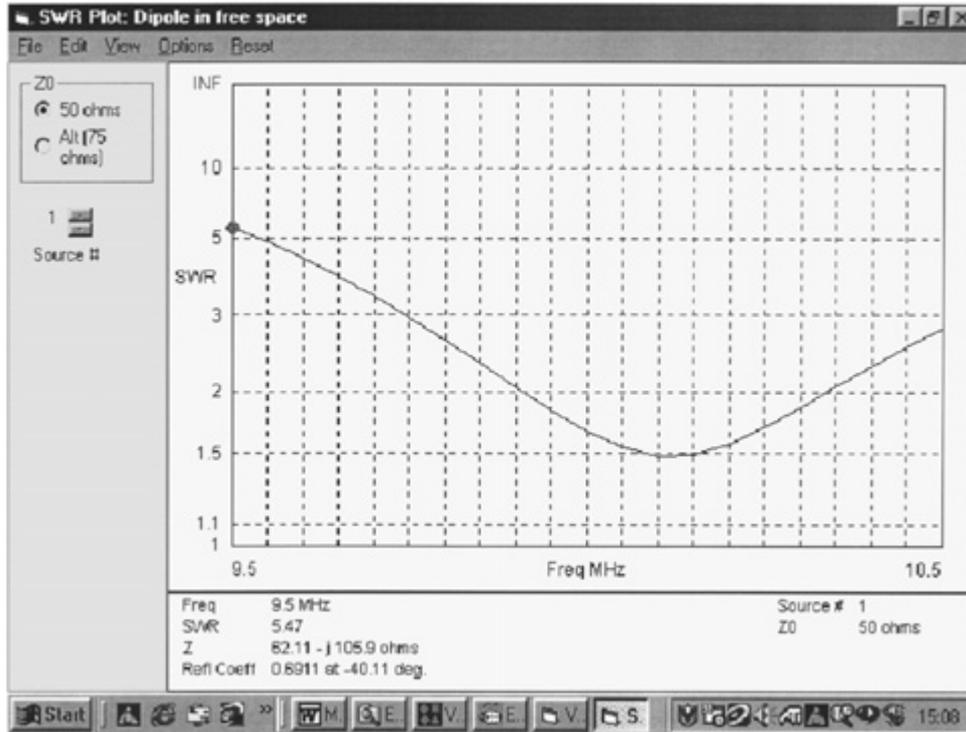
17-10 Result of pressing “FF Plot” button in main window of EZNEC for Windows.

lobe gain, front/sidelobe, cursor azimuth, and gain. You can print out the plot of the radiation pattern with either the data present or the data hidden.

One of the really nice things about EZNEC for Windows is the availability of VSWR plots without going to the professional edition. Figure 17-11 shows the plot for the dipole under consideration. Note that the data given include the starting frequency, the SWR at the starting frequency, the impedance at the starting frequency, the reflection coefficient at the starting frequency, and the source impedance.

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17-11 VSWR plot of a dipole with EZNEC for Windows.

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## 18 CHAPTER

# VHF/UHF transmitting and receiving antennas

THE VHF/UHF SPECTRUM IS COMMONLY ACCEPTED TO RANGE FROM 30 MHz TO 900 MHz, although the upper breakpoint is open to some differences of opinion. The VHF spectrum is 30 MHz to 300 MHz, and the UHF spectrum is 300 MHz to 900 MHz. Above 900 MHz is the microwave spectrum. These bands are used principally for local “line-of-sight” communications, according to the standard wisdom. However, with the advent of OSCAR satellites, the possibility of long-distance direct communications is a reality for VHF/UHF operators. In addition, packet radio is becoming common; this means indirect long-distance possibilities through networking. For the low end of the VHF spectrum (e.g., 6-m amateur band), long-distance communications are a relatively common occurrence.

In many respects, the low-VHF region is much like the 10-m amateur band and 11-m Citizens Band: skip is not an infrequent occurrence. Many years ago, I recall an event where such skip caused many a local police officer to skip a heart beat. In those days, our police department operated on 38.17 MHz, which is between the 6-m and 10-m amateur bands. They received an emergency broadcast concerning a bank robbery at a certain Wilson Boulevard address. After a race to the county line, they discovered that the reported address would be outside of the county, and in fact did not exist even in the neighboring county (a number was skipped). The problem was traced to a police department in a southwest city that also had a Wilson Boulevard, and for them the alarm was real.

The principal difference between the lower frequencies and the VHF/UHF spectrum is that the wavelengths are shorter in the VHF/UHF region. Consider the fact that the wavelengths for these bands range from 10 m to 1 m for the VHF region, and from 1 m to 33 cm for the UHF region. Most antenna designs are based on wavelength, so that fact has some implications for VHF/UHF antenna design. For example, because bandwidth is a function of length/diameter ratio for many classes of antenna, broadbanding an antenna in the VHF/UHF region is relatively easy. If, say, 25-mm (i.e., 1-in) aluminum tubing is used to make a quarter-wavelength vertical, then the approximate  $L/D$  ratio is 790 in the 8-m band and 20 in the 2-m band. This feature is fortunate, because the VHF/UHF bands tend to be wider than the HF bands.

Another point to make is that many of the mechanical chores of antenna design and construction become easier for VHF/UHF antennas. One good example is the delta impedance-matching scheme. At 80 m, the delta-match dimensions are approximately  $36 \times 43$  ft, and at 2 m they are  $9.5 \times 12$  inches. Clearly, delta matching is a bit more practical for most users at VHF than at HF.

## Types of antennas usable for VHF/UHF

The concept “VHF/UHF antenna” is only partially valid because virtually all forms of antenna can be used at HF, MW, and VHF/UHF. The main limitations that distinguish supposedly VHF/UHF designs from others are mechanical: there are some things that are simply much easier to accomplish with small antennas. Besides the delta match mentioned previously, there is the ease of construction for multielement antennas. A 14-element 20-m beam would be a wonderful thing to have in a QRM-laden DX pile-up, but is simply too impractical for all but a few users because of its size. If you look on embassy rooftops around the world you will see many-element Yagi and log periodic HF antennas supported on massive towers...and some of them use a standard Size 25 tower (common for amateur use) as the *antenna boom*! A 14-element 80/75-m Yagi approaches impossibility. But at 2 m, a 14-element Yagi beam antenna can be carried by one person, in one hand, unless the wind is acting up.

**Safety note** Large array beams, even at VHF/UHF, have a relatively high “windsail area,” and even relatively modest winds can apply a lot of force to them. I once witnessed a large, strong technician blown off a ladder by wind acting on a modest, “suburban” sized, TV antenna. *It can happen to you, too. So always install antennas with a helper, and use hoists and other tools to actually handle the array.*

## Lower band antennas on VHF/UHF

Between 1958 and 1962, a friend and I had access to a radio club amateur radio station in a Red Cross chapter house in Virginia. The “antenna farm” consisted of a 14-element 2-m beam, a three-element triband HF beam (10, 15, and 20 m), and a five-band (80 to 10 m) trap dipole. All of the coaxial cables came into the station through a wall; they were kept disconnected and shorted out when not in use because of the senior Red Cross official’s concern over lightning.

One night, attempting to connect the 2-m beam to the Gonset “gooney box” 2-m AM transceiver, my friend accidentally used the cable from the five-band trap dipole instead. We worked a lot of stations that contest weekend, and scored lots of points. Later, we discovered the error, and asked a more technically competent adult (we were teenagers), “Why the good reports?” He then gave us a lesson in longwire antenna theory. A good longwire is many wavelengths long. Consider that a half-wave antenna on 2 m is  $80 \text{ m} / 2 \text{ m}$  or 40 wavelengths shorter than an 80-m half-wave antenna. Thus, the 80-m antenna, counting foreshortening of physical lengths because of the traps, was on the order of 35 to 38 wavelengths long on 2 m. We had a highly directional, but multi-lobed, pattern.

Similarly, 40- to 10-m and 80- to 10-m trap verticals are often usable on VHF/UHF frequencies without any adjustments. Similarly, Citizens Band 11-m antennas, many of which are  $\frac{5}{8}$ -wavelength (18 ft high), will sometimes work on VHF frequencies. Check the VSWR of an HF antenna on 2 m with a reliable VHF/UHF VSWR meter (or RF wattmeter) to discover the truth about any particular antenna. Always use the low-power setting on the transmitter to limit damage in cases where the specific antenna is not usable on a specific frequency.

The lesson to be learned is that antennas are often usable on frequencies much higher than the design frequency, even though useless on nearby bands. Care must be exercised when initially checking out the antenna, but that is not an inordinate difficulty.

## VHF/UHF antenna impedance matching

The VHF/UHF antenna is no more or less immune from the need for impedance matching than lower-frequency antennas. However, some methods are easier (coax baluns, delta match, etc.) and others become either difficult or impossible. An example of the latter case is the tuned  $LC$  impedance-matching network. At 6 m, and even to some limited extent 2 m, inductor and capacitor  $LC$  networks can be used. But above 2 m other methods are more reasonable. We can, however, mimic the  $LC$  tuner by using stripline components, but that approach is not always suited to amateur needs.

The balun transformer makes an impedance transformation between *balanced* and *unbalanced* impedances. Although both 1:1 and 4:1 impedance ratios are possible, the 4:1 ratio is most commonly used for VHF/UHF antenna work. At lower frequencies it is easy to build broadband transformer baluns, but these become more of a problem at VHF and above.

For the VHF/UHF frequencies, a 4:1 impedance ratio coaxial balun (Fig. 18-1A) is normally used. Two sections of identical coaxial cable are needed. One section (*A*) has a convenient length to reach between the antenna and the transmitter. Its characteristic impedance is  $Z_o$ . The other section (*B*) is a half-wavelength long at the center of the frequency range of interest. The physical length is found from

$$L = \frac{5904 V}{F_{\text{MHz}}} \quad [18.1]$$

where

$L$  is the cable length, in inches

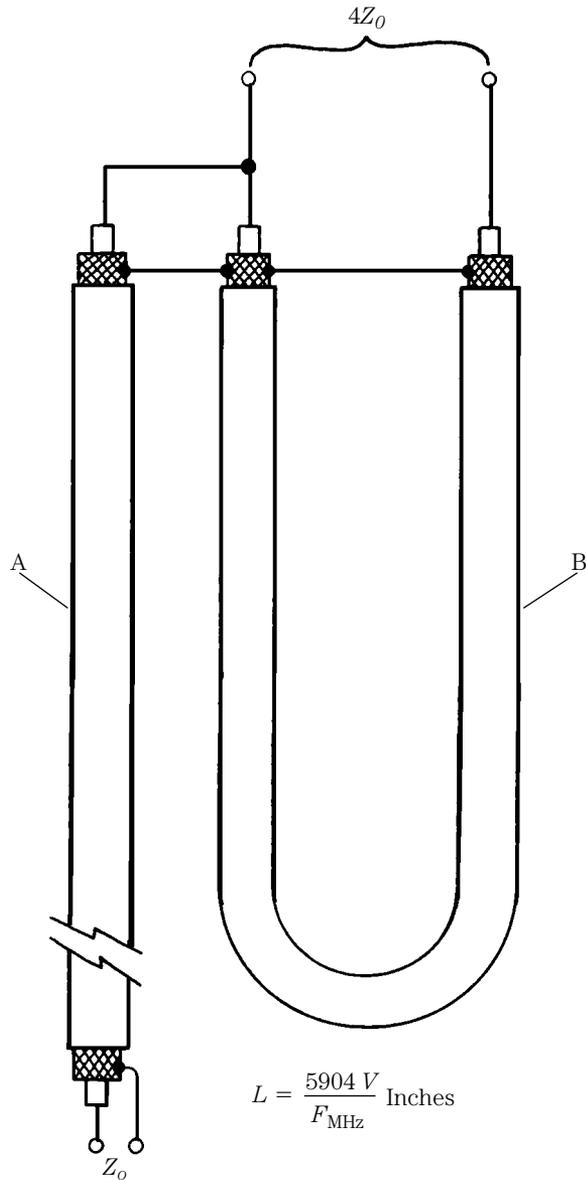
$F_{\text{MHz}}$  is the operating frequency, in megahertz

$V$  is the velocity factor of the coaxial cable

The velocity factors of common coaxial cables are shown in the following table.

### Coaxial cable velocity factors

Regular polyethylene	0.66
Polyethylene foam	0.80
Teflon	0.72



**18-1A** Coaxial 4:1 balun transformer.

**Example 18-1** Calculate the physical length required of a 146-MHz 4:1 balun made of polyethylene foam coaxial cable.

**Solution:**

$$L = \frac{5904 V}{F_{\text{MHz}}}$$

$$\begin{aligned}
 &= \frac{(5904)(0.80)}{146 \text{ MHz}} \\
 &= \frac{4723.2}{146 \text{ MHz}} = 32.4 \text{ in}
 \end{aligned}$$

A mechanical method of joining the coaxial cables is shown in Fig. 18-1B. In this example, three SO-239 coaxial receptacles are mounted on a metal plate. This arrangement has the effect of shorting together the shields of the three ends of coaxial cable. The center conductors are connected in the manner shown. This method is used especially where a mounting bracket is available on the antenna. The lengths of coaxial cable need PL-259 coax connectors installed in order to use this method.

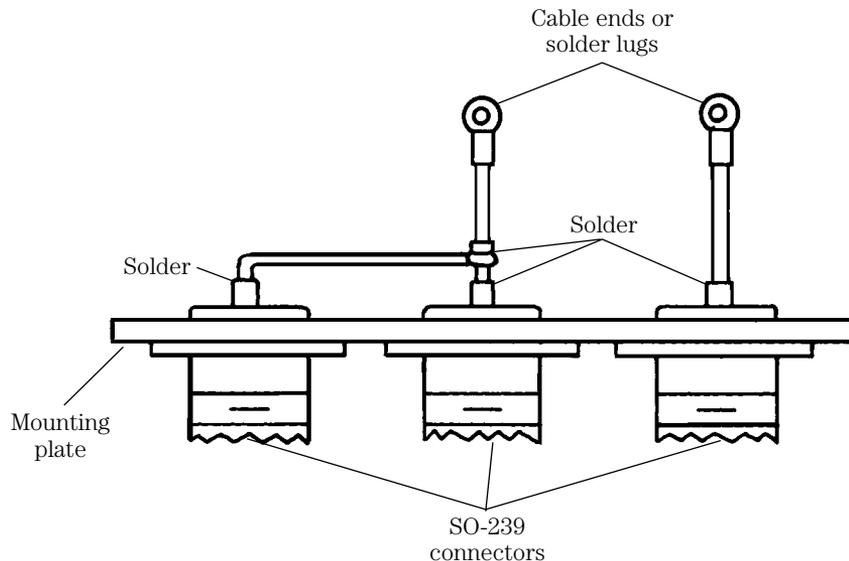
The “delta match” gets its name from the fact that the structure of the matching element has the shape of the Greek letter delta, or a triangle. Figure 18-2A shows the basic delta match scheme. The matching element is attached to the driven element of the antenna (symmetrically, about the center point of the antenna). The width ( $A$ ) of the delta match is given by

$$L = \frac{1416}{F_{\text{MHz}}} \text{ in} \quad [18.2]$$

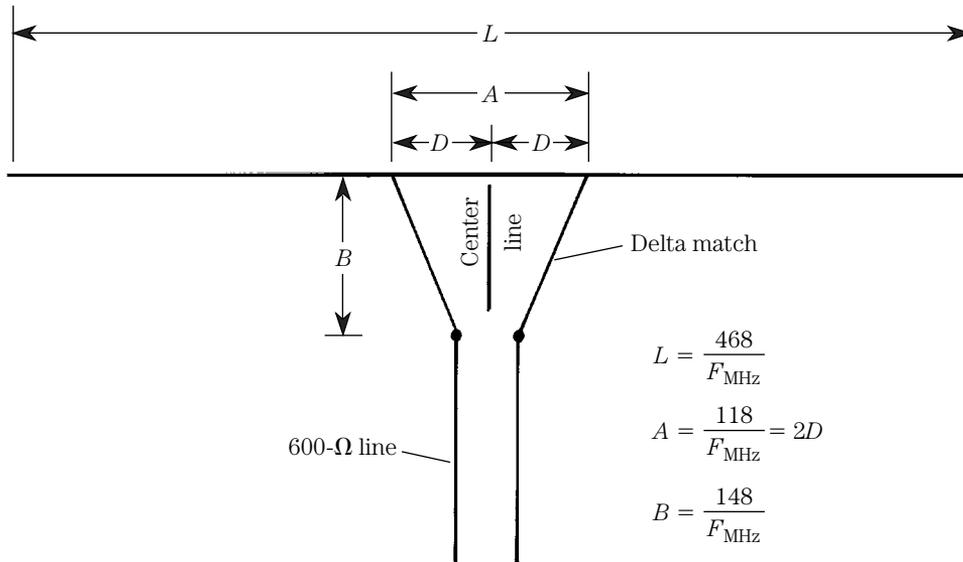
While the height of the match ( $B$ ) is

$$L = \frac{1776}{F_{\text{MHz}}} \text{ in} \quad [18.3]$$

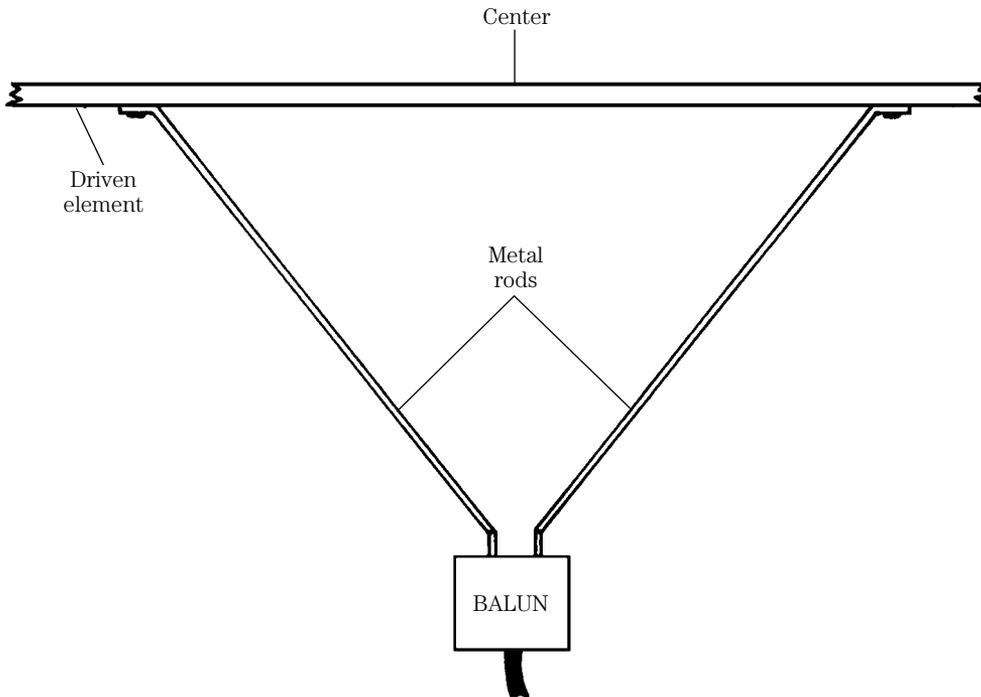
The transmission line feeding the delta match is balanced line, such as parallel transmission line or twin lead. The exact impedance is not terribly critical because the dimensions (especially  $A$ ) can be adjusted to accommodate differences. In general, however, either 450- or 600- $\Omega$  line is used, although 300- $\Omega$  line can also be used. Figure 18-2B



**18-1B** Practical implementation of 4:1 balun using connectors.



**18-2A** Delta feed matching system.



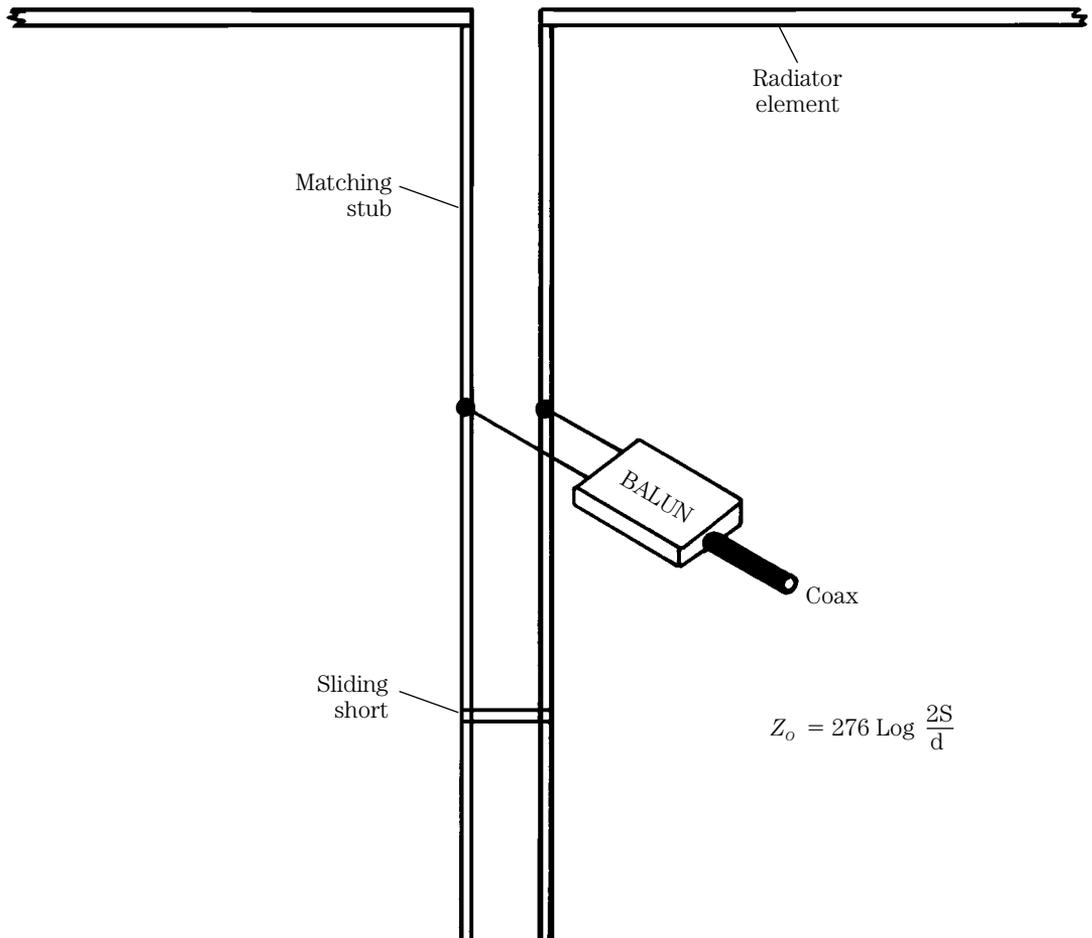
**18-2B** Practical VHF delta match.

shows a method for using coaxial cable with the delta match. The impedance is transformed in a 4:1 balun transformer (see Fig. 18-1A). The elements of the delta match can be made from brass, copper, or aluminum tubing, or a bronze brazing rod bolted to the main radiator element.

A stub-matching system is shown in Fig. 18-3. In this case, the impedance transformation is accomplished despite the half-wavelength shorted stub of transmission line. The exact impedance of the line is not very critical, and is found from

$$Z_o = 276 \log_{10} \frac{2S}{d} \quad [18.4]$$

The matching stub section is made from metal elements such as tubing, wire, or rods (all three are practical at VHF/UHF frequencies). For a  $\frac{3}{16}$ -in rod, the spacing is approximately 2.56 in to make a 450- $\Omega$  transmission line. A sliding short circuit is used to set the electrical length of the half-wave stub. The stub is tapped at a distance



18-3 Stub matching.

from the antenna feedpoint that matches the impedance of the transmission line. In the example shown, the transmission line is coaxial cable, so a 4:1 balun transformer is used between the stub and the transmission line. The two adjustments to make in this system are: (1) the distance of the short from the feedpoint and (2) the distance of the transmission line tap point from the feedpoint. Both are adjusted for minimum VSWR.

The gamma match is basically a half delta match, and operates according to similar principles (Fig. 18-4). The shield (outer conductor) of the coaxial cable is connected to the center point of the radiator element. The center conductor of the coaxial cable series feeds the gamma element through a variable capacitor.

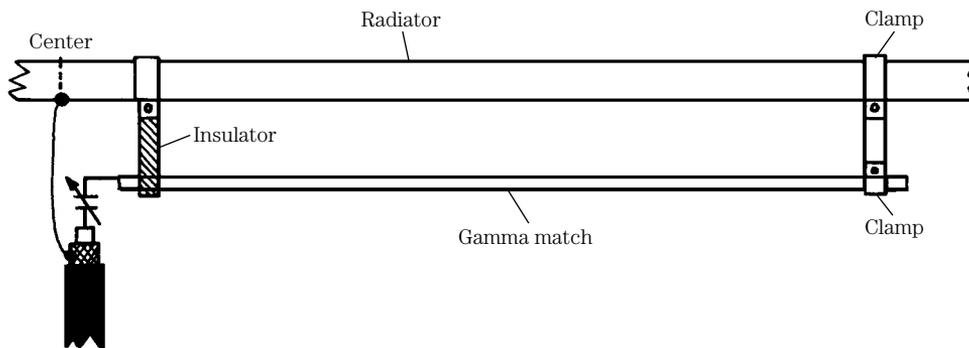
## VHF/UHF antenna examples

Although it is probably not necessary to reiterate the point, VHF/UHF antennas are not substantially different from HF antennas, especially those for the higher bands. However, for various practical reasons there are several forms that are specially suited, or at least popular, in the VHF/UHF bands. In this section we will take a look at some of them.

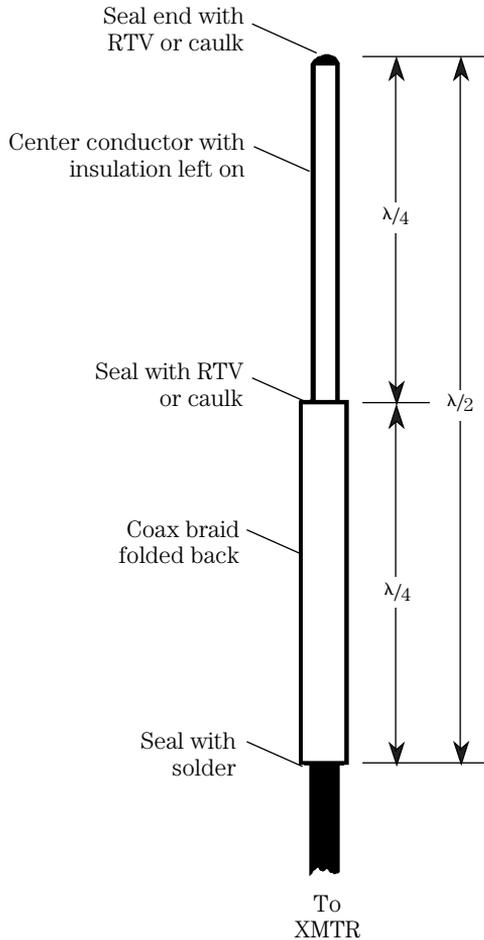
### Coaxial vertical

The coaxial vertical is a quarter-wavelength vertically polarized antenna that is popular on VHF/UHF. There are two varieties. In Fig. 18-5A we see the coaxial antenna made with coaxial cable. Although not terribly practical for long-term installation, the coax-coax antenna is very useful for short-term, portable, or emergency applications. For example, a boater found himself adrift, and in dire trouble, after a storm damaged the boat. The mast-top VHF antenna was washed away, leaving only the end of the coaxial cable dangling loose. Fortunately, the boat operator was a two-way radio technician, and he knew how to strip back the coaxial cable to make an impromptu coaxial vertical.

The coax-coax antenna shown in Fig. 18-5A uses a quarter-wavelength radiator and a quarter-wavelength sleeve. The sleeve consists of the coax braid stripped back and folded down the length of the coax cable. The maximum length is found from the equation below (actual length is trimmed from this maximum):



18-4 Gamma matching.

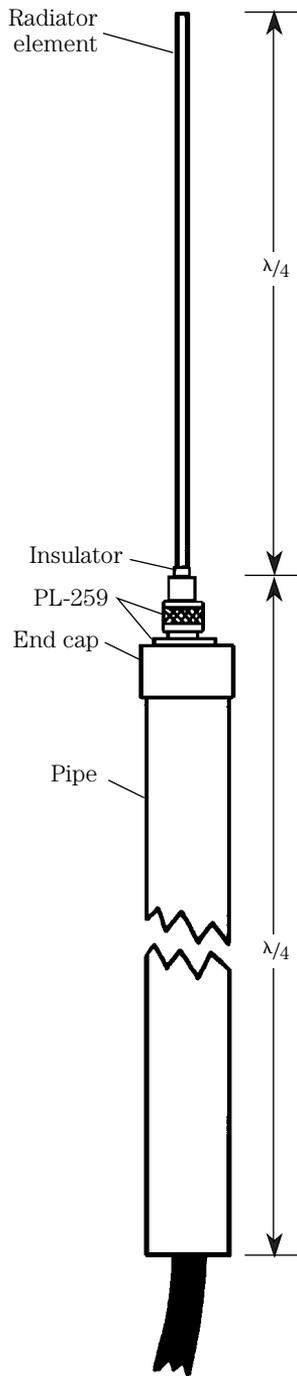


**18-5A** Coaxial vertical based on coaxial cable.

$$L = \frac{2952 V}{F_{\text{MHz}}} \text{ in} \quad [18.5]$$

The antenna is mounted by suspending it from above by a short piece of string, twine, or fishing line. From a practical point of view, the only problem with this form of antenna is that it tends to deteriorate after a few rainstorms. This effect can be reduced by sealing the end, and the break between the sleeve and the radiator, with either silicone *RTV* or bathtub caulk.

A more permanent method of construction is shown in Fig. 18-5B. The sleeve is a piece of copper or brass tubing (pipe) about 1 in in diameter. An end cap is fitted over the end and sweat-soldered into place. The solder is not intended to add mechanical strength, but rather to prevent weathering from destroying the electrical contact between the two pieces. An SO-239 coaxial connector is mounted on the end cap. The coax is connected to the SO-239 inside the pipe, which means making the connection before mounting the end cap.



**18-5B** Tubing coaxial vertical.

The radiator element is a small piece of tubing (or brazing rod) soldered to the center conductor of a PL-259 coaxial connector. An insulator is used to prevent the rod from shorting to the outer shell of the PL-259. (Note: an insulator salvaged from the smaller variety of banana plug can be shaved a small amount with a fine file and made to fit inside the PL-259. It allows enough center clearance for  $\frac{1}{8}$ -inch or  $\frac{3}{16}$ -inch brass tubing.)

Alternatively, the radiator element can be soldered to a banana plug. The normal-size banana plug happens to fit into the female center conductor of the SO-239.

### Collinear vertical

*Gain* in antennas is provided by *directivity*. In other words, by taking the power radiated by the antenna, and projecting it into a limited direction, we obtain the appearance of higher radiated power. In fact, the *effective radiated power* (ERP) of the antenna is merely its feedpoint power multiplied by its gain. Although most antenna patterns are shown in the horizontal dimension (as viewed from above), it is also possible to obtain gain by compressing the vertical aspect. In this manner it is possible to have a vertical antenna that produces gain. Figure 18-6 shows a collinear gain antenna, with vertical polarization and a horizontally omnidirectional pattern. Incidentally, when mounted horizontally the pattern becomes bidirectional.

The collinear antenna shown in Fig. 18-6 is basically a pair of stacked collinear arrays. Each array consists of a quarter-wavelength section *A* and a half-wavelength section *C* separated by a quarter-wavelength phase reversing stub *B*. The phase reversing stub preserves in-phase excitation for the outer element (referenced to the inner element).

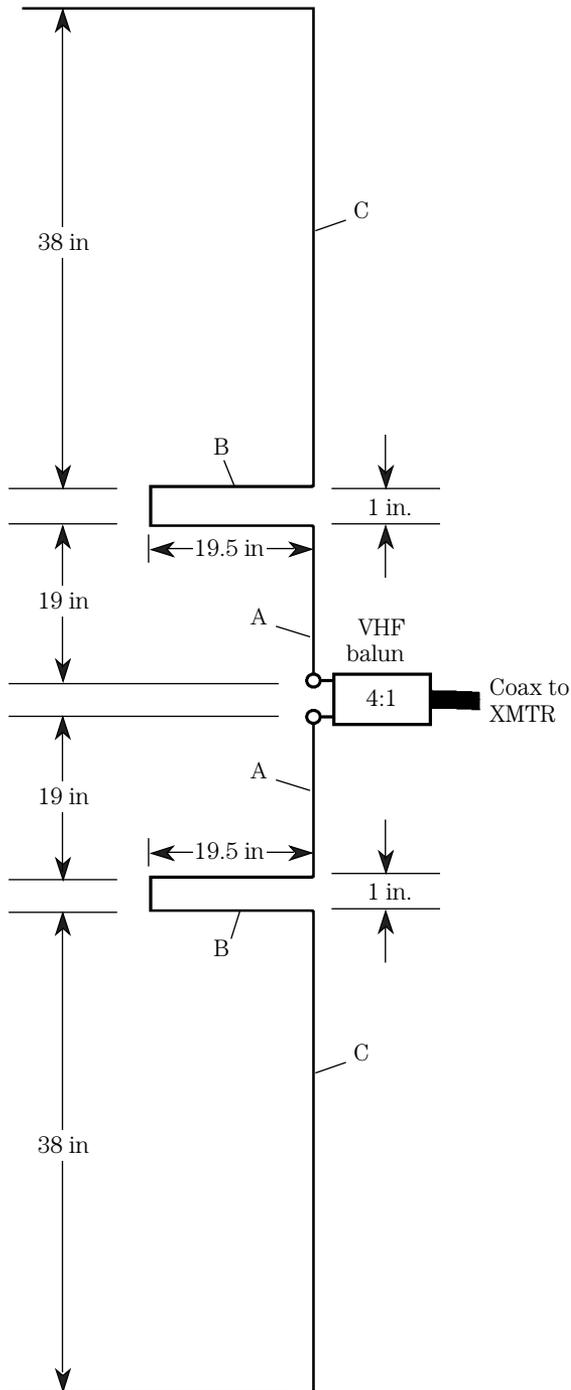
The feedpoint is between the two elements of the array (i.e., between the *A* sections). The coaxial-cable impedance is transformed by a 4:1 balun transformer (see Fig. 18-1A). Alternatively, 300- $\Omega$  twin lead can be used for the transmission line. If this alternative is used, then the use of UHF shielded twin lead is highly recommended. If the transmitter lacks the balanced output needed to feed twin lead, then use a balun at the input end of the twin lead (i.e., right at the transmitter).

### Yagi antennas

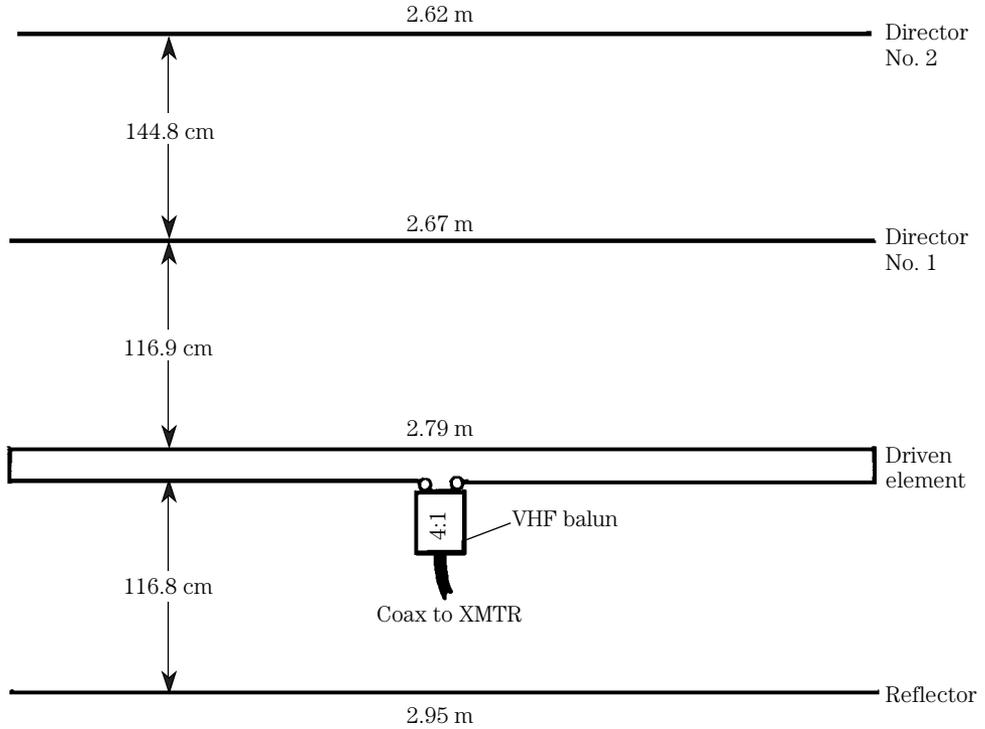
The Yagi beam antenna is a highly directional gain antenna, and is used both in HF and VHF/UHF systems. The antenna is relatively easy to build at VHF/UHF. In fact, it is easier than for HF systems. The basic Yagi was covered in Chap. 12, so we will only show examples of practical VHF devices. A 6-m Yagi antenna is shown in Fig. 18-7. This particular antenna is a four-element model. The reflector and directors can be mounted directly to a metallic boom, because they are merely parasitic. The driven element, however, must be insulated from the metal boom.

The driven element shown in Fig. 18-7 is a folded dipole. While this is common practice at VHF, because it tends to broadband the antenna, it is not strictly necessary. The dimensions of the driven element are found from Eq. 18.4. Set the equation equal to 300  $\Omega$ , select the diameter of the tubing from commercially available sources, and then calculate the spacing.

**Example 18-2** Calculate the spacing of a 300- $\Omega$  folded dipole when  $\frac{3}{4}$ -in tubing is used in its construction.



**18-6** Vertical collinear antenna.



**18-7** Six-meter beam antenna.

Solution:

$$Z_o = 276 \log_{10} \frac{2S}{d}$$

$$300 \Omega = 276 \log_{10} \frac{2S}{d}$$

$$\frac{300}{276} = \log_{10} \frac{2S}{d}$$

$$1.08 = \log_{10} \frac{2S}{d}$$

$$10^{1.08} = \frac{2S}{d}$$

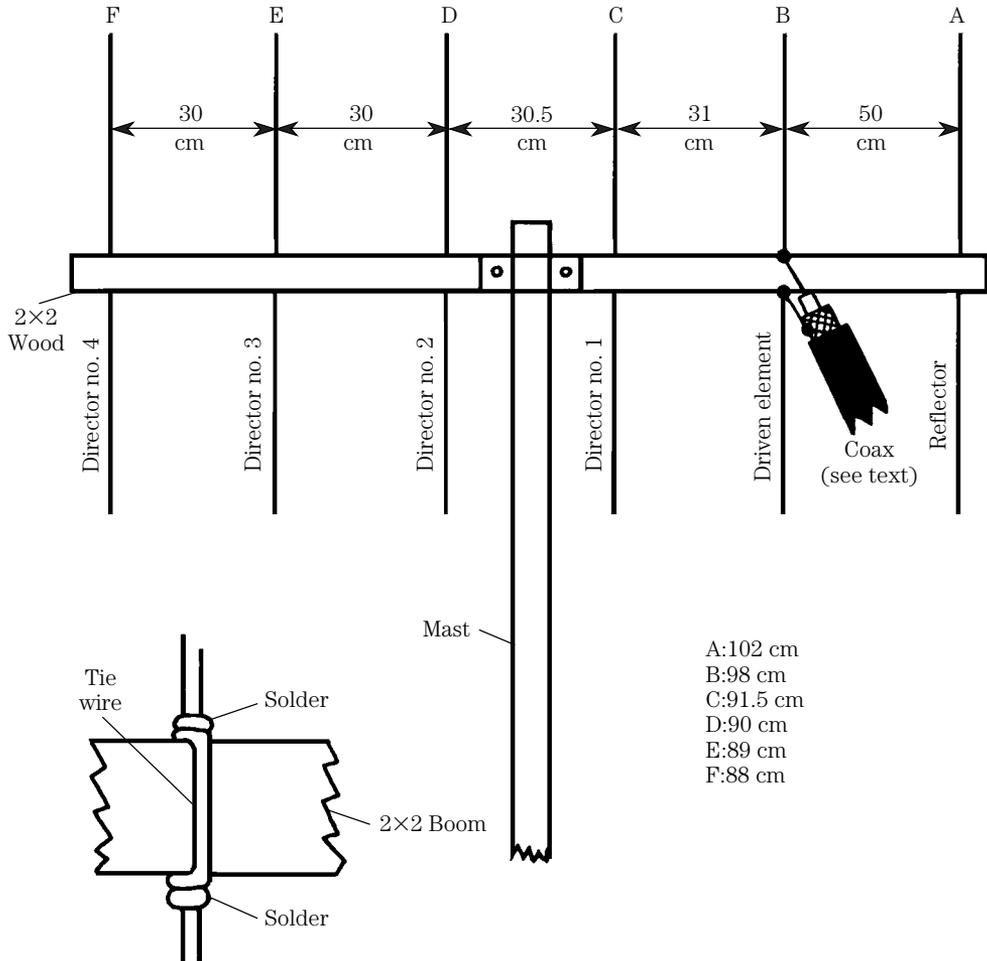
Using 0.75-in pipe results in

$$10^{1.08} = \frac{2S}{0.75}$$

$$12.02 = 2.67S$$

so

$$S = 4.5 \text{ in}$$



18-8 Two-meter vertical beam.

### Two-meter yagi

Figure 18-8 shows the construction details for a six-element 2-meter Yagi beam antenna. This antenna is built using a 2 × 2-in wooden boom and elements made of either brass or copper rod. Threaded brass rod is particularly useful, but not strictly necessary. The job of securing the elements (other than the driven element) is easier when threaded rod is used, because it allows a pair of hex nuts, one on either side of the 2 × 2-in boom, to be used to secure the element. Nonthreaded elements can be secured with RTV sealing a press-fit. Alternatively, tie wires (see inset to Fig. 18-8) can be used to secure the rods. A hole is drilled through the 2 × 2 to admit the rod or tubing. The element is secured by wrapping a tie wire around the rod on either side of the 2 × 2, and then soldering it in place. The tie wire is no. 14 to no. 10 solid wire.

Mounting of the antenna is accomplished by using a mast secured to the boom with an appropriate clamp. One alternative is to use an end-flange clamp, such as

is sometimes used to support pole lamps, etc. The mast should be attached to the boom at the center of gravity, which is also known as the *balance point*. If you try to balance the antenna in one hand unsupported, there is one (and only one) point at which it is balanced (and won't fall). Attach the mast hardware at, or near, this point in order to prevent normal gravitational torques from tearing the mounting apart.

The antenna is fed with coaxial cable at the center of the driven element. Ordinarily, either a matching section of coax, or a gamma match, will be needed because the effect of parasitic elements on the driven element feedpoint impedance is to reduce it.

### $\frac{5}{8}$ -wavelength 2-m antenna

The  $\frac{5}{8}$ -wavelength antenna (Fig. 18-9) is popular on 2 m for mobile operation because it is easy to construct, and it provides a small amount of gain relative to a dipole. The radiator element is  $\frac{5}{8}$ -wavelength, so its physical length is found from:

$$L = \frac{7380}{F_{\text{MHz}}} \text{ in} \quad [18.6]$$

The  $\frac{5}{8}$ -wavelength antenna is not a good match to any of the common forms of coaxial cables. Either a matching section of cable, or an inductor match, is normally used. In Fig. 18-9 an inductor match is used. The matching coil consists of 2 to 3 turns of no. 12 wire, wound over a  $\frac{1}{2}$ -in OD form,  $\frac{1}{2}$ -in long. The radiator element can be tubing, brazing rod, or a length of heavy "piano wire." Alternatively, for low-power systems, it can be a telescoping antenna that is bought as a replacement for portable radios or televisions. These antennas have the advantage of being adjustable to resonance without the need for cutting.

### J-pole antennas

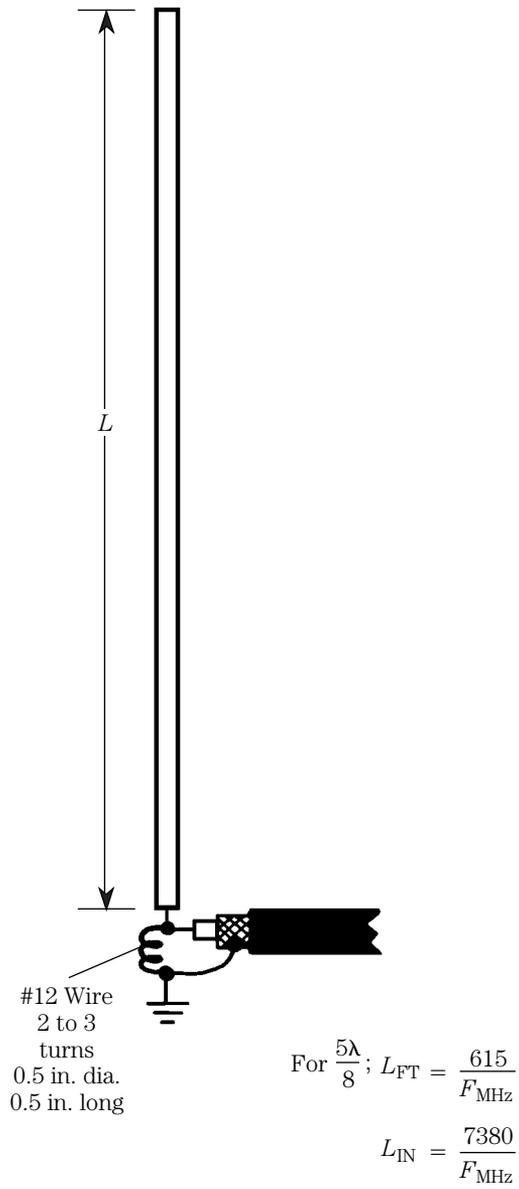
The *J-pole antenna* is another popular form of vertical on the VHF bands. It can be used at almost any frequency, although the example shown in Fig. 18-10 is for 2 m. The antenna radiator is  $\frac{3}{4}$ -wavelength long, so its dimension is found from

$$L = \frac{8838}{F_{\text{MHz}}} \text{ in} \quad [18.7]$$

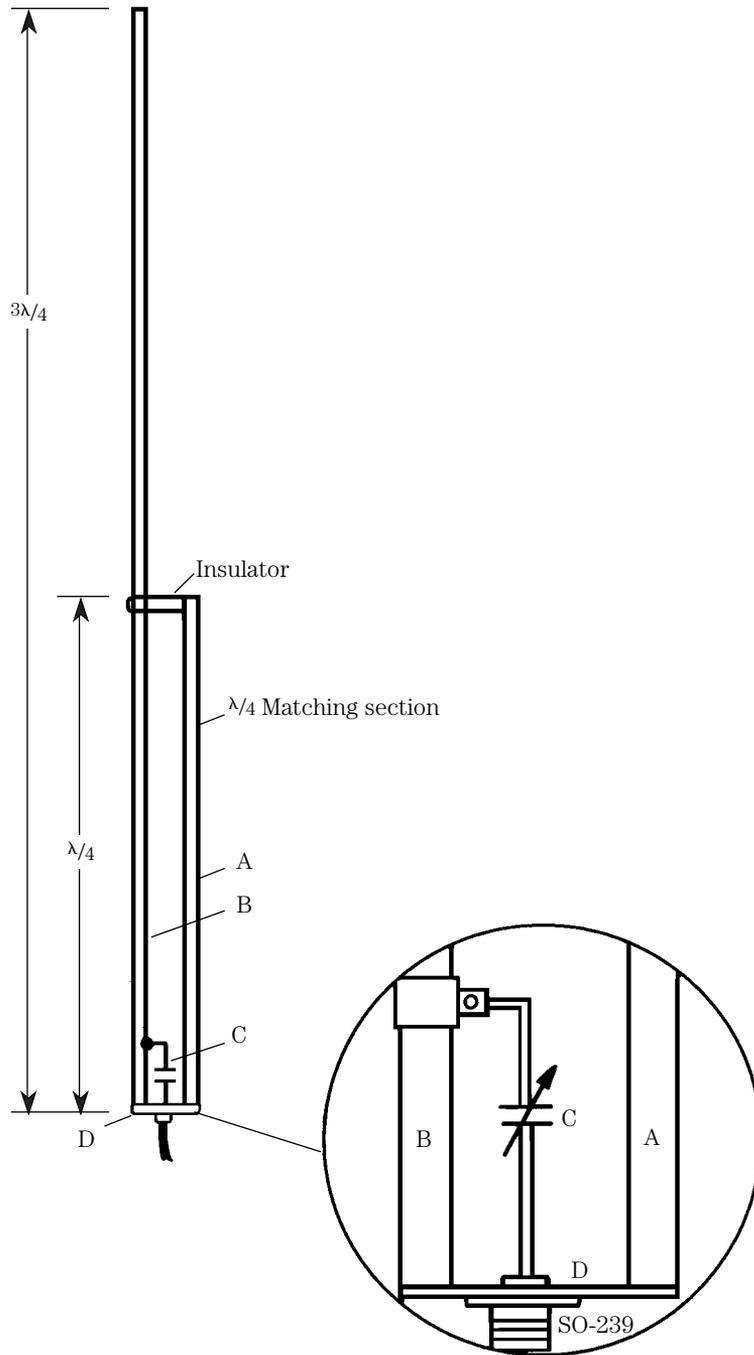
and the quarter-wavelength matching section length from

$$L = \frac{2952}{F_{\text{MHz}}} \text{ in} \quad [18.8]$$

Taken together the matching section and the radiator form a parallel transmission line with a characteristic impedance that is 4 times the coaxial cable impedance. If 50- $\Omega$  coax is used, and the elements are made from 0.5 in OD pipe, then a spacing of 1.5 in will yield an impedance of about 200  $\Omega$ . Impedance matching is accomplished by a gamma match consisting of a 25-pF variable capacitor, connected by a clamp to the radiator, about 6 in (experiment with placement) above the base.



18-9  $\frac{5}{8}$ -wavelength 2-m antenna.



18-10 J-pole antenna.

## Groundplane

The *groundplane antenna* is a vertical radiator situated above an artificial RF ground consisting of quarter-wavelength radiators. Groundplane antennas can be either  $\frac{1}{4}$ -wavelength or  $\frac{5}{8}$ -wavelength (although for the latter case impedance matching is needed—see the previous example).

Figure 18-11 shows how to construct an extremely simple groundplane antenna for 2 m and above. The construction is too lightweight for 6-m antennas (in general), because the element lengths on 6-m antennas are long enough to make their weight too great for this type of construction. The base of the antenna is a single SO-239 chassis-type coaxial connector. Be sure to use the type that requires four small machine screws to hold it to the chassis, and not the single-nut variety.

The radiator element is a piece of  $\frac{3}{16}$ -in or 4-mm brass tubing. This tubing can be bought at hobby stores that sell airplanes and other models. The sizes quoted just happen to fit over the center pin of an SO-239 with only a slight tap from a lightweight hammer—and I do mean *slight* tap. If the inside of the tubing and the connector pin are pretinned with solder, then sweat soldering the joint will make a good electrical connection that is resistant to weathering. Cover the joint with clear lacquer spray for added protection.

The radials are also made of tubing. Alternatively, rods can also be used for this purpose. At least four radials are needed for a proper antenna (only one is shown in Fig. 18-11). This number is optimum because they are attached to the SO-239 mounting holes, and there are only four holes. Flatten one end of the radial, and drill a small hole in the center of the flattened area. Mount the radial to the SO-239 using small hardware (4-40, etc.).

The SO-239 can be attached to a metal L bracket. While it is easy to fabricate such a bracket, it is also possible to buy suitable brackets in any well-equipped hardware store. While shopping at one do-it-yourself type of store, I found several reasonable candidate brackets. The bracket is attached to a length of  $2 \times 2$ -in lumber that serves as the mast.

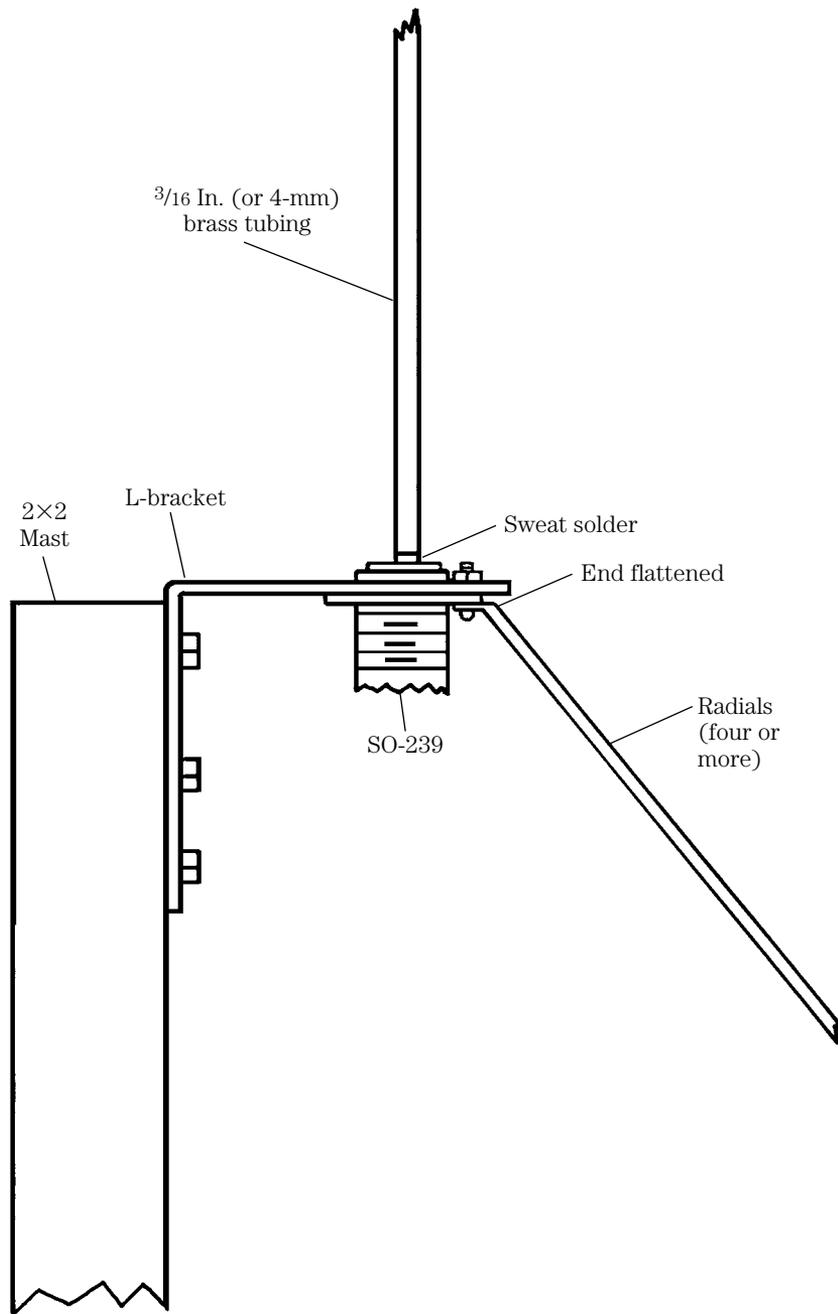
## Halo antennas

One of the more saintly antennas used on the VHF boards is the halo (Fig. 18-12). This antenna basically takes a half-wavelength dipole and bends it into a circle. The ends of the dipole are separated by a capacitor. In some cases, a transmitting-type mica “button” capacitor is used, but in others (and perhaps more commonly), the halo capacitor consists of two 3-in disks separated by a plastic dielectric. While air also serves as a good (and perhaps better) dielectric, the use of plastic allows mechanical rigidity to the system.

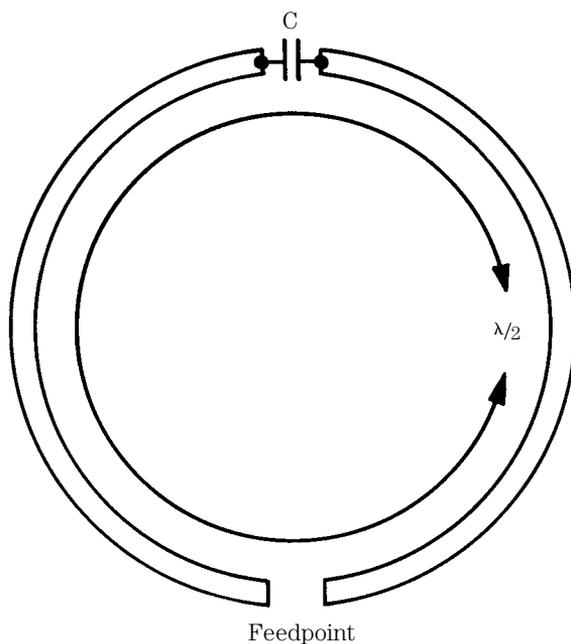
## Quad beam antennas

The quad antenna was introduced in the chapter on beams. It is, nonetheless, also emerging as a very good VHF/UHF antenna. It should go without saying that the antenna is a lot easier to construct at VHF/UHF frequencies than it is at HF frequencies! Figure 18-13 shows a modest example.

There are several methods for building the quad antenna, and Fig. 18-13 represents only one of them. The radiator element can be any of several materials,



18-11 Small VHF groundplane construction.



**18-12** Halo antenna.

including heavy solid wire (no. 8 to no. 12), tubing, or metal rods. The overall lengths of the elements are given by:

Driven element:

$$L = \frac{11,826}{F_{\text{MHz}}} \text{ in} \quad [18.9]$$

Reflector:

$$L = \frac{12,562}{F_{\text{MHz}}} \text{ in} \quad [18.10]$$

Director:

$$L = \frac{11,248}{F_{\text{MHz}}} \text{ in} \quad [18.11]$$

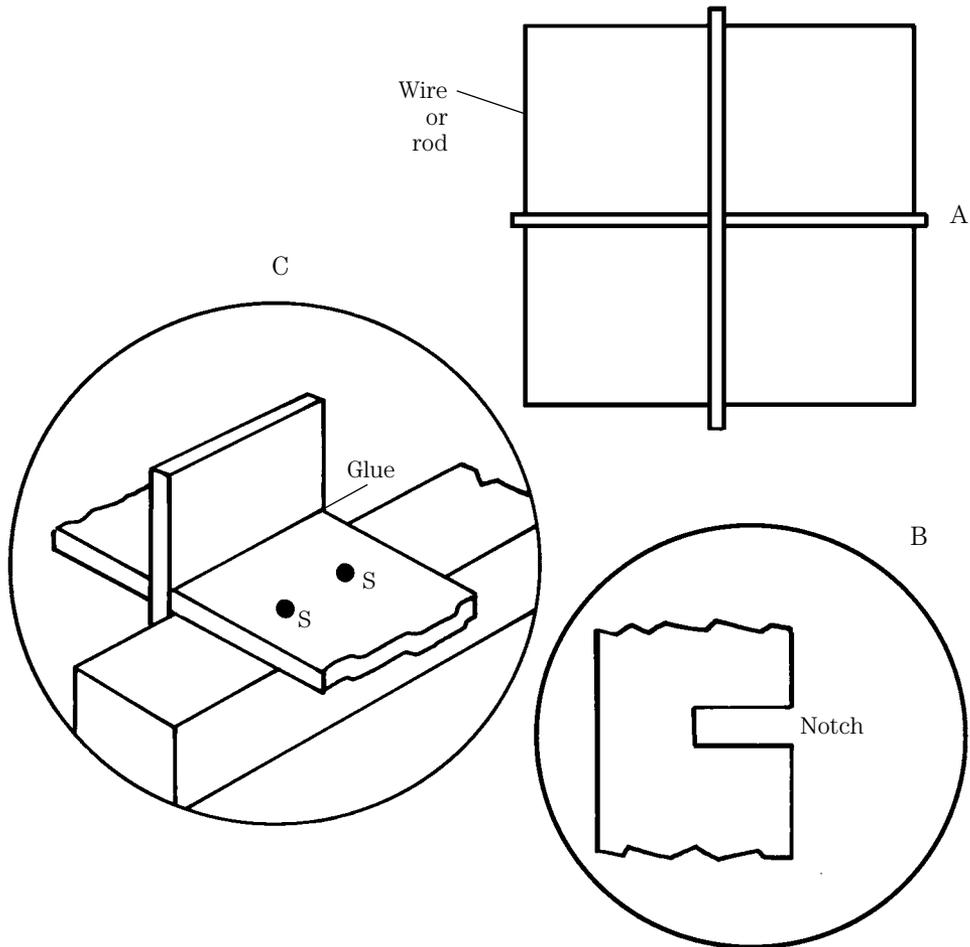
There are several alternatives for making the supports for the radiator. Because of the lightweight construction, almost any method can be adapted for this purpose. In the case shown in Fig. 18-13, the spreaders are made from either 1-in furring strips, trim strips, or (at above 2 m) even wooden paint stirring sticks. The sticks are cut to length, and then half-notched in the center (Fig. 18-13, detail B). The two spreaders for each element are joined together at right angles and glued (Fig. 18-13, detail C). The spreaders can be fastened to the wooden boom at points *S* in detail C. The usual rules regarding element spacing (0.15 to 0.31 wavelength) are followed. See the information on quad antennas in Chap. 12 for further details. Quads have been successfully built for all amateur bands up to 1296 MHz.

Overall lengths:

$$\text{Reflector: } L = \frac{12,562}{F_{\text{MHz}}} \text{ in}$$

$$\text{Director: } L = \frac{11,248}{F_{\text{MHz}}} \text{ in}$$

$$\text{Driven element: } L = \frac{11,826}{F_{\text{MHz}}} \text{ in}$$



18-13 Quad loop construction on VHF.

## VHF/UHF scanner band antennas

The hobby of shortwave listening has always had a subset of adherents who listen exclusively to the VHF/UHF bands. In those bands are found a rich variety of services from commercial two-way radios, police and fire systems, and others. Many of these people are serious DXers, but others have a little of the voyeur in them (they like to listen to what's going on in the community by monitoring the police frequencies). As long as they neither respond to the scene of a police or fire action, nor use the information gained from monitoring in an illegal manner, they are perfectly free to listen in.

A few people have an unusually practical element to their VHF/UHF listening. At least one person known to the author routinely tunes in the local taxicab company's frequency as soon as she orders a cab. She then listens for her own address. She knows from that approximately when to expect the cab.

In the early 1960s, when all VHF monitor receivers were crystal-controlled (and relatively expensive), the listener had a very limited selection of frequencies to choose from. Having worked in various shops that sold those receivers, I can recall that most of our customers tended to be police officers, fire fighters (or volunteers), or journalists covering the local crime news. However, modern scanners operate on both major VHF bands (low and high), plus UHF bands and others.

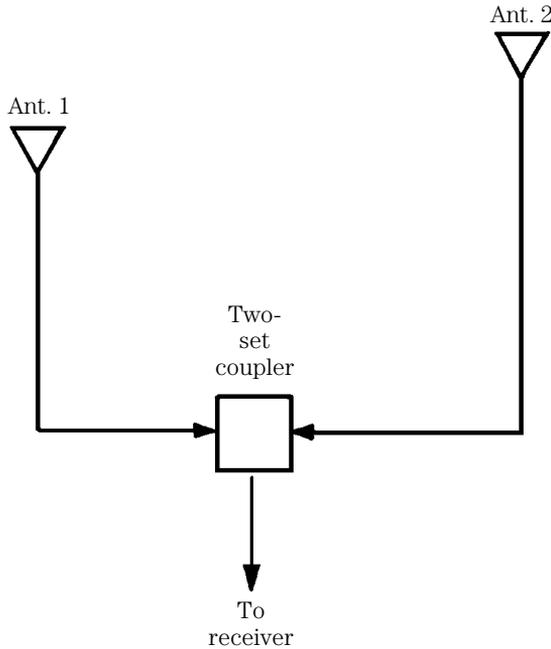
### Scanner-vision antennas

The antennas used by scanner listeners are widely varied, and (in some cases) overpriced. Although it is arguable that a total coverage VHF/UHF antenna is worth the money, and it probably is, there are other possibilities that should be considered.

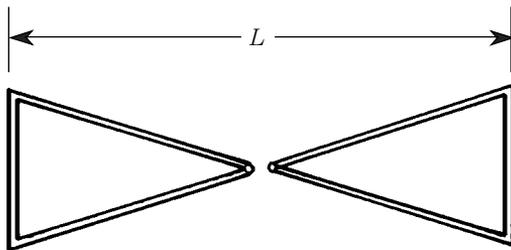
First, don't overlook the use of television antennas for scanner monitoring! The television bands (about 80 channels from 54 MHz to around 800 MHz) encompass most of the ordinarily used scanner frequencies. Although antenna performance is not optimized for the scanner frequencies, it is also not "zero" on those frequencies. If you already have an "all channel" TV antenna installed, then it is a simple matter to connect the antenna to the scanner receiver. (Note: if the antenna uses 300- $\Omega$  twin lead, then install a 4:1 balun transformer that accepts 300  $\Omega$  in and produces 75  $\Omega$  out. These transformers are usually available at TV shops, video shops—including videotape rental places—and Radio Shack stores.)

The directional characteristic of the TV antenna makes it both an advantage and a disadvantage to the scanner user. If the antenna has a rotator, then there is no problem. Just rotate the antenna to the direction of interest. However, if the antenna is fixed, and the station of interest is elsewhere than where the antenna is pointed, then there is a bit of a problem. Nevertheless, some reception is possible for at least three reasons. One is that the main beam of the antenna is not infinitely thin, so the station of interest might well be within the beam—even if off the point a little bit. Second, there are always sidelobes on an antenna. These are areas outside of the main beam that offer reduced, but nonzero, reception characteristics. Finally, the sidelobes and main lobe of the antenna are optimized for the TV bands, and may not obtain the same directivity on certain scanner frequencies. It is, therefore, possible that a TV antenna will have an unusual lobe in the direction of interest for the scanner channel.

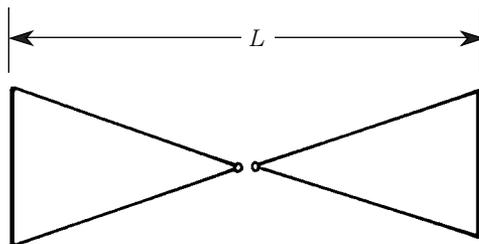
The TV antenna also offers other possibilities. The hardware and electronics, normally found useful on TV antenna systems, are also useful for scanner antennas. The use of the mounting hardware is obvious, but less obvious is the use of components such as multiset couplers, impedance transformers, and wideband amplifiers. For example, Fig. 18-14 shows two scanner antennas joined together into a single



**18-14** Combining two antennas with TV hardware.



**18-15** VHF/UHF "bowtie" dipoles for broadband reception.



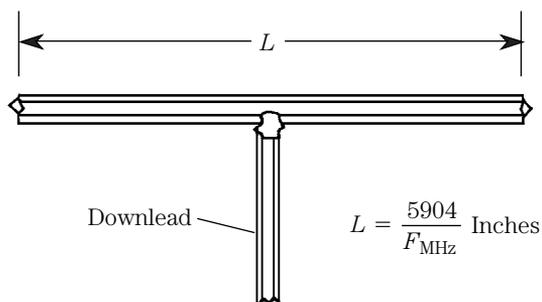
transmission line using a two-set coupler. Although intended to allow two television sets to receive a signal from the same antenna, the device also works to combine two antennas into a single transmission line. Two popular TV uses include two antennas for different directions and VHF and UHF antennas that share the same downlead. Be sure to buy a weatherproof model if you intend to mount the coupler on the antenna mast (the correct place for it) (Fig. 18-15).

TV antenna amplifiers are used to increase the weak signal from a distant station to a level compatible with the TV receiver. Because these amplifiers are wideband designs, they will also work well for the scanner bands that are inclusive within the 54- to 800-MHz TV bands. Some models are intended to be part of the antenna, while others are mounted indoors, some distance from the antenna. The basic rule of thumb is to place the amplifier as close as possible to the “head end” of the system where the antenna is located. This situation will permit signals to be built up to a level considerably above the noise level before being attenuated in the transmission line losses.

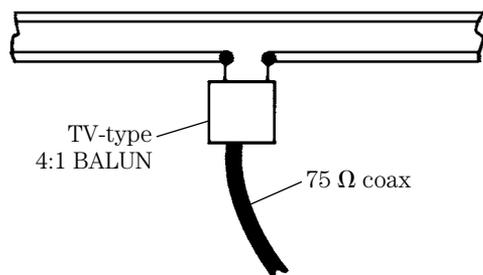
TV/FM receiver folded dipoles made of twin-lead transmission line also work well at other VHF/UHF frequencies. Figures 18-16A and 18-16B show the details of their construction.

### Scanner skyhooks

Other antennas not to be overlooked are the CB, or amateur radio, high-frequency (shortwave) antennas. Many amateur radio operators use their 80-m dipole antenna



**18-16A** Receiver folded dipole antenna made of 300  $\Omega$  twin lead.

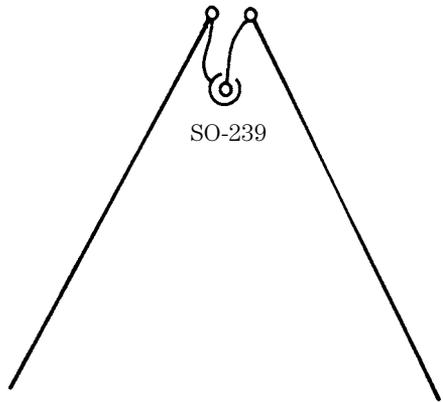


**18-16B** Poles made of twin-lead transmission VHF frequencies.

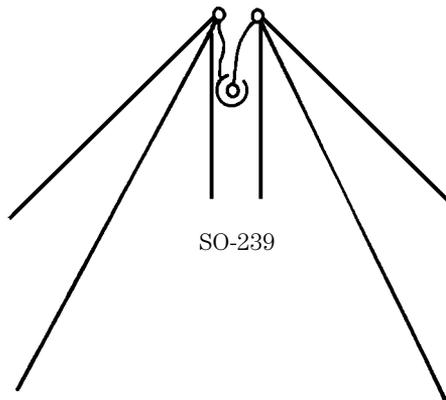
on 2 m (144 to 148 MHz) and find that it has the gain and directivity characteristics of a longwire antenna. Similarly, with vertical “ham” and CB antennas, they have apparent gain on some VHF frequencies and will at least perform to some extent on others.

Even the lowly “random length” wire antenna used for your shortwave receiver ought to turn in decent performance as a VHF/UHF longwire antenna. These antennas are simply a 30- to 150-ft length of no. 14 wire attached to a distant support.

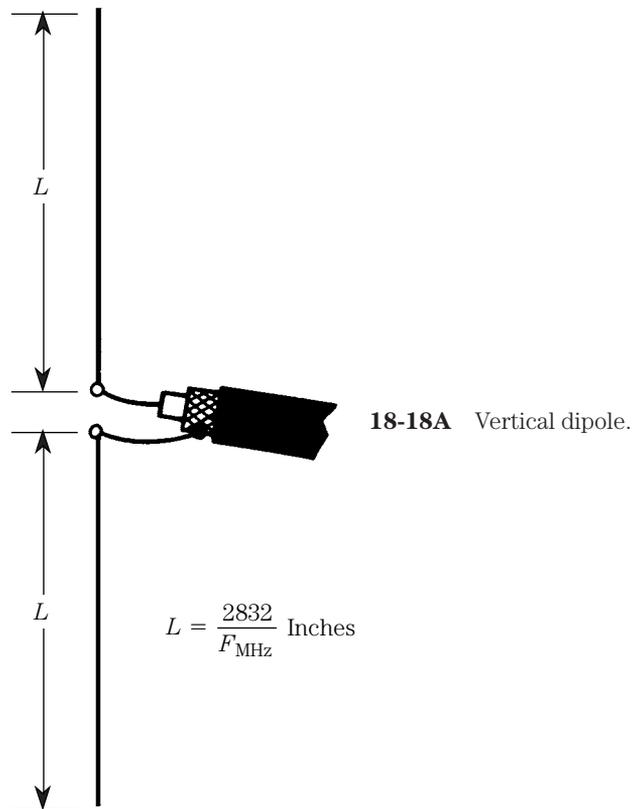
Figure 18-17 shows two variants to an antenna that is popular with VHF/UHF receiver operators: the *drooping dipole*. These antennas are similar to the inverted-vee dipole, in that they consist of two quarter-wavelength radiator elements. Unlike the ordinary dipole, the ends of the radiators, in the single-band drooping dipole of Fig. 18-17A, are lowered. The drooping dipole can be made using an SO-239 UHF coaxial connector as the support for either stiff wire, brass hobbyist tubing, or brazing rod elements cut to  $\lambda/4$  each. The antenna can be made into a multiband, or very wideband, antenna by connecting several pairs of drooping dipoles in parallel, as in Fig. 18-17B.



**18-17A** Drooping dipole.



**18-17B** Multiband drooping dipole.

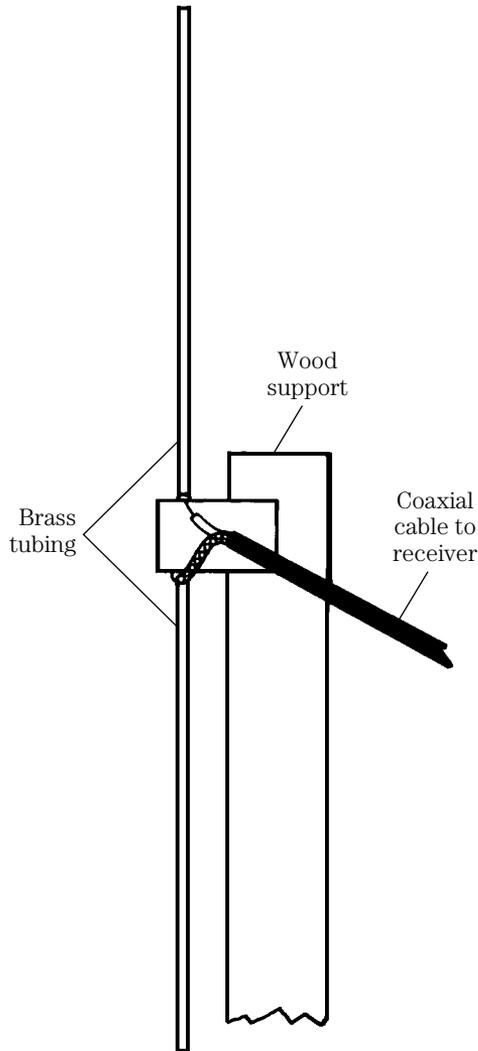


The vertical dipole (Fig. 18-18A) also finds use in the VHF/UHF region. Each element is  $\lambda/4$  long, and its physical length can be found from

$$L = \frac{2832}{F_{\text{MHz}}} \text{ in} \quad \text{[18.12]}$$

This antenna can be built with the same types of materials as the drooping dipole. Although several different construction techniques are popular, the method of Fig. 18-18B is probably the most popular with homebrew builders.

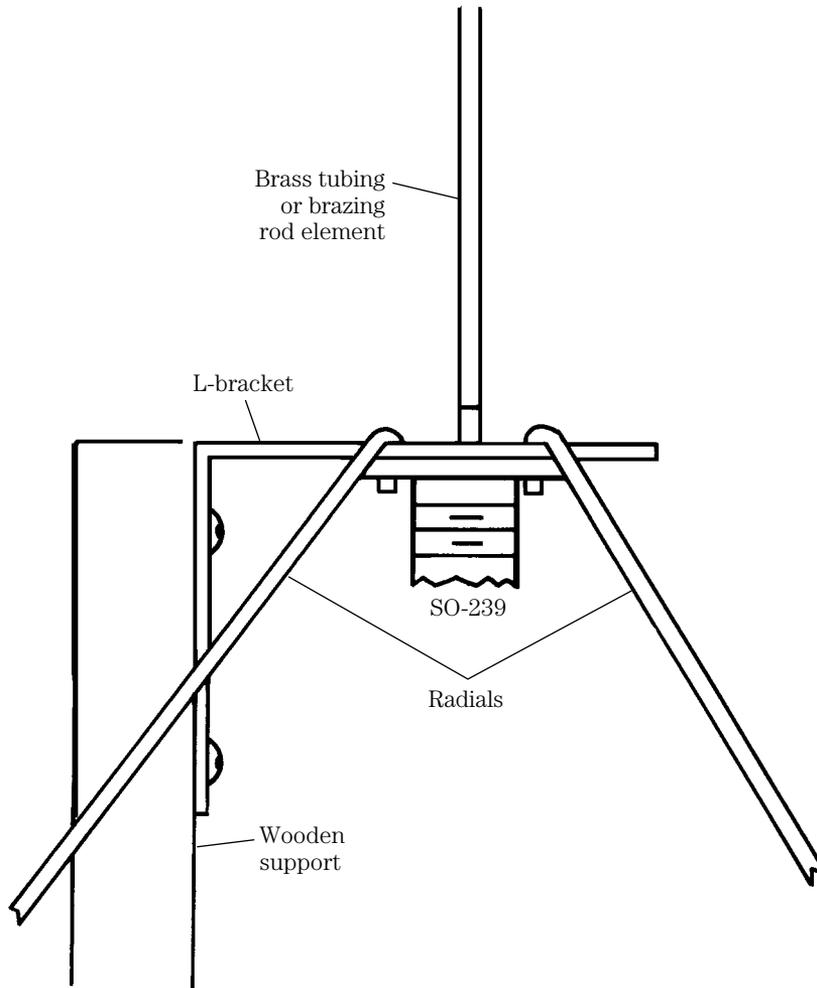
A groundplane antenna for VHF/UHF frequencies is shown in Fig. 18-19. The antenna sits on an L bracket mounted to a  $2 \times 4$ -in wooden mast or support. The L bracket can be manufactured for the purpose, or purchased at Happy Harry's hardware emporium. The base of the antenna is a chassis-mount SO-239 UHF coaxial connector. A quarter-wave radiator element is made from  $1/8$ -in brass tubing, or brazing rod, soldered to the center conductor pin of the SO-239. Similarly, radials can be made from the same type of material, but soldered directly to the body of the SO-239 (this action connects it to the coaxial cable shield). The four mounting holes of the SO-239 make reasonable anchors for the radials. Solder them for electrical integrity.



**18-18B** Construction of VHF vertical dipole.

A coaxial vertical can be made from 1-in brass tubing or copper plumbing pipe, and either brazing rod or brass tubing (Fig. 18-20). The sleeve is fitted with an end cap that has an SO-239 mounted on it (see inset). The radiator element can be made with a mate to the SO-239, namely the PL-259 coaxial plug. The radiator element is soldered to the center conductor, and it is insulated from the shield. You can buy coaxial whips with this type of construction, so you may not have to actually build the radiator element.

Additional gain, about +3 dB, can be achieved by stacking VHF/UHF antennas together. Figure 18-21 shows a typical arrangement in which two half-wavelength dipole antennas are connected together through a quarter-wavelength harness of RG-59/U coaxial cable. This harness is shorter than a quarter wavelength by the velocity factor of the coaxial cable:



**18-19** Construction of VHF groundplane for receivers.

$$L = \frac{2832V}{F_{\text{MHz}}} \quad [18.13]$$

where

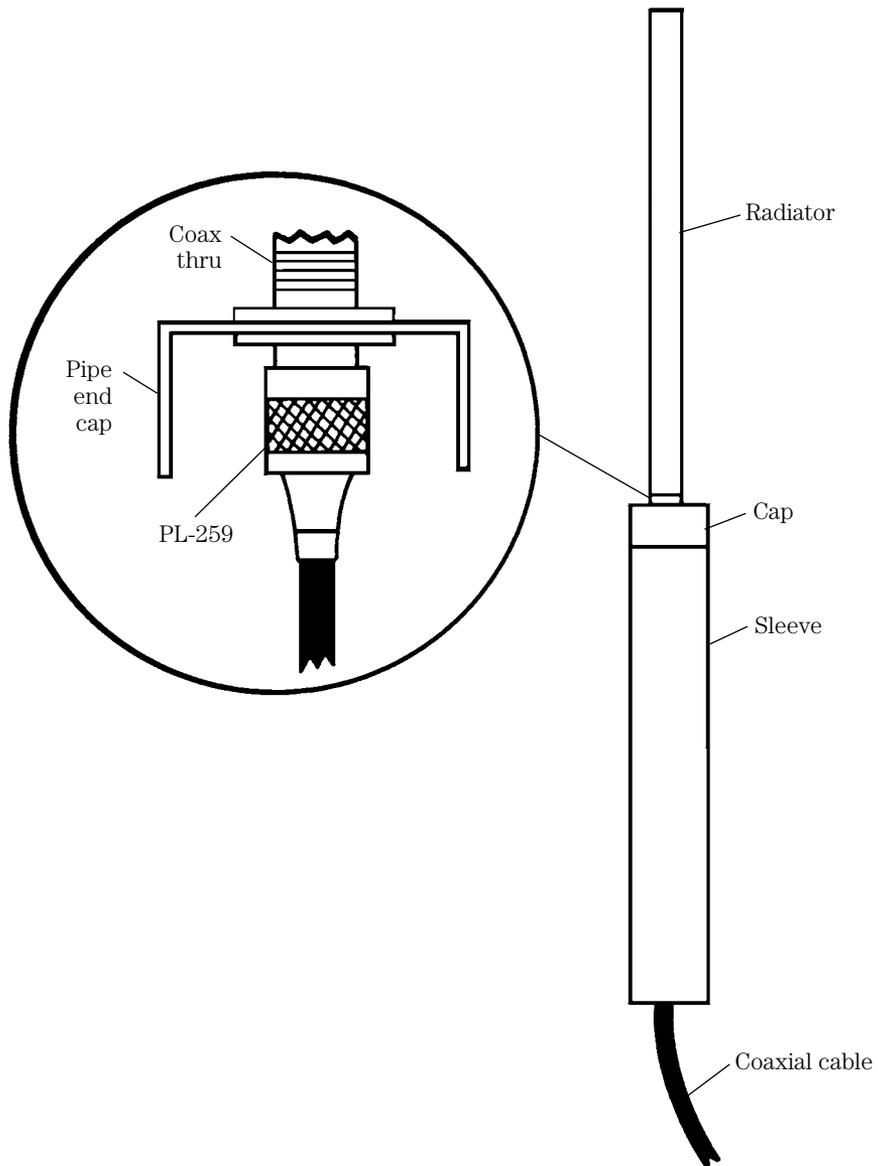
$L$  is the length, in inches

$V$  is the velocity factor (typically 0.66 or 0.80 for common coax)

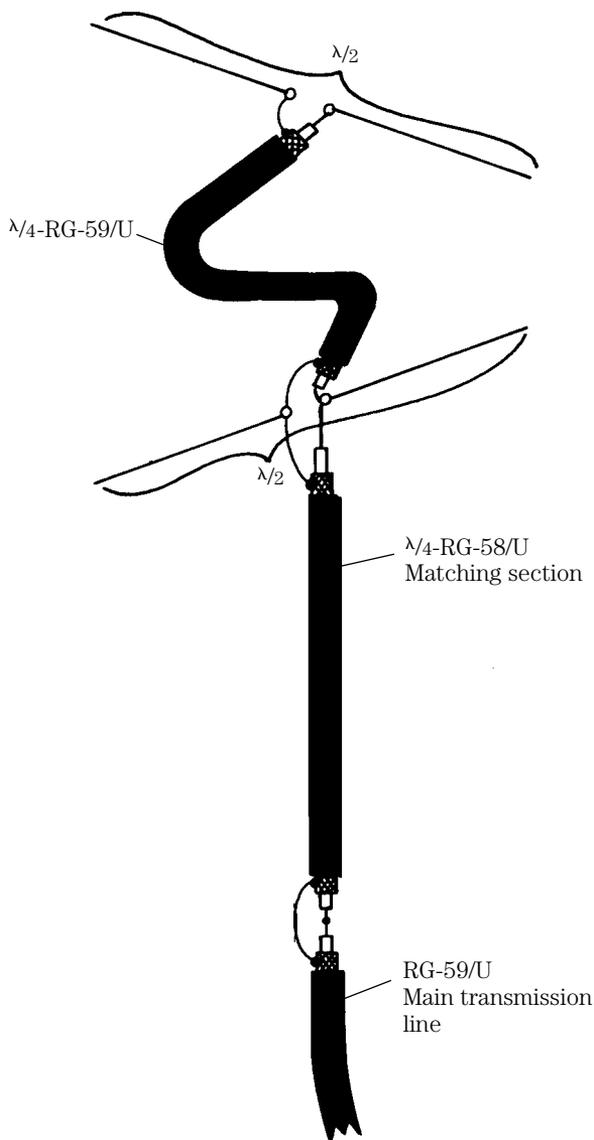
$F_{\text{MHz}}$  is the frequency, in megahertz

The antennas can be oriented in the same direction to increase gain, or orthogonally (as shown in Fig. 18-21) to make it more omnidirectional.

Because the impedance of two identical dipoles, fed in parallel, is one-half that of a single dipole, it is necessary to have an impedance-matching section



18-20 Coaxial dipole vertical.



18-21 Stacking VHF antennas.

made of RG-58/U coaxial cable. This cable is then fed with RG-59/U coax from the receiver.

There is nothing magical about scanner receivers that require any form of antenna that is significantly different from other VHF/UHF antennas. Although the designs might be optimized for VHF or UHF, these antennas are basically the same as others shown in this book. As a matter of fact, almost any antennas, from any chapter, can be used by at least some scanner operators.

## 19 CHAPTER

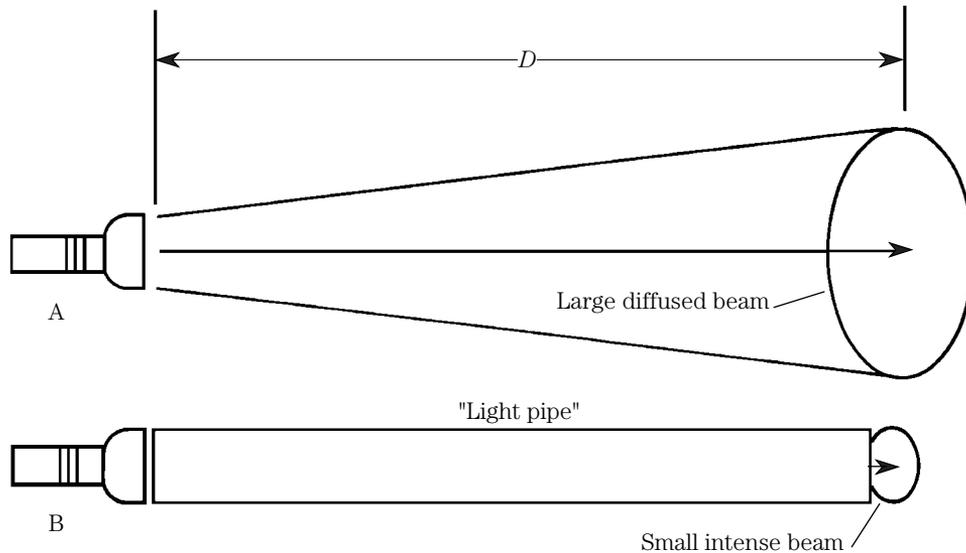
# Microwave waveguides and antennas

THE MICROWAVE PORTION OF THE RADIO SPECTRUM COVERS FREQUENCIES FROM ABOUT 900 MHz to 300 GHz, with wavelengths in free-space ranging from 33 cm down to 1 mm. Transmission lines can be used at frequencies from dc to about 50 or 60 GHz, although, above 5 GHz, only short runs are practical, because attenuation increases dramatically as frequency increases. There are three types of losses in conventional transmission lines: *ohmic*, *dielectric*, and *radiation*. The ohmic losses are caused by the current flowing in the resistance of the conductors making up the transmission lines. Because of the skin effect, which increases resistance at higher frequencies, these losses tend to increase in the microwave region. Dielectric losses are caused by the electric field acting on the molecules of the insulator and thereby causing heating through molecular agitation. Radiation losses represent loss of energy as an electromagnetic wave propagates away from the surface of the transmission line conductor.

Losses on long runs of coaxial transmission line (the type most commonly used) cause concern even as low as the 400-MHz region. Also, because of the increased losses, power handling capability decreases at high frequencies. Therefore, at higher microwave frequencies, or where long runs make coax attenuation losses unacceptable, or where high power levels would overheat the coax, *waveguides* are used in lieu of transmission lines.

What is a waveguide? Consider the light pipe analogy depicted in Fig. 19-1. A flashlight serves as our “RF source,” which (given that light is also an electromagnetic wave) is not altogether unreasonable. In Fig. 19-1A the source radiates into free space and spreads out as a function of distance. The intensity per unit area, at the destination (a wall), falls off as a function of distance ( $D$ ) according to the *inverse square law* ( $1/D^2$ ).

But now consider the transmission scheme in Fig. 19-1B. The light wave still propagates over distance  $D$ , but is now confined to the interior of a mirrored pipe. Almost all of the energy (less small losses) coupled to the input end is delivered to the output end, where the intensity is practically undiminished. Although not



**19-1** Waveguide analogy to light pipe.

perfect, the light pipe analogy neatly summarizes, on a simple level, the operation of microwave waveguides.

Thus, we can consider the waveguide as an “RF pipe” without seeming too serenely detached from reality. Similarly, fiber-optic technology is waveguidelike at optical (IR and visible) wavelengths. In fact, the analogy between fiber optics and waveguide can withstand more rigorous comparison than the simplistic light pipe analogy.

The internal walls of the waveguide are not mirrored surfaces, as in our optical analogy, but are, rather, electrical conductors. Most waveguides are made of aluminum, brass, or copper. In order to reduce ohmic losses, some waveguides have their internal surfaces electroplated with either gold or silver, both of which have lower resistivities than the other metals mentioned above.

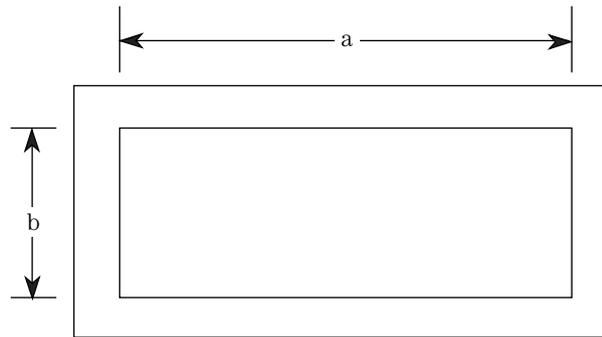
Waveguides are hollow metal pipes, and can have either circular or rectangular cross sections (although the rectangular are, by far, the most common). Figure 19-2 shows an end view of the rectangular waveguide. The dimension  $a$  is the wider dimension, and  $b$  is the narrower. These letters are considered the standard form of notation for waveguide dimensions, and will be used in the equations developed in this chapter.

## Development of the rectangular waveguide from parallel transmission lines

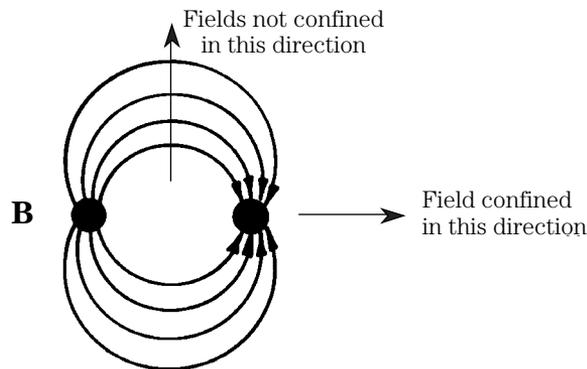
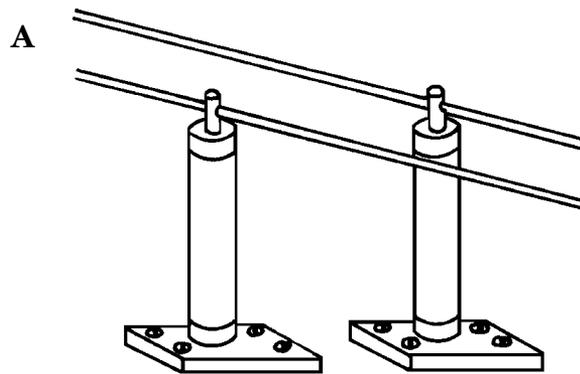
One way of visualizing how a waveguide works is to develop the theory of waveguides from the theory of elementary parallel transmission lines (see Chap. 3). Figure 19-3A shows the basic parallel transmission line which was introduced in Chap. 3. The line

consists of two parallel conductors separated by an air dielectric. Because air won't support the conductors, ceramic or other material insulators are used as supports.

There are several reasons why the parallel transmission line *per se* is not used at microwave frequencies. Skin effect increases ohmic losses to a point that is unacceptable. Also, the insulators supporting the two conductors are significantly more



19-2 Rectangular waveguide (end view).



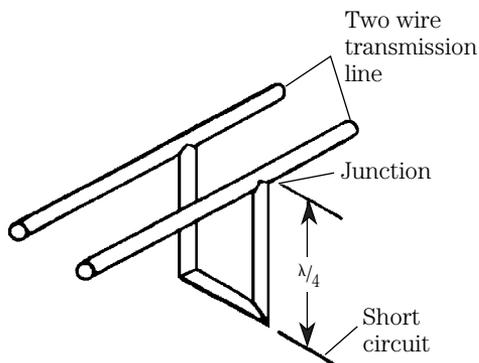
End view of two wire line

19-3 Development of waveguide from parallel line.

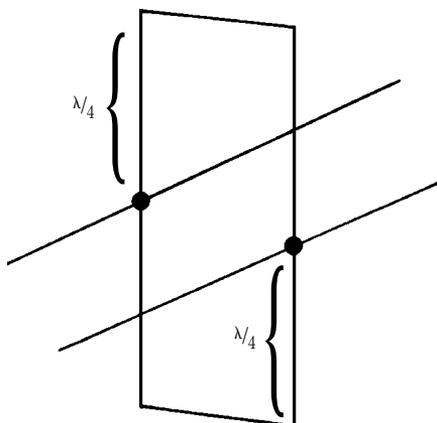
lossy at microwave frequencies than at lower frequencies. Finally, radiation losses increase dramatically. Figure 19-3B shows the electric fields surrounding the conductors. The fields add algebraically (either constructively or destructively), resulting in pinching of the resultant field along one axis, and bulging along the other. This geometry increases radiation losses at microwave frequencies.

Now let's consider the quarter-wavelength shorted stub. The "looking-in" impedance of such a stub is infinite. When placed in parallel across a transmission line (Fig. 19-4A) the stub acts like an insulator. In other words, at its resonant frequency, the stub is a *metallic insulator*, and can be used to physically support the transmission line.

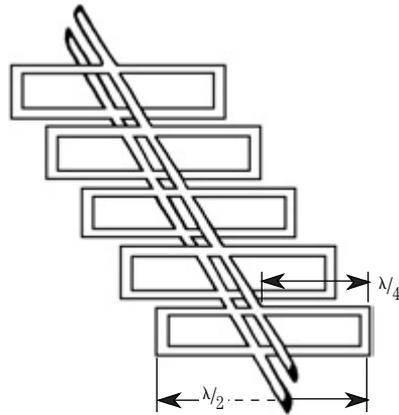
Again, because the impedance is infinite, we can connect two quarter-wavelength stubs in parallel with each other across the same points on the transmission line (Fig. 19-4B) without loading down the line impedance. This arrangement effectively forms a half-wavelength pair. The impedance is still infinite, so no harm is done. Likewise, we can parallel a large number of center-fed half-wavelength pairs along the line, as might be the case when a long line is supported at multiple points. The waveguide is analogous to an *infinite number of*



**19-4A**  
Quarter-wave stub analogy.



**19-4B**  
Quarter-wave stub analogy extended.



**19-4C**  
Quarter-wave stub analogy  
extended even further.

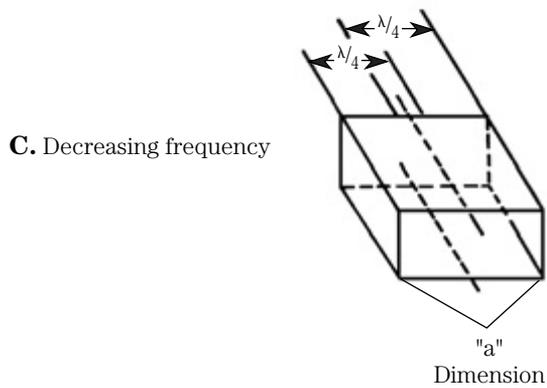
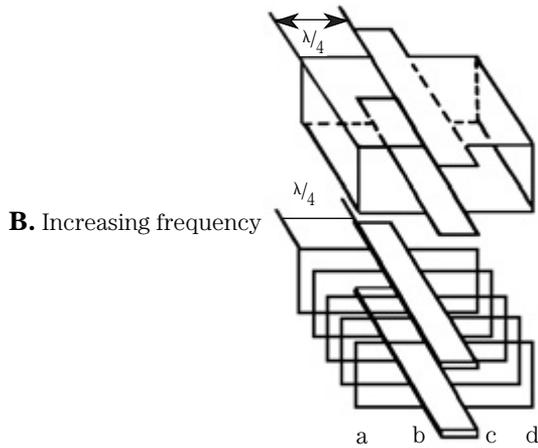
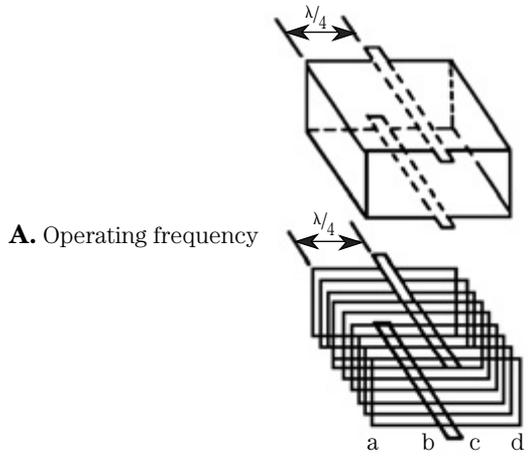
center-fed “half-wave pairs” of quarter-wave shorted stubs connected across the line. The result is the continuous metal pipe structure of the common rectangular waveguide (Fig. 19-4C).

On first glance, relating rectangular waveguide to quarter-wavelength shorted stubs seems to fall down, except at the exact resonant frequency. It turns out, however, that the analogy also holds up at other frequencies, so long as the frequency is higher than a certain minimum cutoff frequency. The waveguide thus acts like a high-pass filter. There is also a practical upper frequency limit. In general, waveguides support a bandwidth of 30 to 40 percent of cutoff frequency. As shown in Fig. 19-5, the center line of the waveguide (which represents the points where the conductors are in the parallel line analogy) becomes a “shorting bar” between segments, and that “bar” widens or narrows according to operating frequency. Thus, the active region is still a quarter-wavelength shorted stub.

Below the cutoff frequency, the structure disappears entirely, and the waveguide acts like a parallel transmission line with a low-impedance inductive reactance shorted across the conductors. When modeled as a pair of quarter-wavelength stubs, the  $a$  dimension of the waveguide is a half-wavelength long. The cutoff frequency is defined as the frequency at which the  $a$  dimension is less than a half-wavelength.

## Propagation modes in waveguides

The signal in a microwave waveguide propagates as an electromagnetic wave, not as a current. Even in a transmission line, the signal propagates as a wave because the current in motion down the line gives rise to electric and magnetic fields, which behave as an electromagnetic field. The specific type of field found in transmission lines, however, is a *transverse electromagnetic (TEM)* field. The term *transverse* implies things at right angles to each other, so the electric and magnetic fields are perpendicular to the direction of travel. In addition to the word “transverse,” these right-angle waves are said to be “normal” or “orthogonal” to the direction of travel—three different ways of saying the same thing: right-angledness.



**19-5**  
Changing frequency does not affect the analogy.

### Boundary conditions

The TEM wave will not propagate in a waveguide because certain *boundary conditions* apply. Although the wave in the waveguide propagates through the air (or inert gas dielectric) in a manner similar to free-space propagation, the phenomenon is bounded by the walls of the waveguide, and that implies certain conditions must be met. The boundary conditions for waveguides are

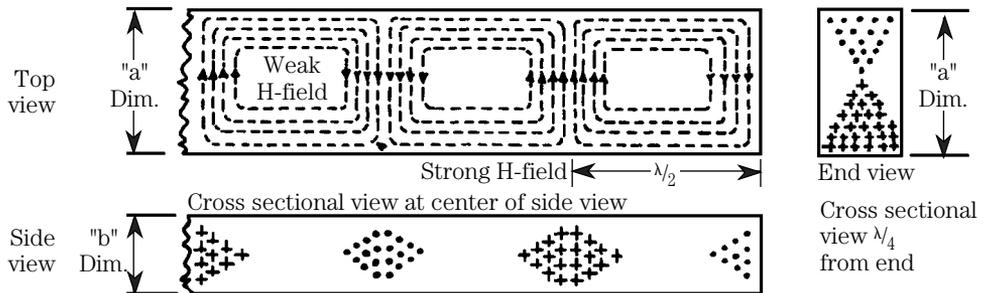
1. The electric field must be orthogonal to the conductor in order to exist at the surface of that conductor.
2. The magnetic field must not be orthogonal to the surface of the waveguide.

In order to satisfy these boundary conditions the waveguide gives rise to two types of propagation modes: *transverse electric mode* (TE mode), and *transverse magnetic mode* (TM mode). The TEM mode violates the boundary conditions because the magnetic field is not parallel to the surface, and so does not occur in waveguides.

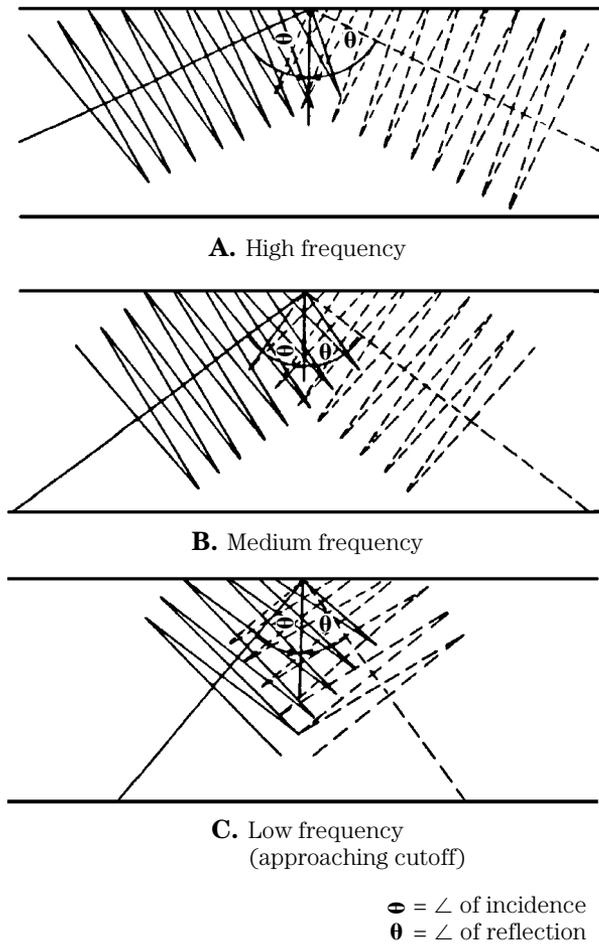
The transverse electric field requirement means that the E field must be perpendicular to the conductor wall of the waveguide. This requirement is met by use of a proper coupling scheme at the input end of the waveguide. A vertically polarized coupling radiator will provide the necessary transverse field.

One boundary condition requires that the magnetic (H) field must not be orthogonal to the conductor surface. Because it is at right angles to the E field, it will meet this requirement (see Fig. 19-6). The planes formed by the magnetic field are parallel to both the direction of propagation and the wide dimension surface.

As the wave propagates away from the input radiator, it resolves into two components that are not along the axis of propagation, and are not orthogonal to the walls. The component along the waveguide axis violates the boundary conditions, so it is rapidly attenuated. For the sake of simplicity, only one component is shown in Fig. 19-7. Three cases are shown in Fig. 19-7: high, medium, and low frequency. Note that the angle of incidence with the waveguide wall increases as frequency drops. The angle rises toward 90° as the cutoff frequency is approached from above. Below the cutoff frequency the angle is 90°, so the wave bounces back and forth between the walls without propagating.



19-6 Magnetic fields in waveguide.



**19-7**  
Frequency effect on propagating wave.

### Coordinate system and dominant mode in waveguides

Figure 19-8 shows the coordinate system used to denote dimensions and directions in microwave discussions. The  $a$  and  $b$  dimensions of the waveguide correspond to the  $x$  and  $y$  axes of a cartesian coordinate system, and the  $z$  axis is the direction of wave propagation.

In describing the various modes of propagation, use a shorthand notation as follows:

$$Tx_{m,n}$$

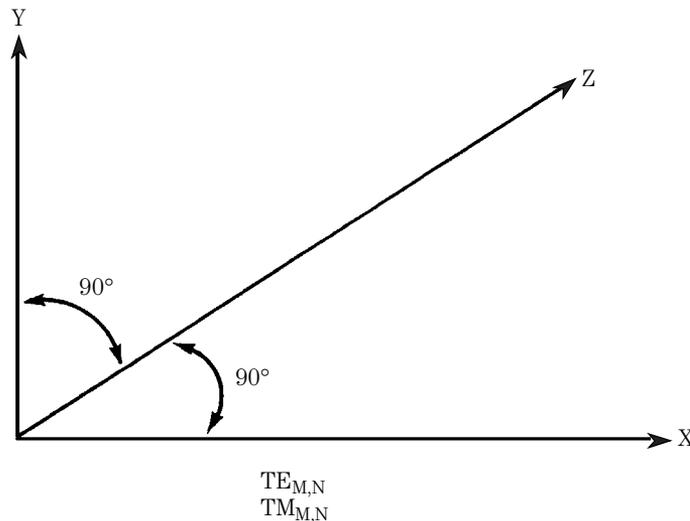
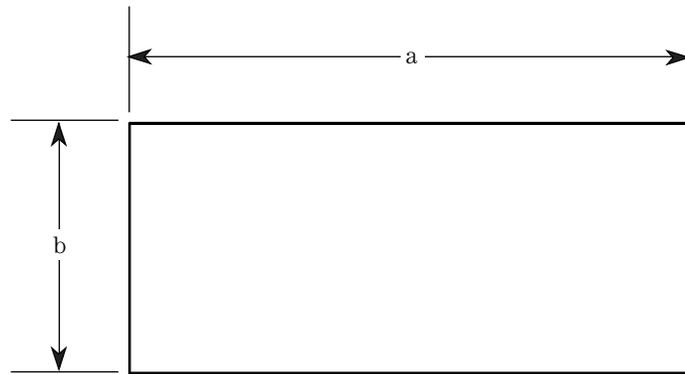
where

- $x$  is E for transverse electric mode, and M for transverse magnetic mode
- $m$  is the number of half-wavelengths along the  $x$  axis (i.e.,  $a$  dimension)
- $n$  is the number of half-wavelengths along the  $y$  axis (i.e.,  $b$  dimension)

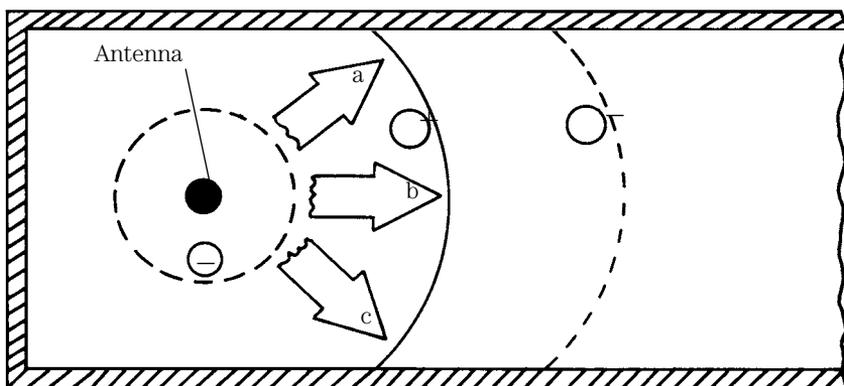
The  $TE_{10}$  mode is called the *dominant mode*, and is the best mode for low attenuation propagation in the  $z$  axis. The nomenclature  $TE_{10}$  indicates that there is one half-wavelength in the  $a$  dimension and zero half-wavelengths in the  $b$  dimension. The dominant mode exists at the lowest frequency at which the waveguide is a half-wavelength.

### Velocity and wavelength in waveguides

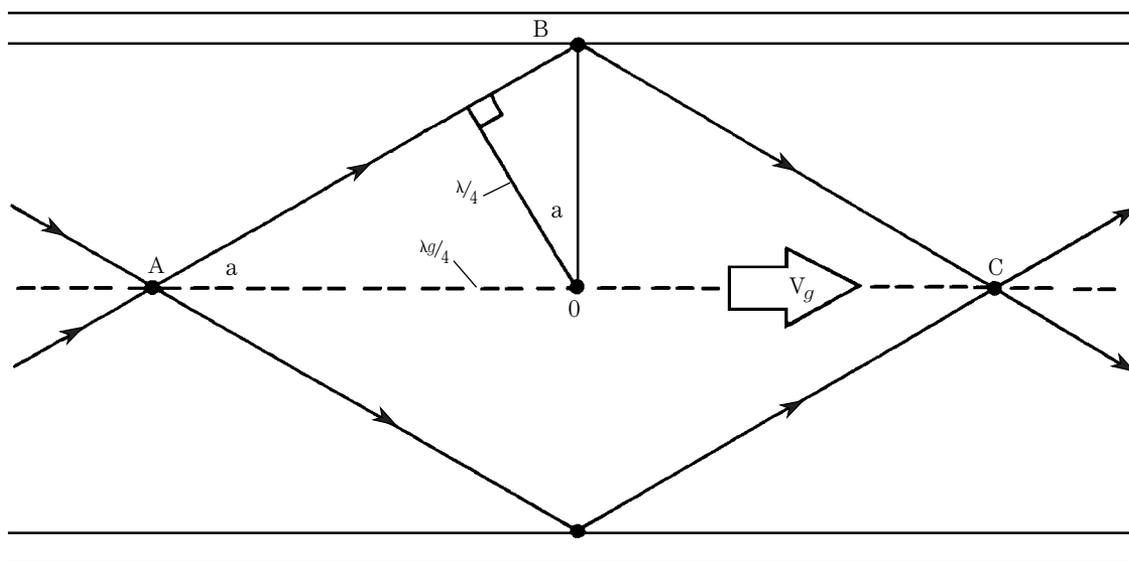
Figures 19-9A and 19-9B show the geometry for two wave components, simplified for the sake of illustration. There are three different wave velocities to consider with respect to waveguides: *free-space velocity*  $c$ , *group velocity*  $V_g$ , and *phase velocity*  $V_p$ . The free-space velocity is the velocity of propagation in unbounded free space (i.e., the speed of light  $c = 3 \times 10^8$  m/s).



**19-8** Rectangular waveguide coordinate system.



**19-9A** Antenna radiator in capped waveguide.



**19-9B** Wave propagation in waveguide.

The *group velocity* is the straight line velocity of propagation of the wave down the center line ( $z$  axis) of the waveguides. The value of  $V_g$  is always less than  $c$ , because the actual path length taken, as the wave bounces back and forth, is longer than the straight line path (i.e., path  $ABC$  is longer than path  $AC$ ). The relationship between  $c$  and  $V_g$  is

$$V_g = c \sin a \quad [19.1]$$

where

$V_g$  is the group velocity, in meters per second  
 $c$  is the free-space velocity ( $3 \times 10^8$  m/s)  
 $a$  is the angle of incidence in the waveguide

The *phase velocity* is the velocity of propagation of the spot on the waveguide wall where the wave impinges (e.g., point  $B$  in Fig. 19-9B). This velocity, depending upon the angle of incidence, can actually be faster than both the group velocity and the speed of light. The relationship between phase and group velocities can be seen in the beach analogy. Consider an ocean beach, on which the waves arrive from offshore at an angle other than  $90^\circ$ . In other words, the arriving wavefronts are not parallel to the shore. The arriving waves have a group velocity  $V_g$ . But as a wave hits the shore, it will strike a point down the beach first, and the “point of strike” races up the beach at a much faster phase velocity  $V_p$ , that is even faster than the group velocity. In a microwave waveguide, the phase velocity can be greater than  $c$ , as can be seen from Eq. 19.2:

$$V_p = \frac{c}{\sin a} \quad [19.2]$$

**Example 19-1** Calculate the group and phase velocities for an angle of incidence of  $33^\circ$ .

**Solution:**

(a) *Group velocity*

$$\begin{aligned} V_g &= c \sin a \\ &= (3 \times 10^8)(\sin 33^\circ) \\ &= (3 \times 10^8)(0.5446) = 1.6 \times 10^8 \text{ m/s} \end{aligned}$$

(b) *Phase velocity*

$$\begin{aligned} V_p &= c/\sin a \\ &= (3 \times 10^8 \text{ m/s})/\sin 33^\circ \\ &= (3 \times 10^8 \text{ m/s})/(0.5446) \\ &= 5.51 \times 10^8 \text{ m/s} \end{aligned}$$

For this problem the solutions are

$$\begin{aligned} c &= 3 \times 10^8 \text{ m/s} \\ V_p &= 5.51 \times 10^8 \text{ m/s} \\ V_g &= 1.6 \times 10^8 \text{ m/s} \end{aligned}$$

We can also write a relationship between all three velocities by combining Eqs. 19.1 and 19.2, resulting in

$$c = \sqrt{V_p V_g} \quad [19.3]$$

In any wave phenomenon the product of *frequency* and *wavelength* is the *velocity*. Thus, for a TEM wave in unbounded free space we know that:

$$c = F \lambda_o \quad [19.4]$$

Because the frequency  $F$  is fixed by the generator, only the wavelength can change when the velocity changes. In a microwave waveguide we can relate phase velocity to wavelength as the wave is propagated in the waveguide:

$$V_p = \frac{\lambda \times c}{\lambda_o} \quad [19.5]$$

where:

$V_p$  is the phase velocity, in meters per second

$c$  is the free-space velocity ( $3 \times 10^8$  m/s)

$\lambda$  is the wavelength in the waveguide, in meters

$\lambda_o$  is the wavelength in free space ( $c/F$ ), in meters (see Eq. 19.4)

Equation 19.5 can be rearranged to find the wavelength in the waveguide:

$$\lambda = \frac{V_p \lambda_o}{c} \quad [19.6]$$

**Example 19-2** A 5.6-GHz microwave signal is propagated in a waveguide. Assume that the internal angle of incidence to the waveguide surfaces is 42 degrees. Calculate (a) phase velocity, (b) wavelength in unbounded free space, and (c) wavelength of the signal in the waveguide.

**Solution:**

(a) *Phase velocity*

$$\begin{aligned} V_p &= \frac{c}{\sin a} \\ &= \frac{3 \times 10^8 \text{ m/s}}{\sin 42^\circ} \\ &= \frac{3 \times 10^8 \text{ m/s}}{0.6991} = 4.5 \times 10^8 \text{ m/s} \end{aligned}$$

(b) *Wavelength in free space*

$$\begin{aligned} \lambda_o &= c/F \\ &= (3 \times 10^8 \text{ m/s}) / (5.6 \times 10^9 \text{ Hz}) \\ &= 0.054 \text{ m} \end{aligned}$$

(c) *Wavelength in waveguide*

$$\begin{aligned}\lambda &= \frac{V_p \lambda_o}{c} \\ &= \frac{(4.5 \times 10^8 \text{ m/s})(0.054 \text{ m})}{3 \times 10^8 \text{ m/s}} = 0.08 \text{ m}\end{aligned}$$

Comparing, we find that the free-space wavelength is 0.054 m, and the wavelength inside of the waveguide increases to 0.08 m.

## Cutoff frequency ( $F_c$ )

The propagation of signals in a waveguide depends, in part, upon the operating frequency of the applied signal. As covered earlier, the angle of incidence made by the plane wave to the waveguide wall is a function of frequency. As the frequency drops, the angle of incidence increases toward  $90^\circ$ .

The propagation of waves depends on the angle of incidence and the associated reflection phenomena. Indeed, both phase and group velocities are functions of the angle of incidence. When the frequency drops to a point where the angle of incidence is  $90^\circ$ , then group velocity is meaningless.

We can define a general mode equation based on our system of notation:

$$\frac{1}{(\lambda_c)^2} = \left(\frac{m}{2a}\right)^2 + \left(\frac{n}{2b}\right)^2 \quad [19.7]$$

where

$\lambda_c$  is the longest wavelength that will propagate

$a$ ,  $b$  are the waveguide dimensions (see Fig. 19-2)

$m$ ,  $n$  are integers that define the number of half-wavelengths that will fit in the  $a$  and  $b$  dimensions, respectively

Evaluating Eq. 19.7 reveals that the longest TE-mode signal that will propagate in the dominant mode ( $\text{TE}_{10}$ ) is given by

$$\lambda_c = 2a \quad [19.8]$$

from which we can write an expression for the cutoff frequency:

$$F_c = \frac{c}{2a} \quad [19.9]$$

where

$F_c$  is the lowest frequency that will propagate, in hertz

$c$  is the speed of light ( $3 \times 10^8$  m/s)

$a$  is the wide waveguide dimension

**Example 19-3** A rectangular waveguide has dimensions of  $3 \times 5$  cm. Calculate the  $TE_{10}$  mode cutoff frequency.

**Solution:**

$$\begin{aligned} F_c &= \frac{c}{2a} \\ &= \frac{(3 \times 10^8 \text{ m/s})}{(2) \left( 5 \text{ cm} \times \frac{1 \text{ m}}{100 \text{ cm}} \right)} \\ &= \frac{3 \times 10^8 \text{ m/s}}{(2)(0.05 \text{ m})} = 3 \text{ GHz} \end{aligned}$$

Equation 19.7 assumes that the dielectric inside the waveguide is air. A more generalized form, which can accommodate other dielectrics, is

$$F_c = \frac{1}{2\sqrt{ue}} \sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2} \quad [19.10]$$

where

$e$  is the dielectric constant

$u$  is the permeability constant

For air dielectrics,  $u = u_o$  and  $e = e_o$ , from which

$$c = \frac{1}{\sqrt{u_o e_o}} \quad [19.11]$$

To determine the cutoff wavelength, we can rearrange Eq. 19.10 to the form:

$$\lambda_c = \frac{2}{\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2}} \quad [19.12]$$

One further expression for air-filled waveguide calculates the actual wavelength in the waveguide from a knowledge of the free-space wavelength and actual operating frequency:

$$\lambda_g = \frac{\lambda_o}{\sqrt{1 - \left(\frac{F_c}{F}\right)^2}} \quad [19.13]$$

where

$\lambda_g$  is the wavelength in the waveguide

$\lambda_o$  is the wavelength in free space

$F_c$  is the waveguide cutoff frequency

$F$  is the operating frequency

**Example 19-4** A waveguide with a 4.5-GHz cutoff frequency is excited with a 6.7-GHz signal. Find (a) the wavelength in free space and (b) the wavelength in the waveguide.

**Solution:**

(a)

$$\begin{aligned}\lambda_o &= c/F \\ &= \frac{3 \times 10^8 \text{ m/s}}{6.7 \text{ GHz} \times \frac{10^9 \text{ Hz}}{1 \text{ GHz}}} \\ &= \frac{3 \times 10^8 \text{ m/s}}{6.7 \times 10^9 \text{ Hz}} = 0.0448 \text{ m}\end{aligned}$$

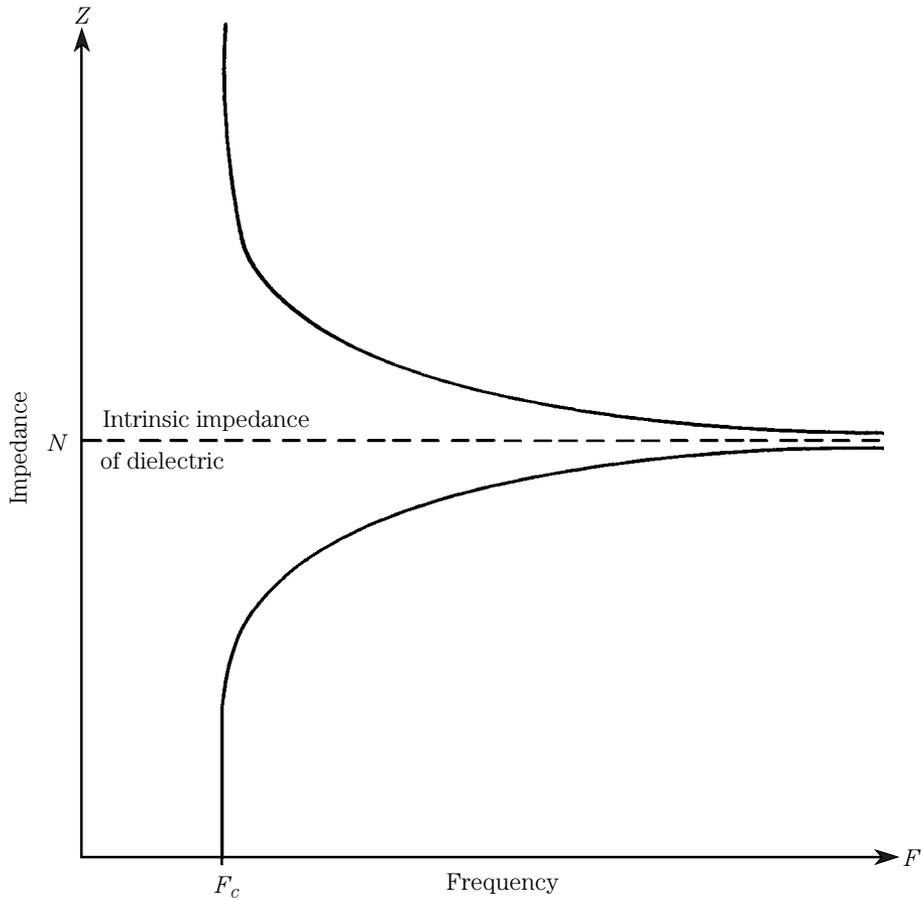
(b)

$$\begin{aligned}\lambda_g &= \frac{\lambda_o}{\sqrt{1 - \left(\frac{F_c}{F}\right)^2}} \\ &= \frac{0.0448 \text{ m}}{\sqrt{1 - \left(\frac{4.5 \text{ GHz}}{6.7 \text{ GHz}}\right)^2}} \\ &= \frac{0.0448 \text{ m}}{1 - 0.67} \\ &= \frac{0.0448}{0.33} = 0.136 \text{ m}\end{aligned}$$

Transverse magnetic modes also propagate in waveguides, but the base  $\text{TM}_{10}$  mode is excluded by the boundary conditions. Thus, the  $\text{TM}_{11}$  mode is the lowest magnetic mode that will propagate.

## Waveguide impedance

All forms of transmission line, including the waveguide, exhibit a characteristic impedance, although in the case of waveguide it is a little difficult to pin down conceptually. This concept was developed for ordinary transmission lines in Chap. 3. For a waveguide, the characteristic impedance is approximately equal to the ratio of the electric and magnetic fields ( $E/H$ ), and converges (as a function of frequency) to the intrinsic impedance of the dielectric (Fig. 19-10). The impedance of the waveguide is a function of waveguide characteristic impedance ( $Z_o$ ) and the wavelength in the waveguide:

**19-10** Impedance versus frequency.

$$Z = \frac{Z_o \lambda_g}{\lambda_o} \quad [19.14]$$

Or, for rectangular waveguide, with constants taken into consideration:

$$Z = \frac{120 \pi \lambda_g}{\lambda_o} \quad [19.15]$$

The *propagation constant*  $B$  for rectangular waveguide is a function of both cutoff frequency and operating frequency:

$$B = W \sqrt{eu} \sqrt{1 - \left(\frac{F_c}{F}\right)^2} \quad [19.16]$$

from which we can express the TE-mode impedance:

$$Z_{\text{TE}} = \frac{\sqrt{ue}}{\sqrt{1 - \left(\frac{F_c}{F}\right)^2}} \quad [19.17]$$

and the TM-mode impedance:

$$Z_{\text{TM}} = 377 \sqrt{1 - \left(\frac{F_c}{F}\right)^2} \quad [19.18]$$

## Waveguide terminations

When an electromagnetic wave propagates down a waveguide, it must eventually reach the end of the guide. If the end is open, then the wave will propagate into free-space. The horn radiator is an example of an unterminated waveguide. If the waveguide terminates in a metallic wall, then the wave reflects back down the waveguide, from whence it came. The interference between incident and reflected waves forms standing waves (see Chap. 3). Such waves are stationary in space, but vary in the time domain.

In order to prevent standing waves, or more properly, the reflections that give rise to standing waves, the waveguide must be *terminated* in a matching impedance. When a properly designed antenna is used to terminate the waveguide, it forms the matched load required to prevent reflections. Otherwise, a *dummy load* must be provided. Figure 19-11 shows several types of dummy load.

The classic termination is shown in Fig. 19-11A. The “resistor” making up the dummy load is a mixture of sand and graphite. When the fields of the propagated wave enter the load, they cause currents to flow, which in turn cause heating. Thus, the RF power dissipates in the sand-graphite rather than being reflected back down the waveguide.

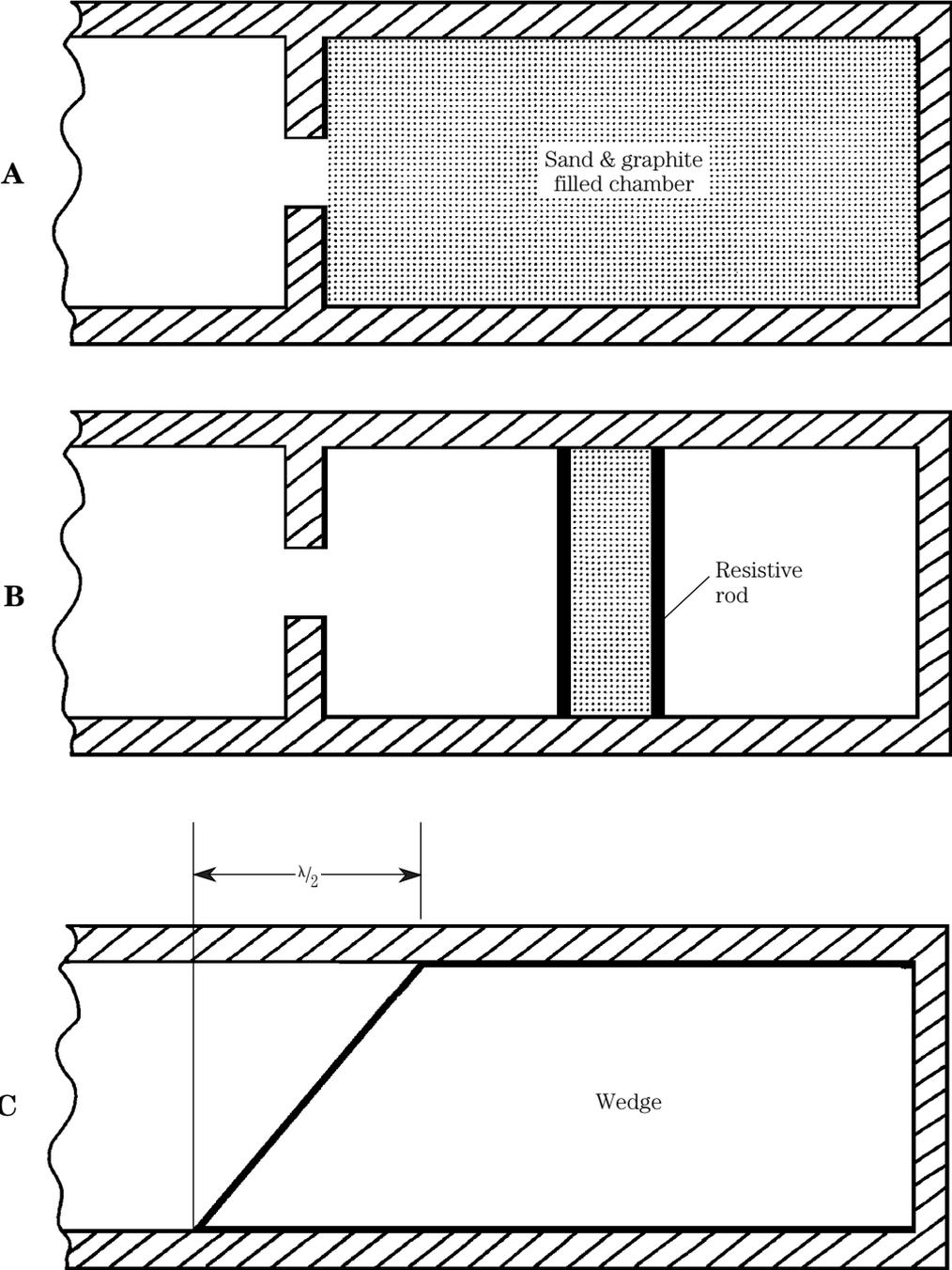
A second dummy load is shown in Fig. 19-11B. The resistor element is a carbonized rod critically placed at the center of the electric field. The E field causes currents to flow, resulting in  $I^2R$  losses that dissipate the power.

Bulk loads, similar to the graphite-sand chamber, are shown in Fig. 19-11C, D, and E. Using bulk material such as graphite or a carbonized synthetic material, these loads are used in much the same way as the sand load (i.e., currents set up, and  $I^2R$  losses dissipate the power).

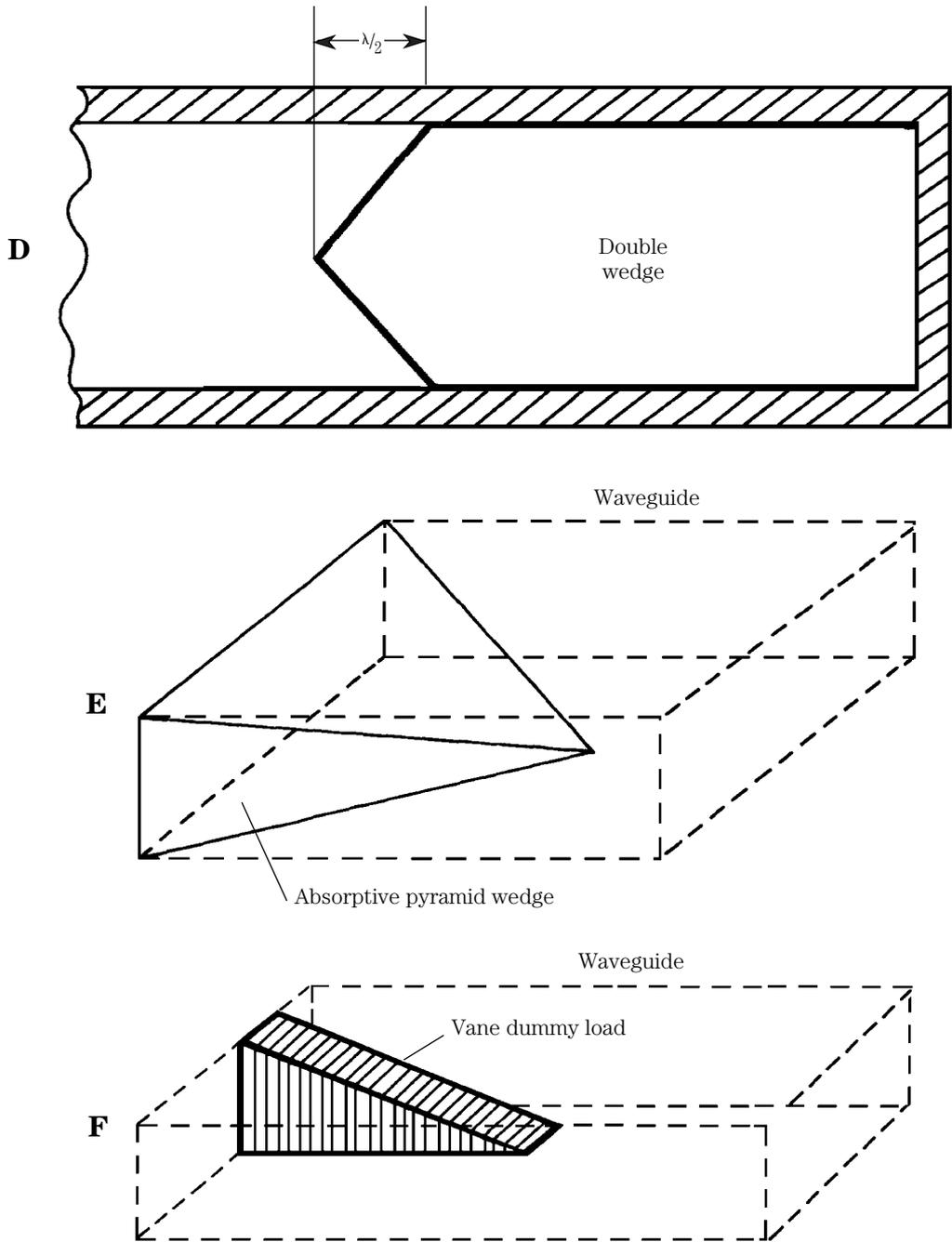
The resistive vane load is shown in Fig. 19-11F. The plane of the element is orthogonal to the magnetic lines of force. When the magnetic lines cut across the vane, currents are induced, which gives rise to the  $I^2R$  losses. Very little RF energy reaches the metallic end of the waveguide, so there is little reflected energy and a low VSWR.

There are situations where it isn’t desirable to terminate the waveguide in a dummy load. Several reflective terminations are shown in Fig. 19-12. Perhaps the simplest form is the permanent end plate shown in Fig. 19-12A. The metal cover must be welded or otherwise affixed through a very-low-resistance joint. At the substantial power levels typically handled in transmitter waveguides, even small resistances can be important.

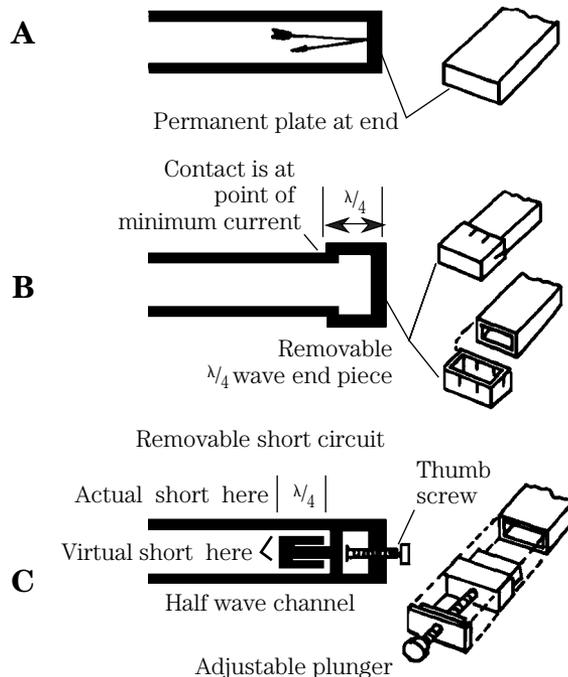
The end plate (shown in Fig. 19-12B) uses a quarter-wavelength cup to reduce the effect of joint resistances. The cup places the contact joint at a point that is a



19-11 Dummy loads: (A) sand-graphite chamber; (B) resistive rod; (C) wedge.



19-11 Dummy loads: (D) double wedge; (E) pyramid wedge; (F) vane dummy load.



**19-12**  
End terminations.

quarter-wavelength from the end. This point is a minimum-current node, so  $I^2R$  losses in the contact resistance become less important.

The adjustable short circuit is shown in Fig. 19-12C. The walls of the waveguide and the surface of the plunger form a half-wavelength channel. Because the metallic end of the channel is a short circuit, the impedance reflected back to the front of the plunger is zero ohms, or nearly so. Thus, a virtual short exists at the points shown. By this means, the contact (or joint) resistance problem is overcome.

## Waveguide joints and bends

Joints and bends in any form of transmission line or waveguide are seen as impedance discontinuities, and so are points at which disruptions occur. Thus, improperly formed bends and joints are substantial contributors to a poor VSWR. In general, bends, twists, joints, or abrupt changes in waveguide dimension can deteriorate the VSWR by giving rise to reflections.

Extensive runs of waveguide are sometimes difficult to make in a straight line. Although some installations do permit a straight waveguide, many others require directional change. This possibility is especially likely on shipboard installations. Figure 19-13A shows the proper ways to bend a waveguide around a corner. In each case, the radius of the bend must be at least two wavelengths at the lowest frequency that will be propagated in the system.

The “twist” shown in Fig. 19-13B is used to rotate the polarity of the E and H fields by  $90^\circ$ . This type of section is sometimes used in antenna arrays for phasing

the elements. As in the case of the bend, the twist must be made over a distance of at least two wavelengths.

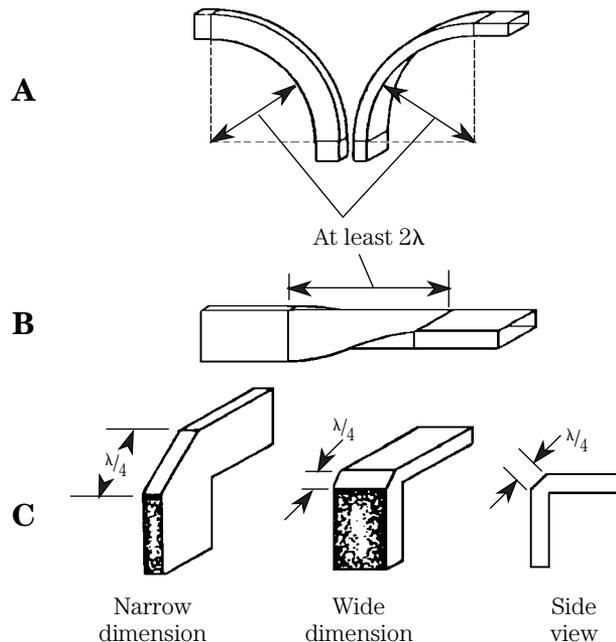
When an abrupt  $90^\circ$  transition is needed, it is better to use two successive  $45^\circ$  bends spaced one-quarter wavelength apart (see Fig. 19-13C). The theory (behind this kind of bend) is to cause the interference of the direct reflection of one bend, with the inverted reflection of the other. The resultant relationship between the fields is reconstructed as if no reflections had taken place.

Joints are necessary in practical waveguides because it simply isn't possible to construct a single length of practical size for all situations. Three types of common joints are used: *permanent*, *semipermanent*, and *rotating*.

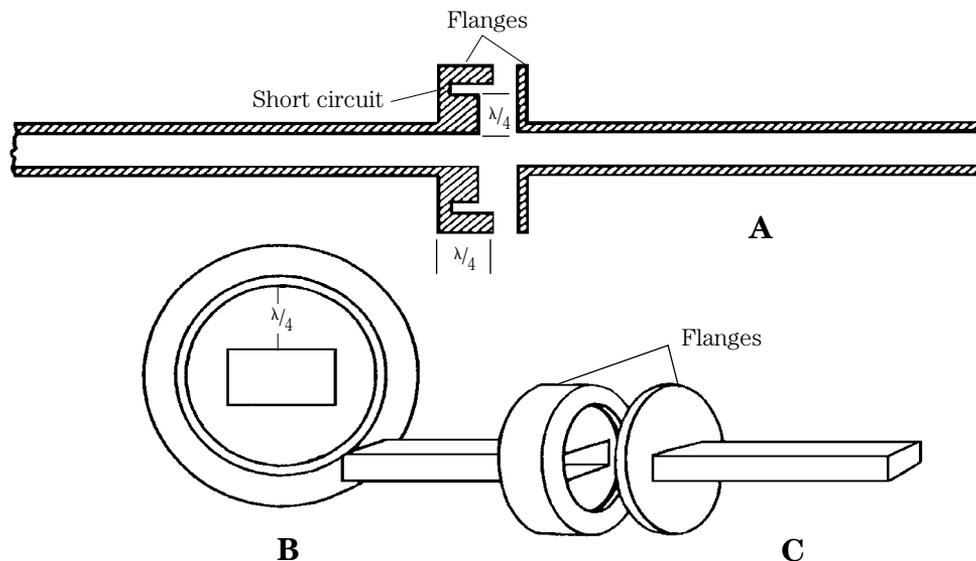
To make a permanent joint the two waveguide ends must be machined extremely flat so that they can be butt-fitted together. A welded or brazed seam bonds the two sections together. Because such a surface represents a tremendous discontinuity, reflections and VSWR will result unless the interior surfaces are milled flat and then polished to a mirror-like finish.

A *semipermanent joint* allows the joint to be disassembled for repair and maintenance, as well as allowing easier on-site assembly. The most common example of this class is the *choke joint* shown in Fig. 19-14.

One surface of the choke joint is machined flat and is a simple butt-end planar flange. The other surface is the mate to the planar flange, but it has a quarter-wavelength circular slot cut at a distance of one-quarter wavelength from the waveguide aperture. The two flanges are shown in side view in Fig. 19-14A, and the slotted end view is shown in Fig. 19-14B. The method for fitting the two ends together is shown in the oblique view in Fig. 19-14C.



**19-13** Bends in waveguide must be gentle.



**19-14** Choke joint.

*Rotating joints* are used in cases where the antenna has to point in different directions at different times. Perhaps the most common example of such an application is the radar antenna.

The simplest form of rotating joint is shown in Fig. 19-15. The key to its operation is that the selected mode is symmetrical about the rotating axis. For this reason, a circular waveguide operating in the  $TM_{01}$  mode is selected. In this rotating choke joint, the actual waveguide rotates but the internal fields do not (thereby minimizing reflections). Because most waveguide is rectangular, however, a somewhat more complex system is needed. Figure 19-16 shows a rotating joint consisting of two circular waveguide sections inserted between segments of rectangular waveguide. On each end of the joint, there is a rectangular-to-circular transition section.

In Fig. 19-16, the rectangular input waveguide operates in the  $TE_{10}$  mode that is most efficient for rectangular waveguide. The E-field lines of force couple with the circular segment, thereby setting up a  $TM_{01}$  mode wave. The  $TM_{01}$  mode has the required symmetry to permit coupling across the junction, where it meets another transition zone and is reconverted to  $TE_{10}$  mode.

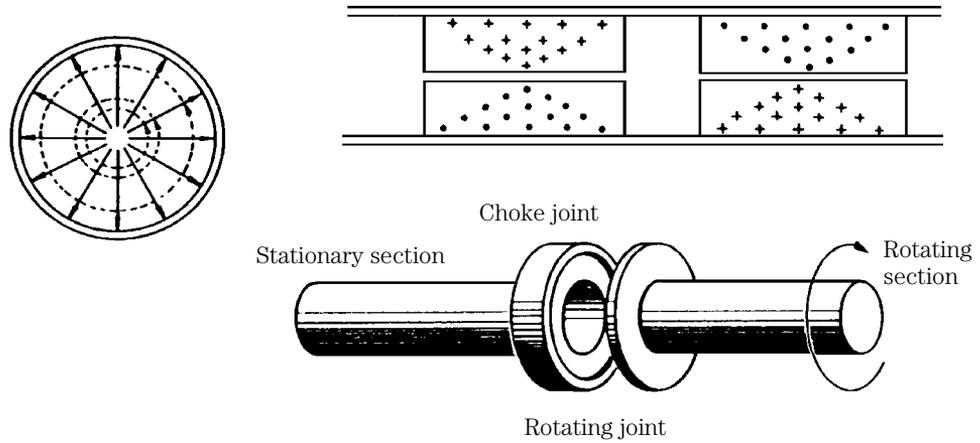
## Waveguide coupling methods

Except possibly for the case where an oscillator exists inside a waveguide, it is necessary to have some form of input or output coupling in a waveguide system. There are three basic types of coupling used in a microwave waveguide: *capacitive* (or *probe*), *inductive* (or *loop*), and *aperture* (or *slot*).

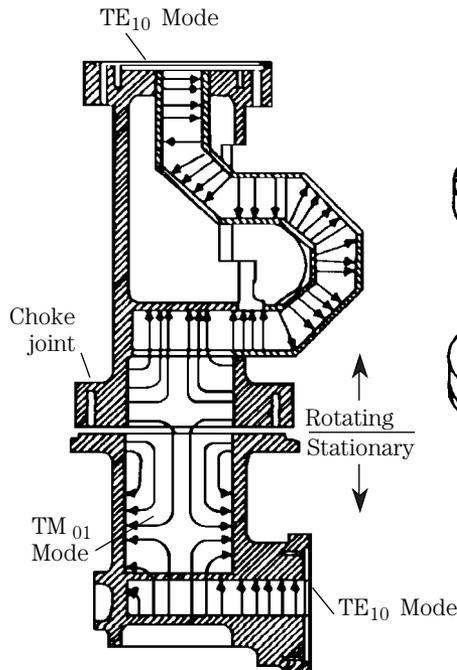
Capacitive coupling is shown in Fig. 19-17. This type of coupling uses a vertical radiator inserted into one end of the waveguide. Typically, the probe is a quarter-

wavelength in a fixed-frequency system. The probe is analogous to the vertical antennas used at lower frequencies. A characteristic of this type of radiator is that the E field is parallel to the waveguide top and bottom surfaces. This arrangement satisfies the first boundary condition for the dominant  $TE_{10}$  mode.

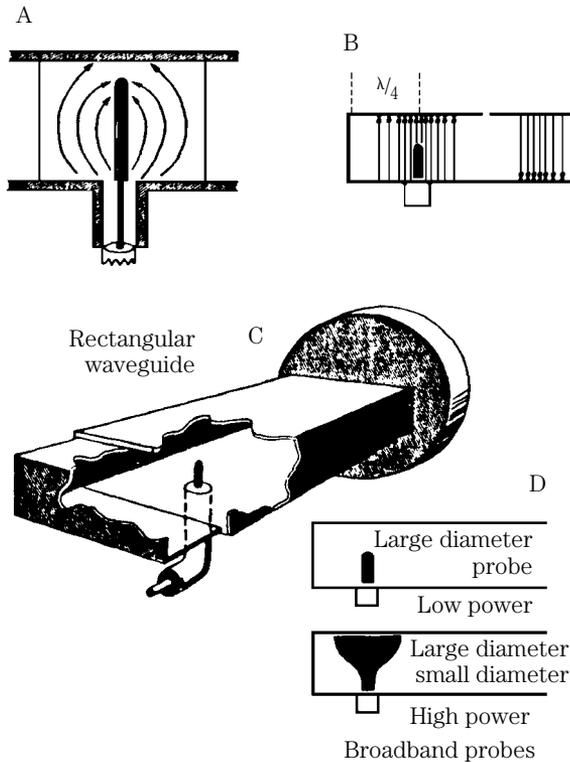
The radiator is placed at a point that is a quarter-wavelength from the rear wall (Fig. 19-17B). By traversing the quarter-wave distance ( $90^\circ$  phase shift), being reflected from the rear wall ( $180^\circ$  phase shift), and then retraversing the



19-15 Basic rotating joint.



19-16 Representative practical rotating joint.



**19-17**  
Probe (capacitive) coupling.

quarter-wavelength distance (another  $90^\circ$  phase shift), the wave undergoes a total phase shift of one complete cycle, or  $360^\circ$ . Thus, the reflected wave arrives back at the radiator in phase to reinforce the outgoing wave. Hence, none of the excitation energy is lost.

Some waveguides have an adjustable end cap (Fig. 19-17C) in order to accommodate multiple frequencies. The end cap position is varied to accommodate the different wavelength signals.

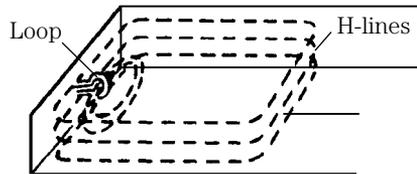
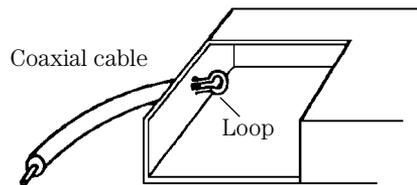
Figure 19-17D shows high- and low-power broadband probes that are typically not a quarter-wavelength except at one particular frequency. Broadbanding is accomplished by attention to the diameter-to-length ratio. The *degree of coupling* can be varied in any of several ways: the *length* of the probe can be varied; the *position* of the probe in the E field can be changed; or *shielding* can be used to partially shade the radiator element.

Inductive, or loop coupling, is shown in Fig. 19-18. A small loop of wire (or other conductor) is placed such that the number of magnetic flux lines is maximized. This form of coupling is popular on microwave receiver antennas, in order to make a waveguide-to-coaxial cable transition. In some cases, the loop is formed by the pig-tail lead of a detector diode that, when combined with a local oscillator, downconverts the microwave signal to an IF frequency in the 30- to 300-MHz region.

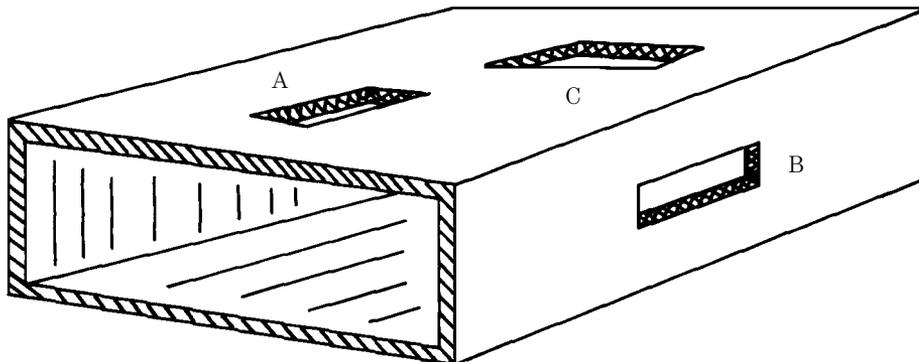
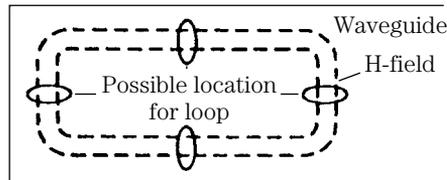
Aperture, or slot coupling, is shown in Fig. 19-19. This type of coupling is used to couple together two sections of waveguide, as on an antenna feed system. Slots

can be designed to couple either electric, magnetic, or electromagnetic fields. In Fig. 19-19, slot *A* is placed at a point where the *E* field peaks, so it allows electrical field coupling. Similarly, slot *B* is at a point where the *H* field peaks, so it allows magnetic field coupling. Finally, we see slot *C*, which allows electromagnetic field coupling.

Slots can also be characterized according to whether they are *radiating* or *non-radiating*. A nonradiating slot is cut at a point that does not interrupt the flow of currents in the waveguide walls. The radiating slot, on the other hand, does interrupt currents flowing in the walls. A radiating slot is the basis for several forms of antenna, which are discussed at the end of this chapter.



**19-18**  
Loop (inductive) coupling.



**19-19** Slot coupling.

## Microwave antennas

Antennas are used in communications and radar systems at frequencies from the very lowest to the very highest. In both theory and practice, antennas are used until frequencies reach infrared and visible light, at which point optics becomes more important. Microwaves are a transition region between ordinary “radio waves” and “optical waves,” so (as might be expected) microwave technology makes use of techniques from both worlds. For example, both dipoles and parabolic reflectors are used in microwave systems.

The purpose of an antenna is to act as a *transducer* between either electrical oscillations or propagated “guided waves” (i.e., in transmission lines or waveguides) and a propagating electromagnetic wave in free space. A principal function of the antenna is to act as an *impedance matcher* between the waveguide, or transmission line, impedance and the impedance of free space.

Antennas can be used equally well for both receiving and transmitting signals because they obey the *law of reciprocity*. That is, the same antenna can be used to both receive and transmit with equal success. Although there might be practical or mechanical reasons to prefer specific antennas for one or the other mode, electrically they are the same.

In the transmit mode, the antenna must radiate electromagnetic energy. For this job, the important property is *gain*  $G$ . In the receive mode, the job of the antenna is to gather energy from impinging electromagnetic waves in free space. The important property for receiver antennas is the *effective aperture*  $A_e$ , which is a function of the antenna’s physical area. Because of reciprocity, a large gain usually infers a large effective aperture and vice versa. *Effective aperture* is defined as the area of the impinging radio wavefront that contains the same power as is delivered to a matched resistive load across the feedpoint terminals.

## The isotropic “antenna”

Antenna definitions and specifications can become useless unless a means is provided for putting everything on a common footing. Although a variety of systems exist for describing antenna behavior, the most common system compares a specific antenna with a theoretical construct, called the *isotropic radiator*.

An isotropic radiator is a spherical point source that radiates equally well in all directions. By definition, the directivity of the isotropic antenna is unity (1), and all antenna gains are measured against this standard. Because the geometry of the sphere and the physics of radiation are well known, we can calculate field strength and power density at any point. These figures can then be compared with the actual values from an antenna being tested. From spherical geometry, we can calculate isotropic power density at any distance  $R$  from the point source:

$$P_d = \frac{P}{4\pi R^2} \quad [19.19]$$

where

$P_d$  is the power density, in watts per square meter

$P$  is the power in watts input to the isotropic radiator

$R$  is the radius in meters at which point power density is measured

**Example 19-5** Calculate the power density at a distance of 1 km (1000 m) from a 1000-W isotropic source.

**Solution:**

$$\begin{aligned} P_d &= \frac{P}{4\pi R^2} \\ &= \frac{(1000 \text{ W})}{4\pi (1000 \text{ m})^2} \\ &= 7.95 \times 10^{-5} \text{ W/m}^2 \end{aligned}$$

The rest of this chapter covers antenna gains and directivities that are relative to isotropic radiators.

### Near field and far field

Antennas are defined in terms of *gain* and *directivity*, both of which are measured by looking at the radiated field of the antenna. There are two fields to consider: *near field* and *far field*. The patterns published for an antenna tend to reflect far-field performance. The far field for most antennas falls off according to the *inverse square law*. That is, the intensity falls off according to the square of the distance ( $1/R^2$ ), as in Eq. 19.19.

The near field of the antenna contains more energy than the far field because of the electric and magnetic fields close to the antenna radiator element. The near field tends to diminish rapidly according to a  $1/R^4$  function. The minimum distance to the edge of the near field is a function of both the wavelength of the radiated signals and the antenna dimensions:

$$r_{\min} = \frac{2d^2}{\lambda} \quad [19.20]$$

where

$r_{\min}$  is the near-field distance

$d$  is the largest antenna dimension

$\lambda$  is the wavelength of the radiated signal (all factors in the same units)

**Example 19-6** An antenna with a length of 6 cm radiates a 12-cm wavelength signal. Calculate the near-field distance.

**Solution:**

$$\begin{aligned}
 r_{\min} &= \frac{2d^2}{\lambda} \\
 &= \frac{(2)(6 \text{ cm})^2}{12 \text{ cm}} \\
 &= \frac{72}{12} = 6 \text{ cm}
 \end{aligned}$$

**Antenna impedance**

Impedance represents the total opposition to the flow of alternating current (e.g., RF), and includes both resistive and reactive components. The reactive components can be either capacitive or inductive, or a combination of both. Impedance can be expressed in either of two notations:

$$Z = \sqrt{R^2 + (X_L - X_C)^2} \quad [19.21]$$

or

$$Z = R \pm jX \quad [19.22]$$

Of these, Eq. 19.22 is perhaps the more commonly used in RF applications. The reactive part of antenna impedance results from the magnetic and electrical fields (close to the radiator) returning energy to the antenna radiator during every cycle. The resistive part of impedance consists of two elements: *ohmic losses*  $R_o$  and *radiation resistance*  $R_r$ . The ohmic losses are due to heating of the antenna conductor elements by RF current passing through, as when current passes through any conductor.

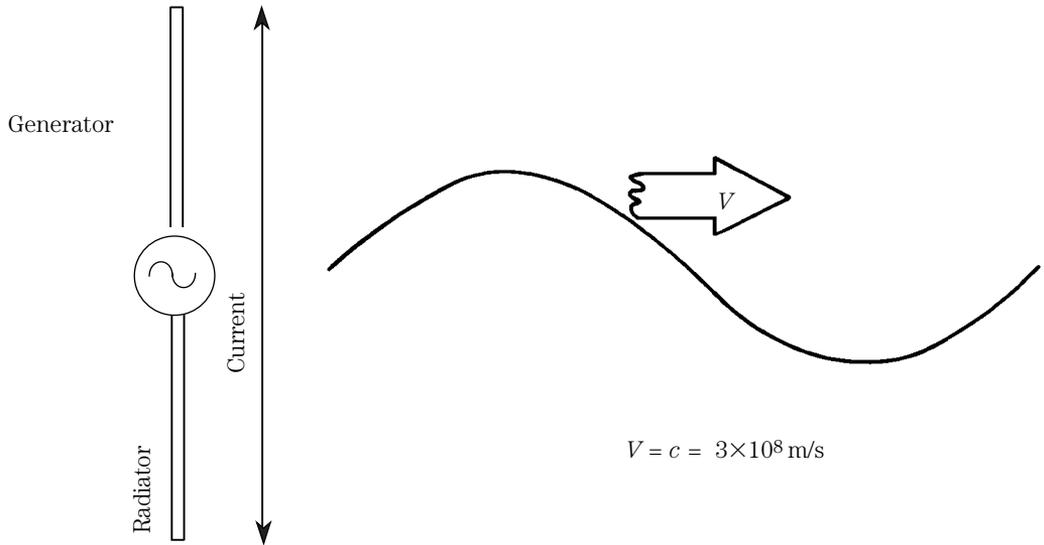
The radiation resistance relates to the radiated energy. An efficiency factor  $k$  compares the loss and radiation resistances:

$$k = \frac{R_r}{R_r + R_o} \quad [19.23]$$

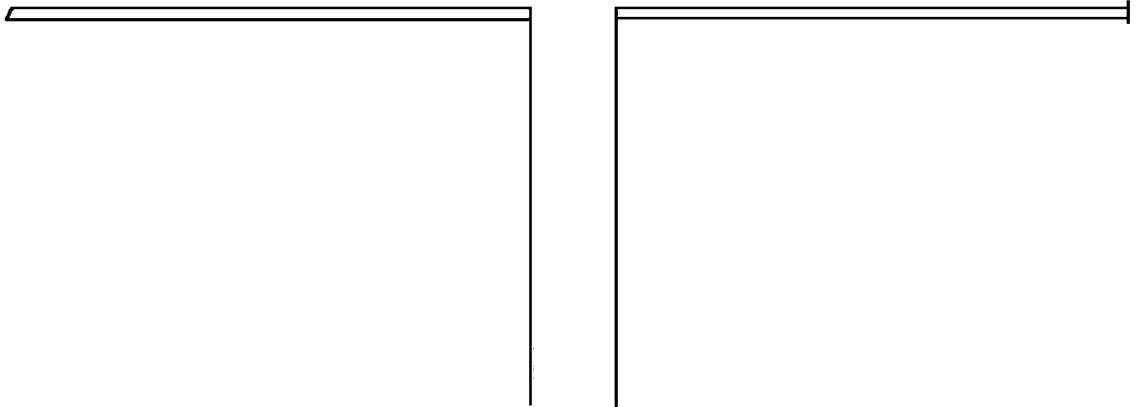
The goal of the antenna designer is to reduce  $R_o$  to a minimum. The value of  $R_r$  is set by the antenna design and installation, and is defined as the quotient of the voltage over the current at the feedpoint, less losses.

**Dipole antenna elements**

The *dipole* is a two-pole antenna (Fig. 19-20) that can be modeled as either a single radiator fed at the center (Fig. 19-20A) or a pair of radiators fed back to back (Fig. 19-20B). RF current from the source oscillates back and forth in the radiator element, causing an electromagnetic wave to propagate in a direction *perpendicular* to the radiator element. The polarity of any electromagnetic field is the direction of the electrical field vector (see Chap. 2). In the dipole, the polarization is parallel to the radiator element: a horizontal element produces a horizontally polarized signal, while a vertical element produces a vertically polarized signal.

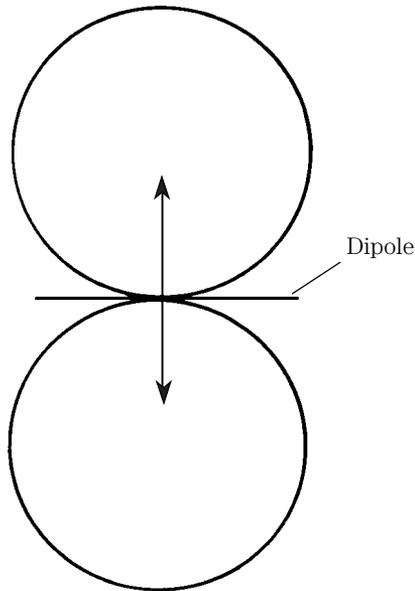


**19-20A** Basic dipole antenna showing propagation.

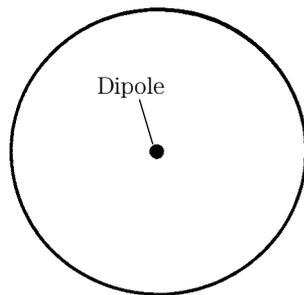


**19-20B** Basic dipole antenna.

Figure 19-21 shows the radiator patterns for the dipole viewed from two perspectives. Figure 19-21A shows the pattern of a horizontal half-wavelength dipole as viewed from above. This plot shows the *directivity* of the dipole: maximum radiation is found in two lobes perpendicular to the radiator length. The plot in Fig. 19-21B shows the end-on pattern of the dipole. This omnidirectional pattern serves for a vertically polarized dipole viewed from above. The end-on pattern of a horizontal dipole would be similar, except that it is distorted by ground effects unless the antenna is a very large number of wavelengths above the ground.



**19-21A**  
Dipole pattern (horizontal).



**19-21B**  
Dipole pattern (vertical).

A microwave dipole is shown in Fig. 19-22. The antenna radiator element consists of a short conductor at the end of a section of waveguide. Although most low-frequency dipoles are a half-wavelength, microwave dipoles might be either a half-wavelength, less than a half-wavelength, or greater than a half-wavelength, depending upon application. For example, because most microwave dipoles are used to illuminate a reflector antenna of some sort, the length of the dipole depends upon the exact illumination function required for proper operation of the reflector. Most, however, will be a half-wavelength.

### **Antenna directivity and gain**

The dipole discussed illustrated a fundamental property of the type of antenna generally used at microwave frequencies: *directivity* and *gain*. These two concepts are different but so interrelated that they are usually discussed at the same time. Because of the directivity, the antenna focuses energy in only two directions, which

means that all of the energy is found in those directions (Fig. 19-21A), rather than being distributed over a spherical surface. Thus, the dipole has a gain approximately 2.1 dB greater than isotropic. In other words, the measured power density at any point will be 2.1 dB higher than the calculated isotropic power density for the same RF input power to the antenna.

**Directivity** The directivity of an antenna is a measure of its ability to direct RF energy in a limited direction, rather than in all (spherical) directions equally. As shown in Fig. 19-21A, the horizontal directivity of the dipole forms a bidirectional figure-8 pattern. Two methods for showing unidirectional antenna patterns are shown in Fig. 19-23. The method of Fig. 19-23A is a polar plot viewed from above. The main lobe is centered on 0°. The plot of Fig. 19-23B is a rectangular method for displaying the same information. This pattern follows a  $(\sin x)/x$  function or, for power,  $[(\sin x)/x]^2$ .

Directivity  $D$  is a measure of relative power densities:

$$D = \frac{P_{\max}}{P_{\text{av}}} \quad [19.24]$$

Or, referenced to isotropic,

$$D = \frac{4 \pi}{\Phi} \quad [19.25]$$

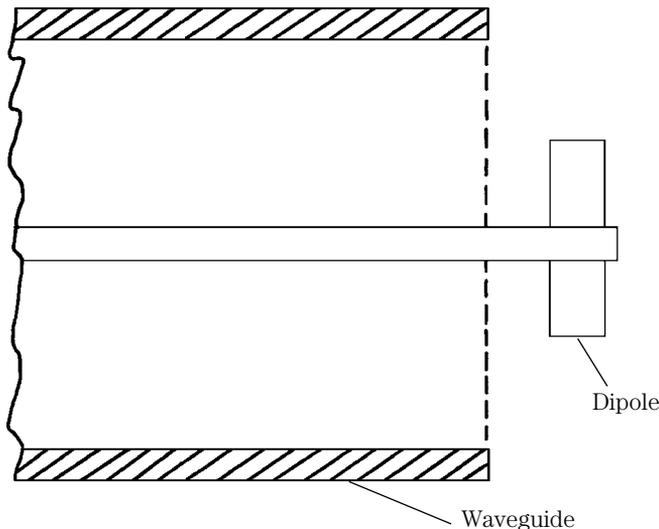
where

$D$  is the directivity

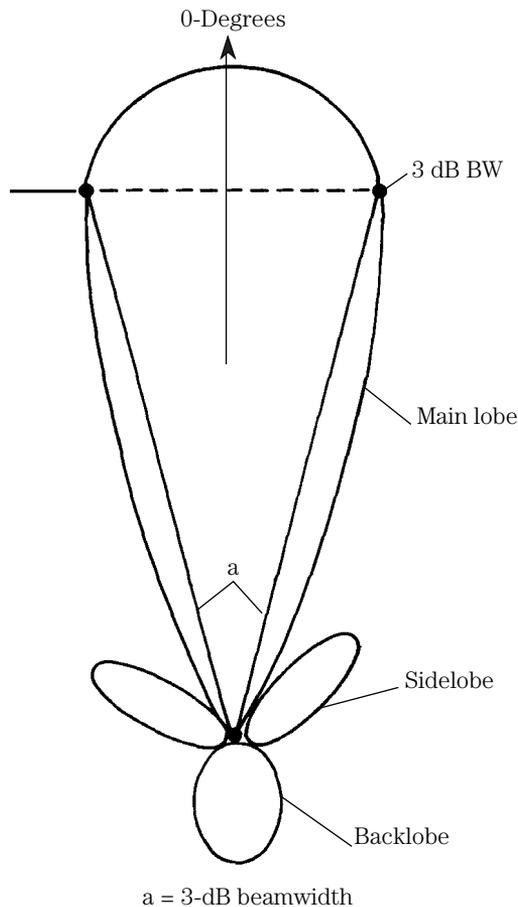
$P_{\max}$  is the maximum power

$P_{\text{av}}$  is the average power

$\Phi$  is the solid angle subtended by the main lobe



**19-22**  
Microwave dipole radiator.



**19-23A**  
Directional antenna pattern (top view).

The term  $\Phi$  is a solid angle, which emphasizes the fact that antenna patterns must be examined in at least two extents: horizontal and vertical.

A common method for specifying antenna directivity is *beamwidth* (BW). The definition of BW is the angular displacement between points on the main lobe (see Figs. 19-23A and 19-23B), where the power density drops to one-half ( $-3$  dB) of its maximum main lobe power density. This angle is shown in Fig. 19-23A as  $a$ .

In an ideal antenna system, 100 percent of the radiated power is in the main lobe, and there are no other lobes. But in real antennas certain design and installation anomalies cause additional minor lobes, such as the *sidelobes* and *backlobe* shown in Fig. 19-23A. Several problems derive from the minor lobes. First is the loss of usable power. For a given power density required at a distant receiver site, the transmitter must supply whatever additional power is needed to make up for the minor lobe losses.

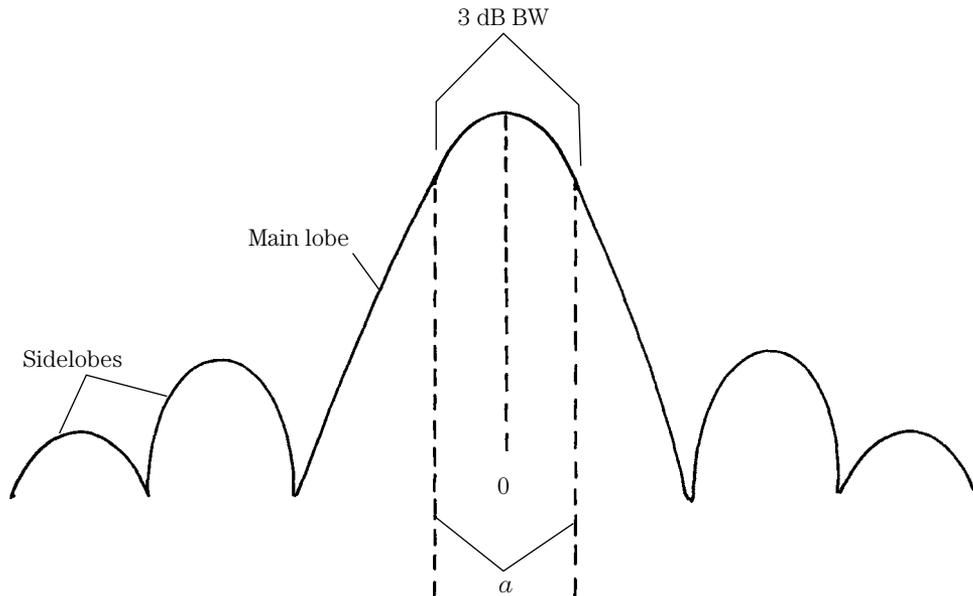
The second problem is intersystem interference. A major application of directional antennas is the prevention of mutual interference between nearby cochan-

nel stations. In radar systems, high sidelobes translate to errors in detected targets. If, for example, a sidelobe is large enough to detect a target, then the radar display will show this off-axis target as if it was in the main lobe of the antenna. The result is an azimuth error that could be important in terms of marine and aeronautical navigation.

**Gain** Antenna gain derives from the fact that energy is squeezed into a limited space instead of being distributed over a spherical surface. The term *gain* implies that the antenna creates a higher power when, in fact, it merely concentrates the power into a single direction that would otherwise be spread out over a larger area. Even so, it is possible to speak of an apparent increase in power. Antenna-transmitter systems are often rated in terms of *effective radiated power* (ERP). The ERP is the product of the transmitter power and the antenna gain. For example, if an antenna has a gain of +3 dB, the ERP will be twice the transmitter output power. In other words, a 100-W output transmitter connected to a +3-dB antenna will produce a power density at a distant receiver equal to a 200-W transmitter feeding an isotropic radiator. There are two interrelated gains to be considered: *directivity gain*  $G_d$  and *power gain*  $G_p$ .

The directivity gain is defined as the quotient of the maximum radiation intensity over the average radiation intensity (note the similarity to the directivity definition). This measure of gain is based on the shape of the antenna radiation pattern, and can be calculated with respect to an isotropic radiator ( $D=1$ ) from:

$$G_d = \frac{4\pi P_a}{P_r} \quad [19.26]$$



19-23B Graphically presented pattern.

where

$G_d$  is the directivity gain

$P_a$  is the maximum power radiated per unit of solid angle

$P_r$  is the total power radiated by the antenna

The power gain is similar in nature, but slightly different from directivity gain; it includes dissipative losses in the antenna. Not included in the power gain are losses caused by cross-polarization or impedance mismatch between the waveguide (or transmission line) and the antenna. There are two commonly used means for determining power gain:

$$G_p = \frac{4\pi P_a}{P_n} \quad [19.27]$$

and

$$G_p = \frac{P_{ai}}{P_i} \quad [19.28]$$

where

$P_a$  is the maximum radiated per unit solid angle

$P_n$  is the net power accepted by the antenna (i.e., less mismatch losses)

$P_{ai}$  is the average intensity at a distant point

$P_i$  is the intensity at the same point from an isotropic radiator fed the same RF power level as the antenna

(Equations assume equal power to antenna and comparison isotropic source.)

Provided that ohmic losses are kept negligible, the relationship between directivity gain and power gain is given by:

$$G_p = \frac{P_r G_d}{P_n} \quad [19.29]$$

(All terms as previously defined.)

**Relationship of gain and aperture** Antennas obey the law of reciprocity, which means that any given antenna will work as well to receive as to transmit. The function of the receiver antenna is to gather energy from the electromagnetic field radiated by the transmitter antenna. The aperture is related to, and often closely approximates, the physical area of the antenna. But in some designs the effective aperture  $A_e$  is less than the physical area  $A$ , so there is an effectiveness factor  $n$  that must be applied. In general, however, a high-gain transmitter antenna also exhibits a high receiving aperture, and the relationship can be expressed as

$$G = \frac{4\pi A_e n}{\lambda^2} \quad [19.30]$$

where

$A_e$  is the effective aperture

$n$  is the aperture effectiveness ( $n=1$  for a perfect, lossless antenna)

$\lambda$  is the wavelength of the signal

### Horn antenna radiators

The horn radiator is a tapered termination of a length of waveguide (see Fig. 19-24 (A-C) that provides the impedance transformation between the waveguide impedance and the free-space impedance. Horn radiators are used both as antennas in their own right, and as illuminators for reflector antennas. Horn antennas are not a perfect match to the waveguide, although standing wave ratios of 1.5:1 or less are achievable. The gain of a horn radiator is proportional to the area  $A$  of the flared open flange ( $A = ab$  in Fig. 19-24B), and inversely proportional to the square of the wavelength:

$$G = \frac{10A}{\lambda^2} \quad [19.31]$$

where

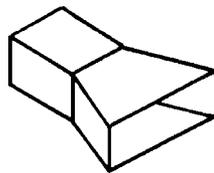
$A$  is the flange area

$\lambda$  is the wavelength (both in same units)

The  $-3$ -dB beamwidth for vertical and horizontal extents can be approximated from:

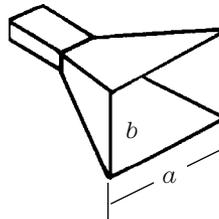
*Vertical*

$$\Phi_v = \frac{51\lambda}{b} \quad \text{degrees} \quad [19.32]$$



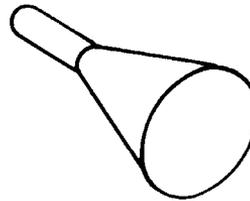
Rectangular

**A**



Pyramidal

**B**



Conical

**C**

19-24 Horn radiators.

*Horizontal*

$$\Phi_h = \frac{70\lambda}{a} \quad \text{degrees} \quad [19.33]$$

where

- $\Phi_v$  is the vertical beamwidth, in degrees
- $\Phi_h$  is the horizontal beamwidth, in degrees
- $a, b$  are dimensions of the flared flange
- $\lambda$  is the wavelength

A form of antenna, related to the horn, is the cavity antenna of Fig. 19-25. In this type of antenna, a quarter-wavelength radiating element extends from the waveguide (or transmission line connector) into a resonant cavity. The radiator element is placed a quarter-wavelength into a resonant cavity; and is spaced a quarter-wavelength from the rear wall of the cavity. A tuning disk is used to alter cavity dimensions in order to provide a limited tuning range for the antenna. Gains to about 6 dB are possible with this arrangement.

### Reflector antennas

At microwave frequencies, it becomes possible to use *reflector antennas* because of the short wavelengths involved. Reflectors are theoretically possible at lower frequencies, but because of the longer wavelengths, the antennas would be so large that they become impractical. Several forms of reflector are used (Figs. 19-26 and 19-27). In Fig. 19-26 we see the *corner reflector* antenna, which is used primarily in the high-UHF and low-microwave region. A dipole element is placed at the “focal point” of the corner reflector, so it receives (in phase) the reflected wavefronts from the surface. Either solid metallic reflector surfaces or wire mesh may be used. When mesh is used, however, the holes in the mesh must be  $\frac{1}{2}$ -wavelength or smaller.

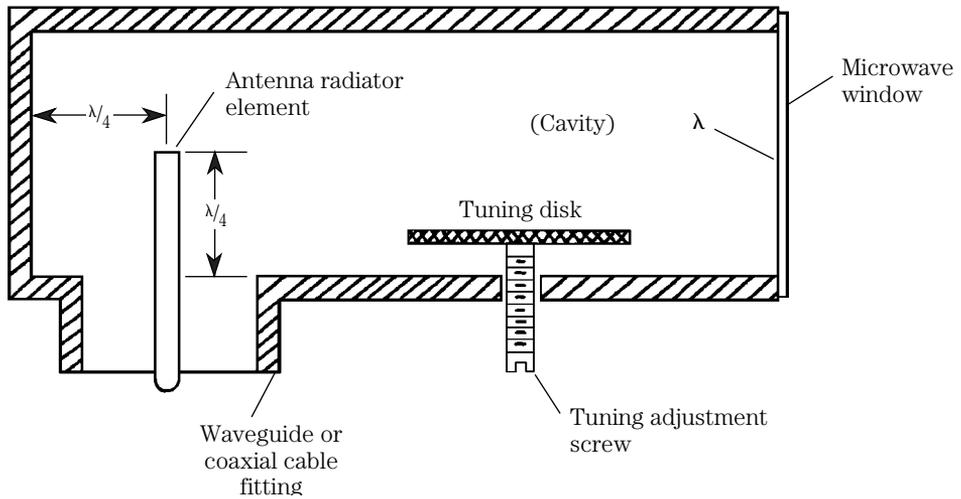
Figure 19-27 shows several other forms of reflector surface shape, most of which are used in assorted radar applications.

### Parabolic “dish” antennas

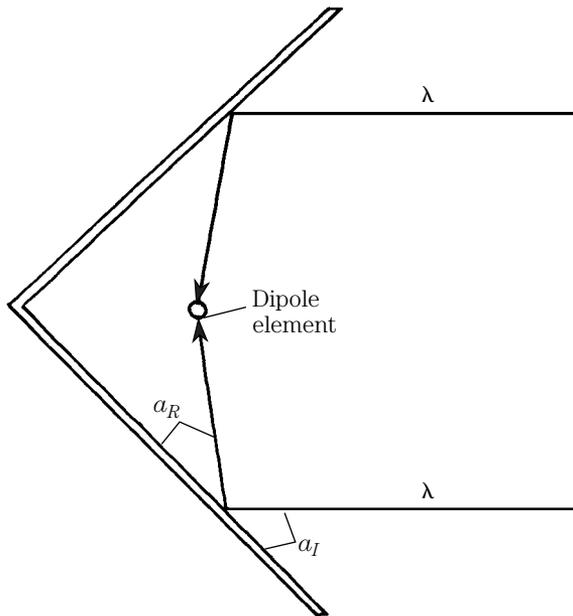
The parabolic reflector antenna is one of the most widespread of all the microwave antennas, and is the type that normally comes to mind when thinking of microwave systems. This type of antenna derives its operation from physics similar to optics, and is possible because microwaves are in a transition region between ordinary radio waves and infrared/visible light.

The dish antenna has a paraboloid shape as defined by Fig. 19-28. In this figure, the dish surface is positioned such that the center is at the origin (0,0) of an  $x$ - $y$  coordinate system. For purposes of defining the surface, we place a second vertical axis called the *directrix* ( $y'$ ) a distance behind the surface equal to the focal length ( $u$ ). The paraboloid surface follows the function  $y^2 = 4ux$ , and has the property that a line from the focal point  $F$  to any point on the surface is the same length as a line from that same point to the directrix (in other words,  $MN = MF$ ).

If a radiator element is placed at the focal point  $F$ , then it will illuminate the reflector surface, causing wavefronts to be propagated away from the surface in



19-25 Cavity antenna.



19-26 Corner reflector.

phase. Similarly, wavefronts, intercepted by the reflector surface, are reflected to the focal point.

**Gain** The gain of a parabolic antenna is a function of several factors: dish diameter, feed illumination, and surface accuracy. The dish diameter  $D$  should be large compared with its depth. Surface accuracy refers to the degree of surface irregularities. For commercial antennas,  $\lambda/8$ -wavelength surface accuracy is usually sufficient,



A. Paraboloid



B. Truncated paraboloid (surface search)



C. Truncated paraboloid (height finding)



D. Orange-peel paraboloid



E. Cylindrical paraboloid



F. Corner reflector

19-27 Reflector antennas.

although on certain radar antennas the surface accuracy specification must be tighter.

The feed illumination refers to how evenly the feed element radiates to the reflector surface. For circular parabolic dishes, a circular waveguide feed produces optimum illumination, and rectangular waveguides are less than optimum. The  $TE_{11}$  mode is desired. For best performance, the illumination should drop off evenly from the center to the edge, with the edge being 210 dB down from the center. The diameter, length, and beamwidth of the radiator element (or horn) must be optimized for the specific  $F/d$  ratio of the dish. The cutoff frequency is approximated from

$$f_{\text{cutoff}} = \frac{175,698}{d_{\text{mm}}} \quad [19.34]$$

where

$f_{\text{cutoff}}$  is the cutoff frequency  
 $d$  is the inside diameter of the circular feedhorn

The gain of the parabolic dish antenna is found from

$$G = \frac{k (\pi D)^2}{\lambda^2} \quad [19.35]$$

where

$G$  is the gain over isotropic  
 $D$  is the diameter  
 $\lambda$  is the wavelength (same units as  $D$ )  
 $k$  is the reflection efficiency (0.4 to 0.7, with 0.55 being most common)

The  $-3$ -dB beamwidth of the parabolic dish antenna is approximated by

$$BW = \frac{70 \lambda}{D} \quad [19.36]$$

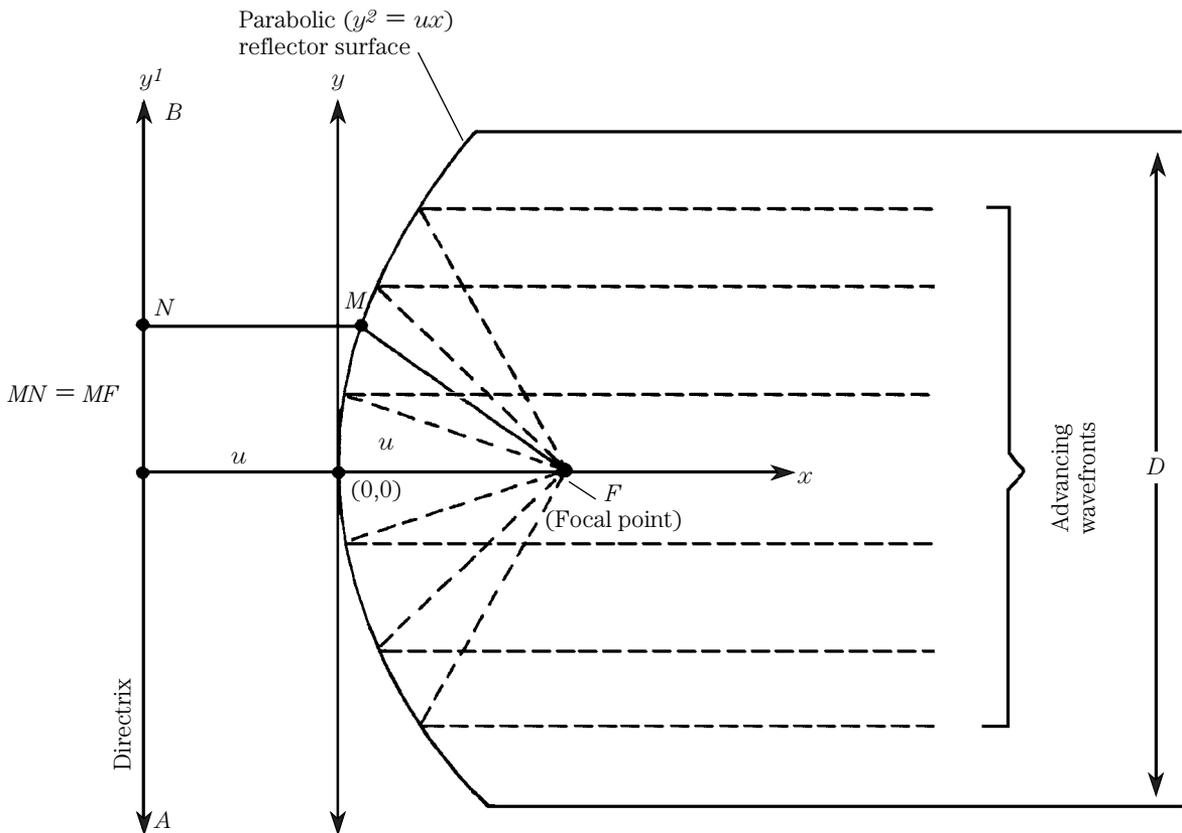
and the focal length by

$$F = \frac{D^2}{16d} \quad [19.37]$$

For receiving applications, the effective aperture is the relevant specification and is found from

$$A_e = k\pi(D/2)^2 \quad [19.38]$$

The antenna pattern radiated by the antenna is similar to Fig. 19-23B. With horn illumination, the sidelobes tend to be 23 to 28 dB below the main lobe, or 10 to 15 dB below isotropic. It is found that 50 percent of the energy radiated by the parabolic dish is within the  $-3$ -dB beamwidth, and 90 percent is between the first nulls on either side of the main lobe.

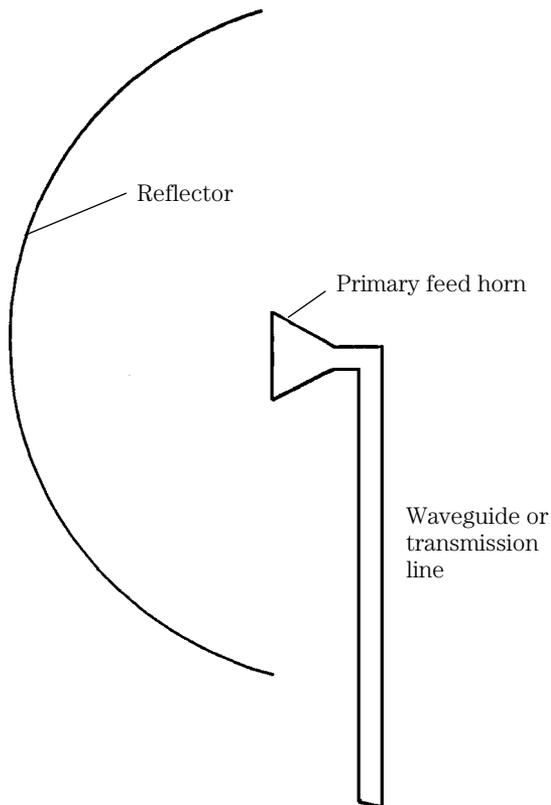


19-28 Ray tracing shows operation of parabolic antenna.

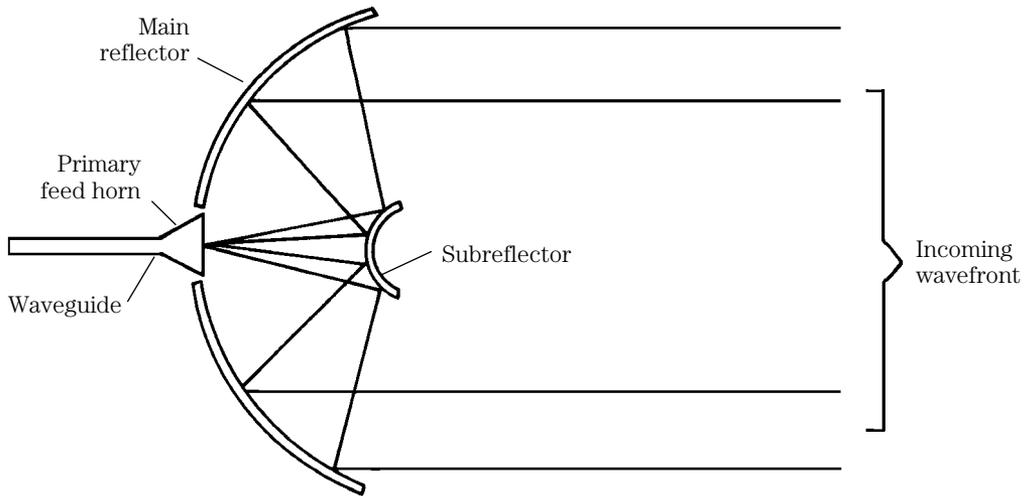
If a dipole element is used for the feed device, then a *splash plate* is placed  $\frac{1}{4}$ -wavelength behind the dipole in order to improve illumination. The splash plate must be several wavelengths in diameter, and is used to reflect the backlobe back toward the reflector surface. When added to the half-wave phase reversal inherent in the reflection process, the two-way quarter-wavelength adds another half-wavelength and thereby permits the backwave to move out in phase with the front lobe wave.

**Parabolic dish feed geometries** Figure 19-29 shows two methods for feeding parabolic dish antennas, regardless of which form of radiator (horn, dipole, etc.) is used. In Fig. 19-29A we see the method in which the radiator element is placed at the focal point, and a waveguide (or transmission line) is routed to it. This method is used in low-cost installations such as home satellite TV receive-only (TVRO) antennas.

Figure 19-29B shows the *Cassegrain feed* system. This system is modeled after the Cassegrain optical telescope. The radiator element is placed at an opening at the center of the dish. A hyperbolic subreflector is placed at the focal point, and it is used to reflect the wavefronts to the radiator element. The Cassegrain system results in lower-noise operation because of several factors: less transmission line length, lower sidelobes, and the fact that the open horn sees sky instead of earth



**19-29A**  
Parabolic antenna conventional feed.



**19-29B** Parabolic antenna Cassegrain feed.

(which has lower temperature); on the negative side, galactic and solar noise might be slightly higher on a Cassegrain dish.

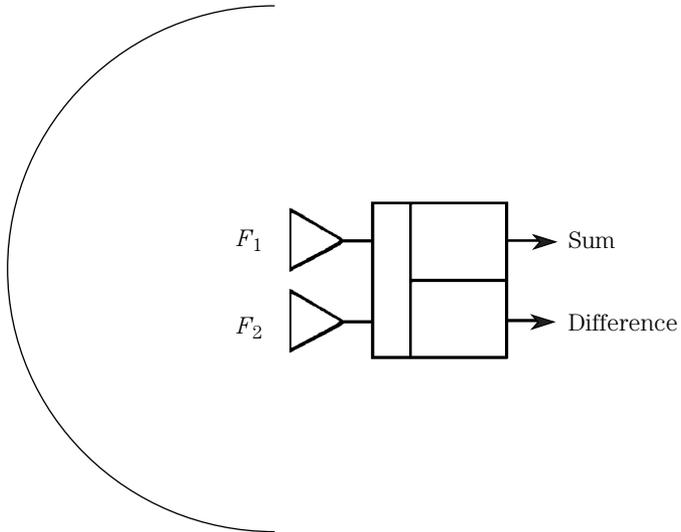
Figure 19-30A shows the *monopulse* feed geometry. In this system, a pair of radiator elements are placed at the focal point, and fed to a power splitter network that outputs both *sum* and *difference* signals (Fig. 19-30B). When these are combined, the resultant beam shape has improved  $-3$ -dB beamwidth as a result of the algebraic summation of the two.

### Array antennas

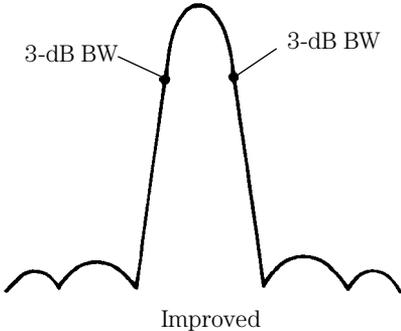
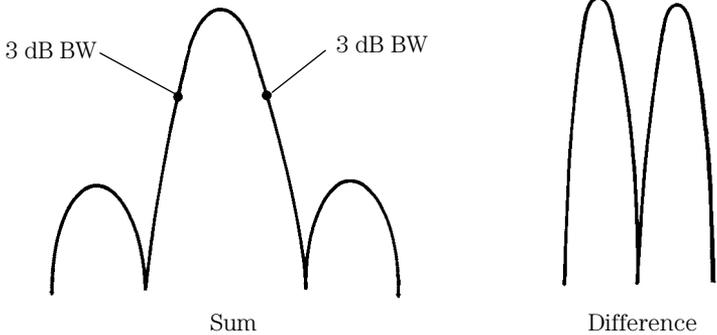
When antenna radiators are arranged in a precision array, an increase in gain occurs. An array might be a series of dipole elements, as in the broadside array of Fig. 19-31 (which is used in the UHF region), or a series of slots, horns, or other radiators. The overall gain of an array antenna is proportional to the number of elements, as well as the details of their spacing. In this, and other antennas, a method of *phase shifting* is needed. In Fig. 19-31, the phase shifting is caused by the crossed feeding of the elements, but in more modern arrays, other forms of phase shifter are used.

Two methods of feeding an array are shown in Fig. 19-32. The *corporate feed* method connects all elements, and their phase shifters, in parallel with the source. The *branch feed* method breaks the waveguide network into two (or more) separate paths.

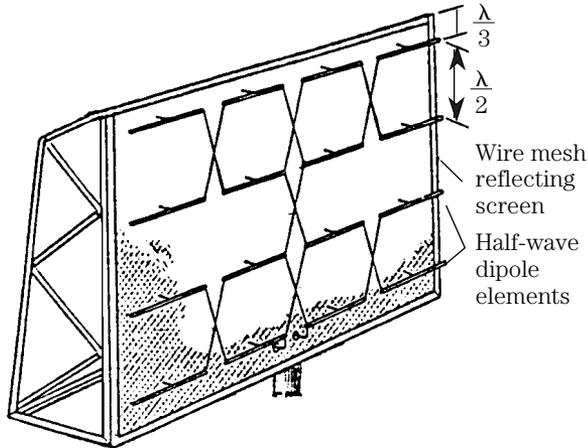
**Solid-state array antennas** Some modern radar sets use solid-state array antennas consisting of a large number of elements, each of which is capable of shifting the phase of a microwave input signal. Two forms are known: *passive* (Fig. 19-33A) and *active* (Fig. 19-33B). In the passive type of element, a ferrite (or PIN diode) phase shifter is placed in the transmission path between the RF input and the radiator element (usually a slot). By changing the phase of the RF signal selectively, it is possible to form and steer the beam at will. A 3-bit (i.e., three discrete state) phase



**19-30A**  
Monopulse feed.



**19-30B**  
Monopulse patterns.



**19-31**  
Reflector array antenna.

shifter allows the phase to shift in  $45^\circ$  increments, while a 4-bit phase shifter allows  $22.5^\circ$  increments of phase shift.

The active element contains, in addition to a phase shifter, a transmit power amplifier (1 or 2 W) and a low-noise amplifier (LNA) for receiving. A pair of transmit/receive (T/R) switches select the path to which the RF signal is directed. The total output power of this antenna is the sum of all output powers from all elements in the array. For example, an array of one thousand 2-W elements makes a 2000-W system.

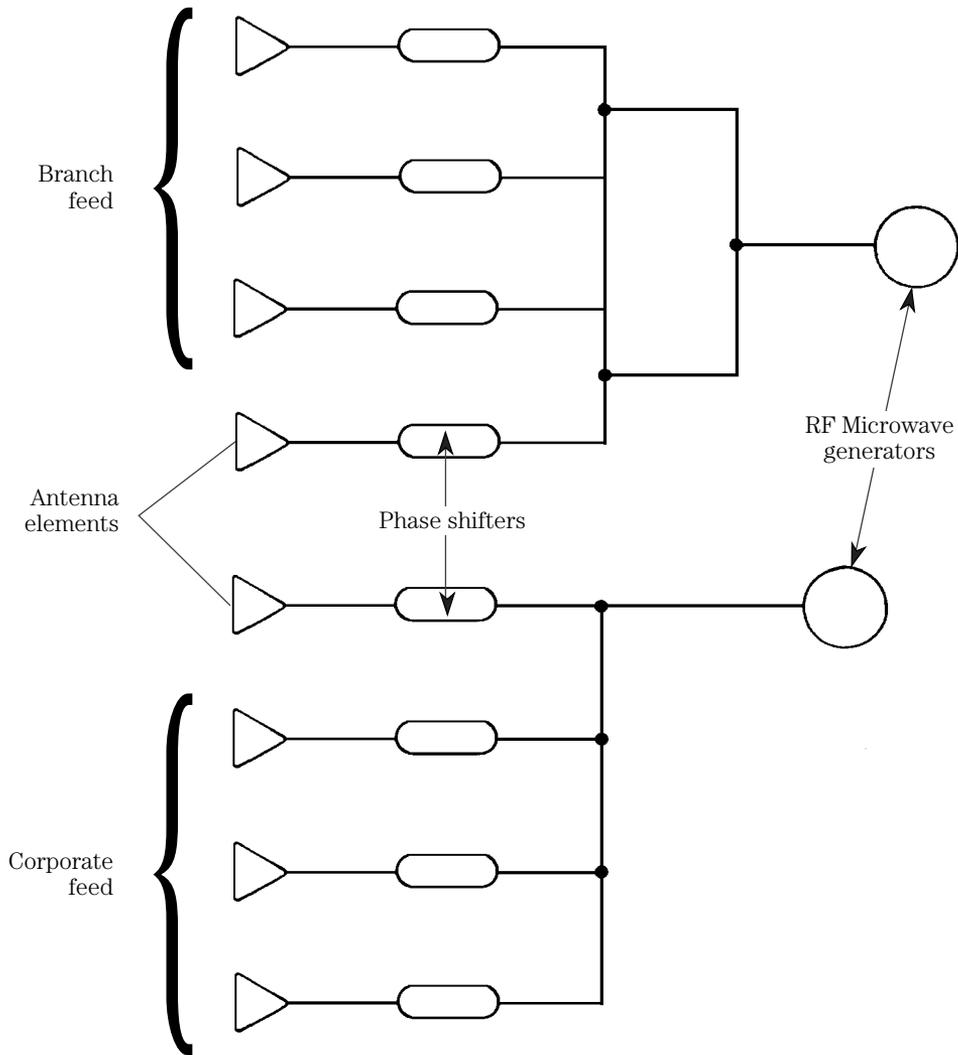
### Slot array antennas

A resonant slot (cut into a wall of a section of waveguide) is somewhat analogous, if not identical, to a dipole. By cutting several slots in the waveguide, we obtain the advantages of an array antenna in which the elements are several slot radiators. Slot array antennas are used for marine navigation radars, telemetry systems, and the reception of microwave television signals in the Multipoint Distribution Service (MDS) on 2.145 GHz.

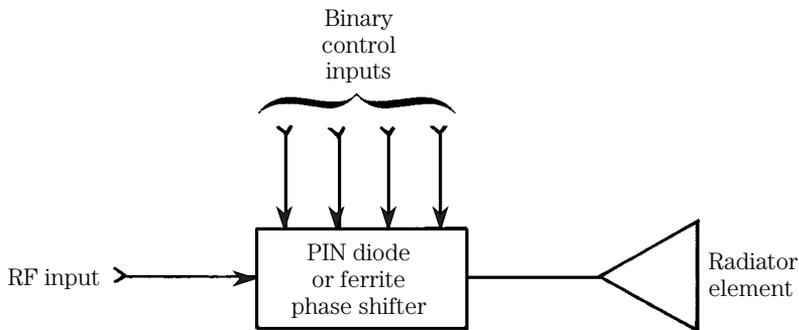
Figure 19-34 shows a simple slot antenna used in telemetry applications. A slotted section of rectangular waveguide is mounted to a right-angle waveguide flange. An internal wedge (not shown) is placed at the top of the waveguide and serves as a matching-impedance termination to prevent internal reflected waves. Directivity is enhanced by attaching flanges to the slotted section of waveguide parallel to the direction of propagation (see end view of Fig. 19-34).

Figure 19-35 shows two forms of *flatplate array* antennas constructed from slotted waveguide radiator elements (shown as insets). Figure 19-35A shows the rectangular array, and Fig. 19-35B shows the circular array. These flatplate arrays are used extensively in microwave communications and radar applications.

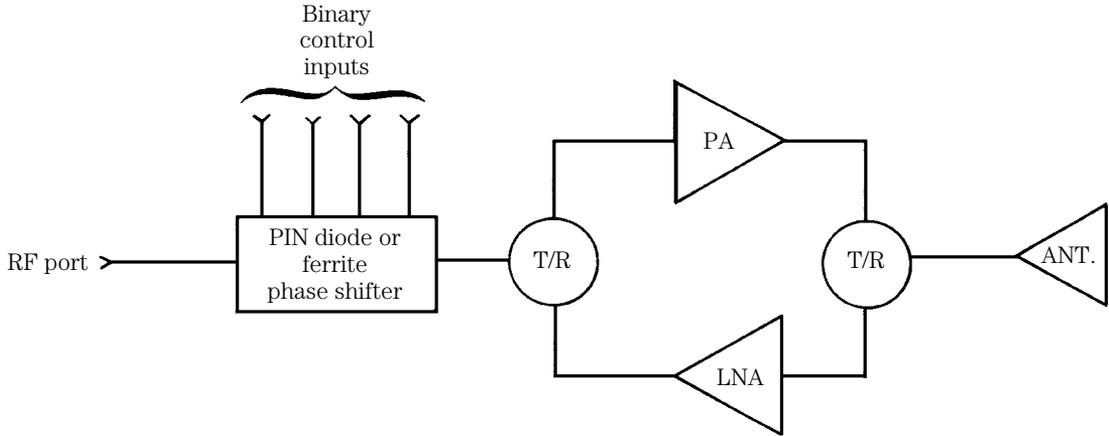
The feed structure for a flatplate array is shown in Fig. 19-35C. The antenna element is the same sort as shown in Figs. 19-35A and 19-35B. A distribution waveguide is physically mated with the element, and a coupling slot is provided between the two waveguides. Energy propagating in the distribution system waveguide is



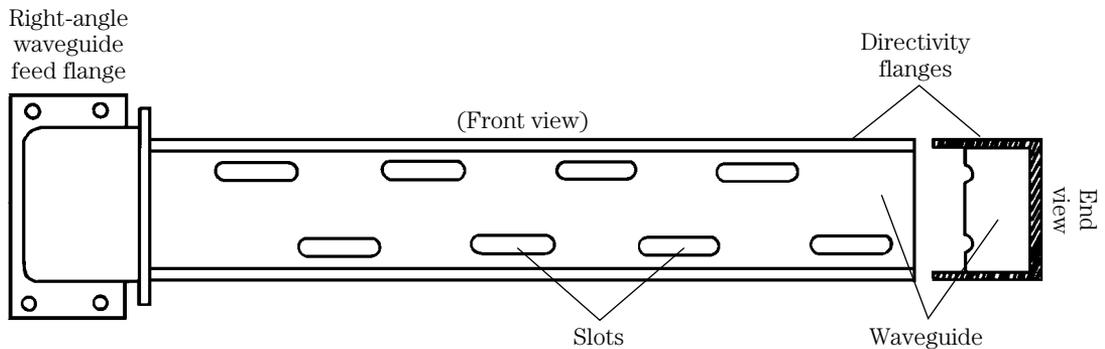
**19-32** Branch feed and corporate feed.



**19-33A** Phase shifter.



19-33B Phase-shift T/R element.



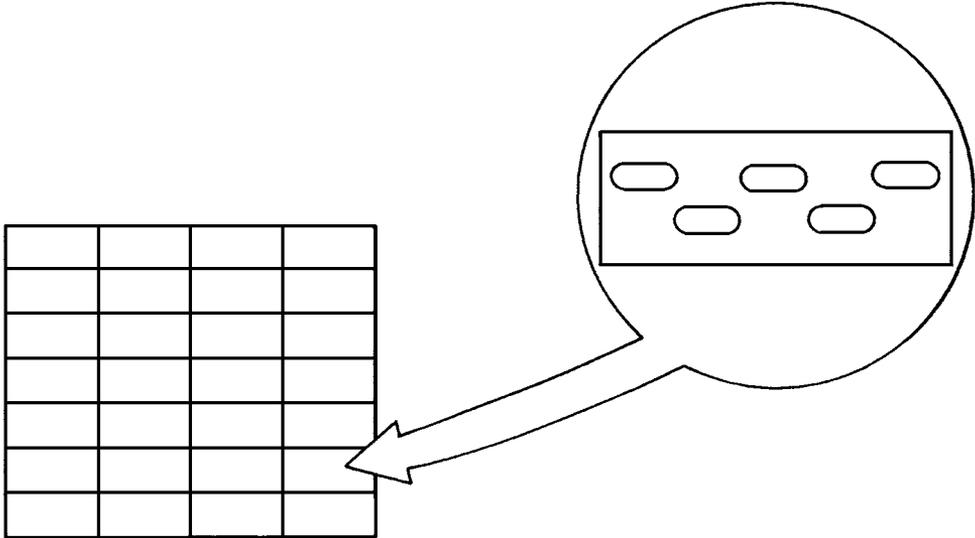
19-34 Slot antenna (front view).

coupled into the antenna radiator element through this slot. In some cases metallic or dielectric phase-shifting stubs are also used in order to fine-tune the antenna radiation pattern.

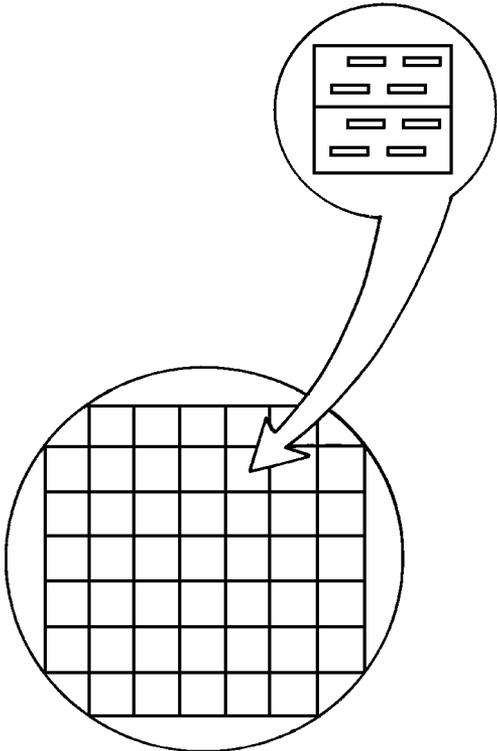
## Microwave antenna safety note

Microwave RF energy is dangerous to your health. Anything that can cook a roast beef can also cook you! The U.S. government sets a safety limit for microwave exposure of  $10 \text{ mW/cm}^2$  averaged over 6 minutes; some other countries use a level one-tenth of the U.S. standard. The principal problem is tissue heating, and eyes seem especially sensitive to microwave energy. Some authorities believe that cataracts form from prolonged exposure. Some authorities also believe that genetic damage to offspring is possible as well as other long-term effects as a result of cumulative exposure.

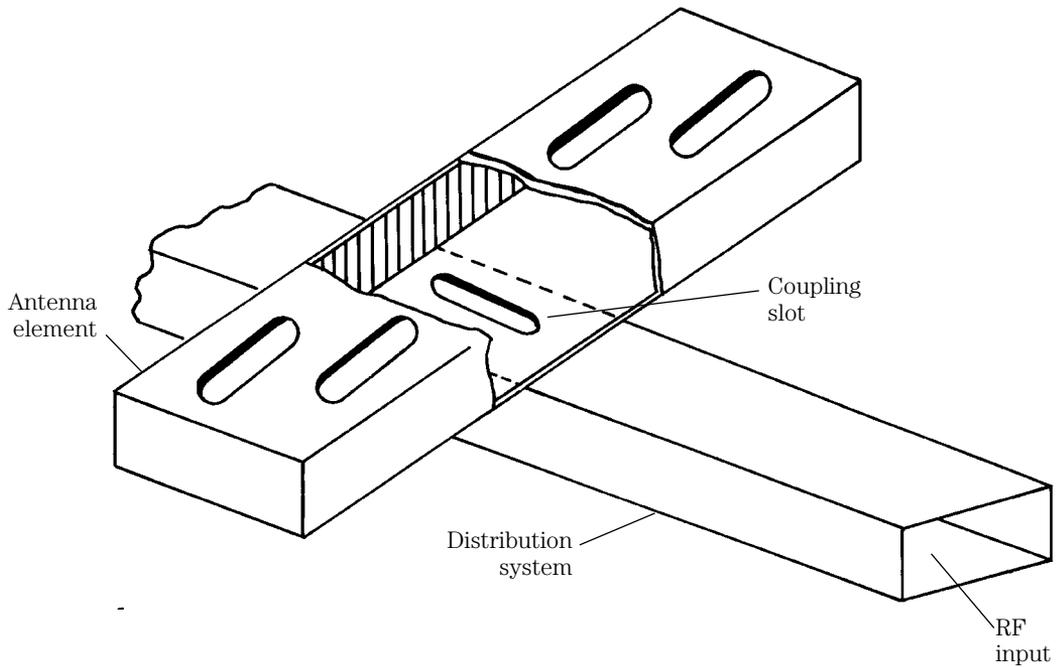
Because of their relatively high gain, microwave antennas can produce hazardous field strengths in close proximity—even at *relatively low RF input power levels*. At



**19-35A** Flatplate slot array.



**19-35B**  
Flatplate antenna.



**19-35C** Flatplate antenna feed coupling.

least one technician in a TV satellite earth station suffered abdominal adhesions, solid matter in the urine, and genital dysfunction after servicing a 45-m-diameter 3.5-GHz antenna with RF power applied.

Be very careful around microwave antennas. Do not service a radiating antenna. When servicing nonradiating antennas, be sure to stow them in a position that prevents the inadvertent exposure of humans, should power accidentally be applied. A *Radiation Hazard* sign should be prominently displayed on the antenna. Good design practice requires an interlock system that prevents radiation in such situations. “Hot” transmitter service should be performed with a shielded dummy load replacing the antenna.

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## 20 CHAPTER

# Antenna noise temperature

RADIO RECEPTION IS ESSENTIALLY A MATTER OF SIGNAL-TO-NOISE RATIO (SNR). SIGNALS must be some amplitude above the noise floor of the system in order to be received properly. All electronic systems (receivers and antennas included) have inherent noise, even if there is no power flowing in them. One of the goals of the antenna designer is to minimize the noise so that weak signals are not obscured. One of the basic forms of noise seen in systems is the *thermal noise*. Even if the amplifiers in the receiver add no additional noise (they will!), there will be thermal noise at the input.

If you replace the antenna with a resistor matched to the system impedance that is totally shielded, there will still be noise present. The noise is produced by the random motion of electrons inside the resistor. At all temperatures above absolute zero (about  $-273.16^{\circ}\text{C}$ ) the electrons in the resistor material are in random motion. At any given instant there will be a huge number of electrons in motion in all directions. The reason why there is no discernible current flow in one direction is that the motions cancel each other out even over short time periods. The noise power present in a resistor is:

$$P_N = KTBR \quad [20.1]$$

where

$P_N$  is the noise power, in watts

$T$  is the temperature, in kelvins (K)

$K$  is Boltzmann's constant ( $1.38 \times 10^{-23}$  K)

$B$  is the bandwidth, in hertz

$R$  is the resistance, in ohms

*Note:* By international agreement  $T$  is set to 290 K.

Consider a receiver with a 1-MHz bandwidth and an input resistance of  $50 \Omega$ . The noise power is  $(1.38 \times 10^{-23} \text{ K}) \times (290 \text{ K}) \times (1,000,000 \text{ Hz}) \times (50 \Omega) = 2 \times 10^{-13} \text{ W}$ . This noise is called *thermal noise*, *thermal agitation noise*, or *Johnson noise*.

A resonant antenna can be modeled as an impedance consisting solely of a resistor with a value equal to the feedpoint impedance. If an antenna has a  $50\text{-}\Omega$  feedpoint

impedance, then it will generate exactly the same amount of thermal noise as a resistor of the same value.

## Noise factor, noise figure, and noise temperature

The noise performance of a receiving system can be defined in three different, but related, ways: *noise factor*  $F_n$ , *noise figure* (NF), and *equivalent noise temperature*  $T_e$ ; these properties are definable as a simple ratio, decibel ratio, or kelvin temperature, respectively.

### Noise factor ( $F_n$ )

For components such as resistors, the noise factor is the ratio of the noise produced by a real resistor to the simple thermal noise of an ideal resistor. The noise factor of a radio receiver (or any system) is the ratio of output noise power  $P_{no}$  to input noise power  $P_{ni}$ :

$$F_n = \frac{P_{no}}{P_{ni}} \Big|_{T=290 \text{ K}} \quad [20.2]$$

In order to make comparisons easier, the noise factor is usually measured at the standard temperature ( $T_o$ ) of 290 K (standardized room temperature), although in some countries 299 K or 300 K is commonly used (the differences are negligible).

It is also possible to define noise factor  $F_n$  in terms of output and input signal-to-noise ratios:

$$F_n = \frac{S_{ni}}{S_{no}} \quad [20.3]$$

where

$S_{ni}$  is the input signal-to-noise ratio

$S_{no}$  is the output signal-to-noise ratio

### Noise figure (NF)

The noise figure is a frequently used measure of a receiver's "goodness," or its departure from "idealness." Thus, it is a *figure of merit*. The noise figure is the noise factor converted to decibel notation:

$$\text{NF} = 10 \log F_n \quad [20.4]$$

where

NF is the noise figure, in decibels

$F_n$  is the noise factor

LOG refers to the system of base 10 logarithms

### Noise temperature ( $T_e$ )

The noise "temperature" is a means for specifying noise in terms of an equivalent temperature. Evaluating the noise equations shows that the noise power is directly

proportional to temperature in kelvins, and also that noise power collapses to zero at the temperature of absolute zero (0 K).

Note that the equivalent noise temperature  $T_e$  is *not* the physical temperature of the amplifier, but rather a theoretical construct that is an *equivalent* temperature that produces that amount of noise power. The noise temperature is related to the noise factor by

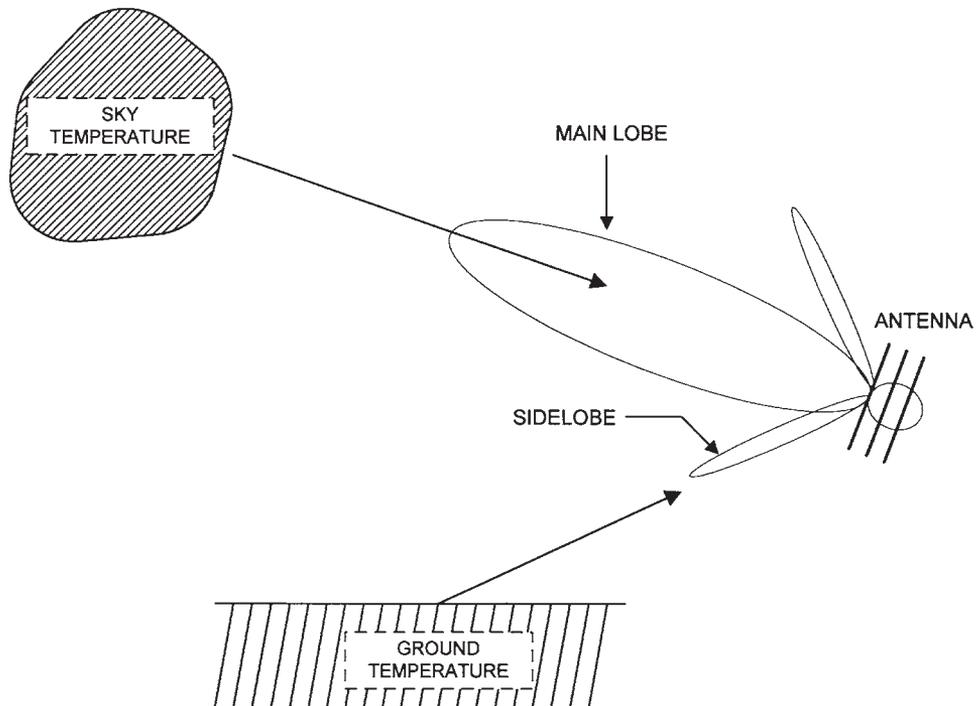
$$T_e = (F_n - 1) T_o \quad [20.5]$$

and to noise figure by

$$T_e = KT_o \log^{-1} \left( \frac{NF}{10} \right) - 1 \quad [20.6]$$

Noise temperature is often specified for receivers and amplifiers in combination with, or in lieu of, the noise figure. The noise temperature concept is also applied to antennas where it is related to the amount of thermal noise generated by the resistive component of the antenna feedpoint impedance.

The antenna-receiver system will be afflicted by three different noise sources external to the receiver. The first is the thermal noise temperature of the feedpoint impedance ( $T_R$ ). The sky exhibits a noise temperature that depends on where the antenna main lobe is pointed. Similarly, the ground has a noise temperature that consists of components reflected from the sky as well as components of its own caused by whatever thermal agitation exists. In a typical system (Fig. 20-1) the



20-1 Contributors to antenna noise temperature.

**420** *Antenna noise temperature*

main lobe will be pointed toward the sky noise source, while the sidelobes will pick up noise from the ground. The total noise temperature of the antenna is

$$T_{\text{ANT}} = (M \times T_{\text{SKY}}) + \xi (1 - M)T_{\text{GND}} + T_R \quad \mathbf{[20.7]}$$

where

$T_{\text{ANT}}$  is the equivalent noise temperature of the antenna

$T_{\text{SKY}}$  is the noise temperature of the sky

$T_{\text{GND}}$  is the noise temperature of the ground

$T_R$  is the feedpoint resistance noise temperature

$M$  is the fraction of the total energy that enters the main lobe

$\xi$  is the fraction of sidelobes that are viewing the ground (only one of several sidelobes is shown in Fig. 20-1)

## 21 CHAPTER

# Antennas for radio astronomy

FOR CENTURIES ASTRONOMERS HAVE SCANNED THE HEAVENS WITH OPTICAL TELESCOPES. But today, astronomers have many more tools in their bag, and one of them is *radio* astronomy. The field of radio astronomy emerged in the 1930s and 1940s through the work of Grote Reber and Carl Jansky. Even during World War II, progress was made as many tens of thousands of operators were listening to frequencies from dc to near daylight (well, actually, the low-end microwave bands). British radar operators noted during the Battle of Britain that the distance at which they could detect German aircraft dropped when the Milky Way was above the horizon.

Although there is a lot of amateur radio astronomy being done, most of it requires microwave equipment with low-noise front ends. However, there are several things that almost anyone can do.

The topic of antennas for radio astronomy can include nearly all forms of directional gain antenna. It is common to see Yagis, ring Yagis, cubical quads, and other antennas for lower-frequency use (18 to 1200 MHz). Microwave gain antennas can be used for higher frequencies. Indeed, many amateur radio astronomers appropriate TV receive-only (TVRO) satellite dish antennas for astronomy work. In this chapter we will cover some antennas that are not found in other chapters, at least not in this present context.

## Jupiter reception

Jupiter is a strong radio source (next to the sun, it's the strongest in the sky). It produces noiselike signals over the spectrum 5 to 40 MHz, with peaks between 18 and 24 MHz. One source claims that the radio signals come from massive storms on the largest planet's surface, apparently triggered by the transit of the jovian moons through the planet's magnetic field. The signals are plainly audible on the HF band any time Jupiter is above the horizon, day or night. However, in order to eliminate the possibility of both local and terrestrial skip signals from interfering, Jupiter DX-ers prefer to listen during the hours after 2100 [or whenever the maximum usable frequency (MUF) drops significantly below 18 MHz] and local sunup. Listen to the

amateur 17- or 15-m bands. If you hear zilch activity, then it's a good bet that the MUF has dropped enough to make listening worthwhile. Even during the day, however, it is possible to hear jovian signals, but differentiating them from other signals or solar noise is difficult.

So what do you need to DX Jupiter? It would help to have a decent radio receiver that works well over the range 18 to 24 MHz. Some cheaper radio receivers are not desirable, but most modern communications receivers are fine. The radio signals are rising and falling "swooshing" noises. The chances of receiving a signal from Jupiter are about 1 in 6, according to several radio astronomers.

The antenna can be a simple dipole cut for the middle of the 18- to 24-MHz band, which happens to be a 15-m amateur radio band antenna. The antenna should be installed in the normal manner for any dipole, except that the wire must run east-west in order to pick up the southerly rising planet.

Figure 21-1 shows a broadband dipole that covers the entire frequency region of interest (18 to 24 MHz) by paralleling three different dipoles: one cut for 18 MHz, one cut for 21 MHz, and one cut for 24 MHz. The dimensions are

A	19.5 ft	24 MHz
B	22.3 ft	21 MHz
C	26 ft	18 MHz

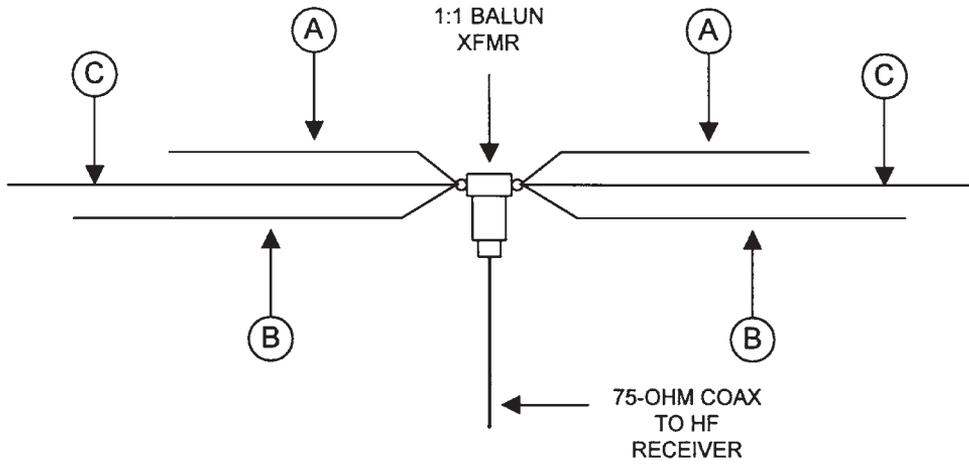
There are several approaches to making this type of antenna. One is to use three-conductor wire, and cut the wires to the lengths indicated above. Another is to use a homemade spacer to spread the wires apart.

## Ring antenna

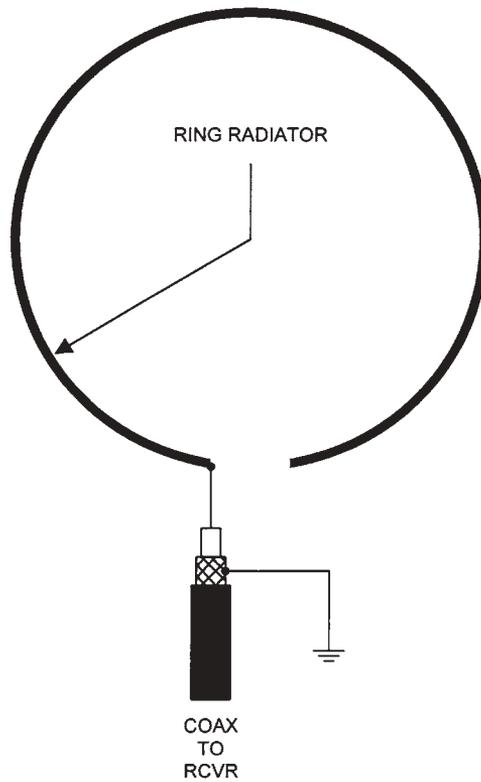
Another popular jovian radio antenna is the ring radiator, two versions of which are shown in Fig. 21-2. This antenna is made of a 5-ft-diameter loop of  $\frac{1}{2}$ -in-diameter soft-drawn copper plumbing pipe. The single-ended version is shown in Fig. 21-2A. In this antenna the loop is open-ended. The center conductor of the coaxial cable feedline is connected to one end of the ring radiator, while the coax shield is connected to the chicken wire ground plane. The balanced version (Fig. 21-2B) has an RF transformer ( $T_1$ ) at the feedpoint.

The ring radiator antenna should have a bandpass preamplifier. The preamp is needed because of the low pickup of this kind of antenna, and that preamp should be mounted as close as possible to the antenna. The bandpass characteristic is obtained by filtering. The idea is to get rid of terrestrial signals in the adjacent bands. The typical preamplifier is broadbanded (which is what we want), so will pick up not only the desired signals but others as well (which is what we don't want). Even a 5-W CB transmitter a few blocks away can drive the preamplifier into saturation, so it's wise to eliminate the undesired signals before they get into the preamplifier. In the case of the single-ended amplifier, a single-ended preamplifier is used. But for the balanced version (Fig. 21-2B) a differential preamplifier is used.

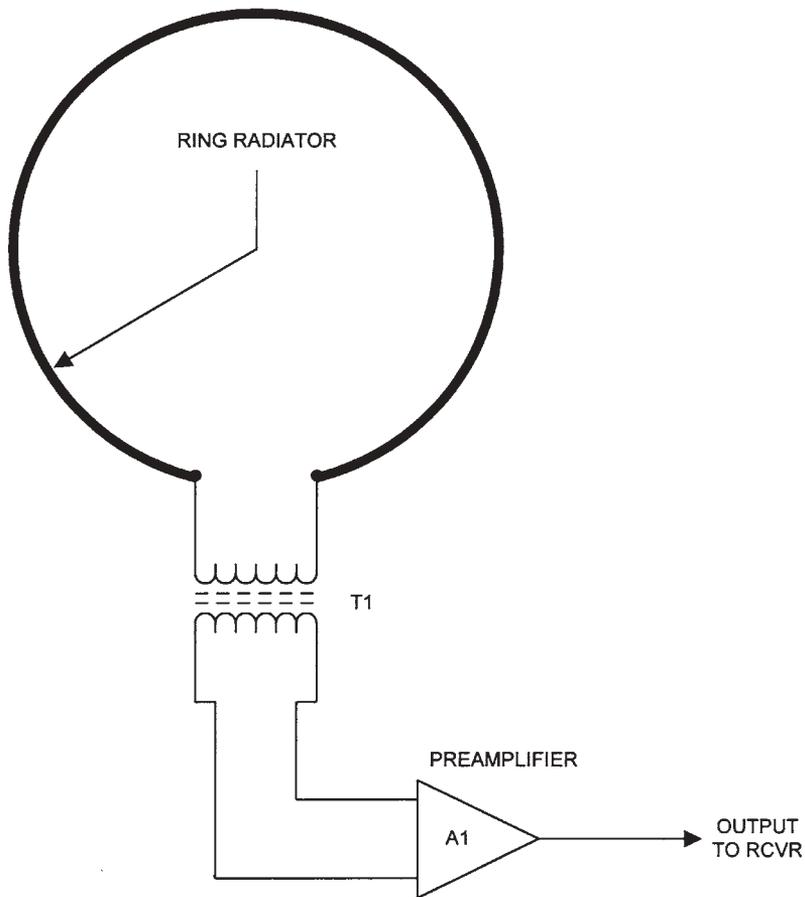
The loop (both versions) is mounted about 7 or 8 in above a groundplane made of chicken wire, metal window screen, copper sheeting, or copper foil. The copper sheeting



21-1 Wideband dipole such as might be used for Jupiter reception.



21-2A Ring radiator: single-ended.



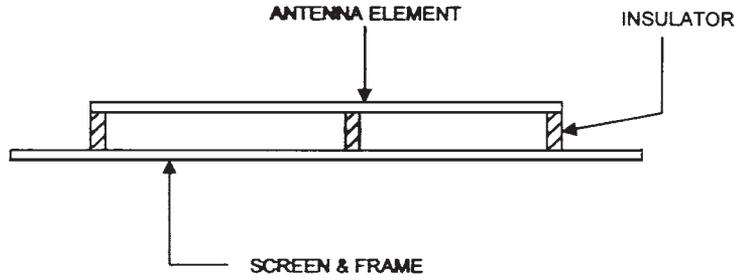
**21-2B** Ring radiator: balanced.

or foil is best, but costs a lot of money, and turns ugly green after a couple weeks in the elements. If you use screen, make sure that it is a metallic screen. Some window and porch screening material is made of synthetic materials that are insulators.

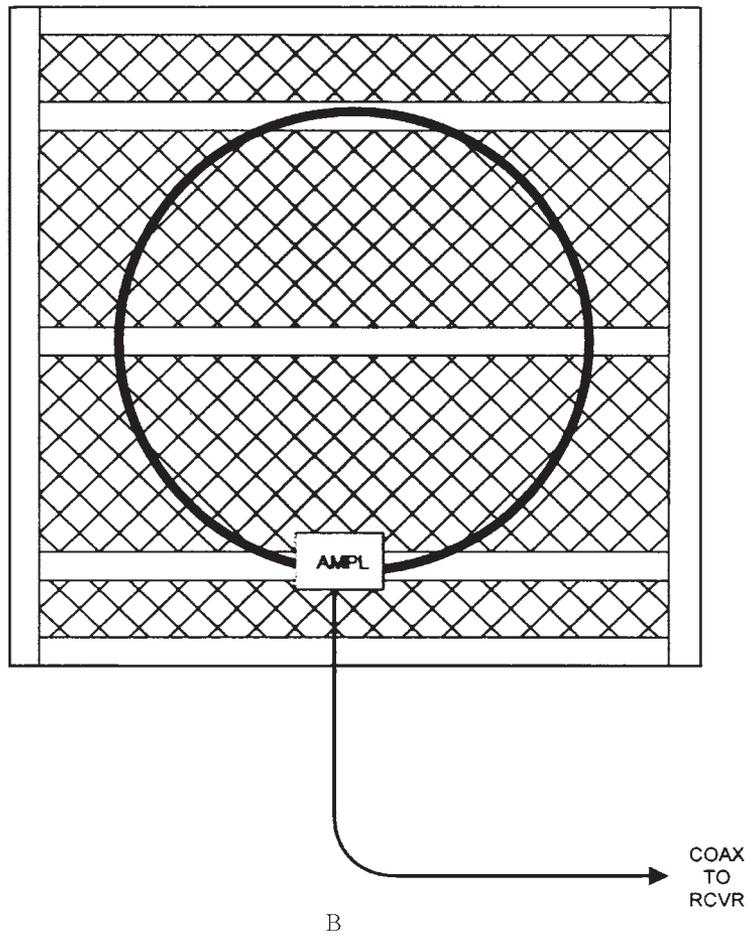
Figure 21-3A is a mechanical side view of the ring radiator antenna, while Fig. 21-3B is a side view. The antenna is mounted above the screen with insulators. These can be made of wood, plastic, or any other material. The frame holding the ground-plane screen (Fig. 21-3B) can be made from  $1 \times 2$ -in lumber. Note that the frame has interior crosspieces to support the antenna, as well as the outer perimeter. The larger outer perimeter is needed because the screen groundplane should extend beyond the diameter of the radiator element by about 10 to 15 percent.

## DDRR

The directional discontinuity ring radiator (DDRR) antenna is shown in Fig. 21-4A, while a side view showing the mounting scheme is shown in Fig. 21-4B. It is typically

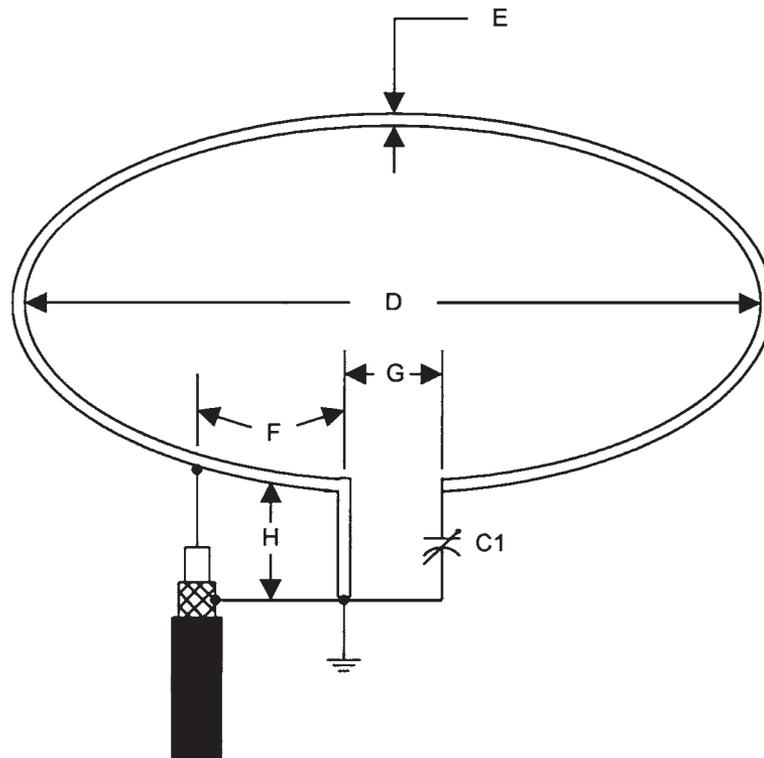


A

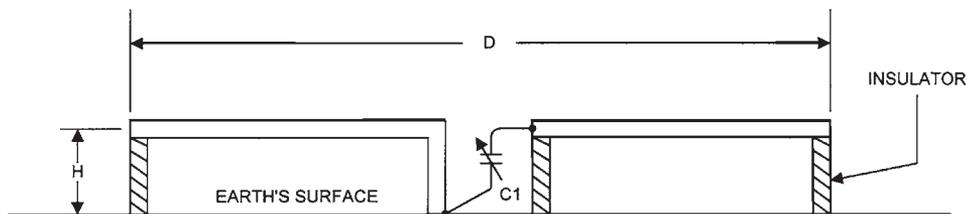


B

21-3 Ring radiator mounted over a ground screen.



21-4A DDRR antenna.



21-4B DDRR antenna: side view.

mounted about 1 ft off the ground ( $H=12$  in). The DDRR consists of two sections, one vertical and one horizontal. The short vertical section has a length equal to the height  $H$  of the antenna above the ground. One end of the vertical segment is grounded. The horizontal section is a loop with a diameter  $D$ .

The conductor diameter  $E$  is at least 0.5 in at 28 MHz, and 4 in at 4 MHz. Because of the loop, some people call this the hula hoop antenna. One author recommends using a 2-in automobile exhaust pipe bent into the correct shape by an auto muffler dealer. The far end of the loop is connected to ground through a small-value tuning

capacitor  $C_1$ . The actual value of  $C_1$  is found experimentally and is used to resonate the antenna to a particular operating frequency.

The feedline of the DDRR antenna is coaxial cable connected such that the shield is grounded at the bottom end of the vertical section. The center conductor is connected to the ring radiator a distance  $F$  from the vertical section. The length  $F$  is determined by the impedance that must be matched. The radiation resistance is approximated by

$$R_T = \frac{2620H^2}{\lambda^2} \quad [21.1]$$

where

$R_T$  is the radiation resistance, in ohms

$H$  is the height of the antenna off the ground

$\lambda$  is the wavelength

(Both  $H$  and  $\lambda$  are in the same units.)

The approximate values for the various dimensions of the DDRR are given below in general terms, with examples in Table 21-1:

$D$	$0.078\lambda$
$H$	$0.11D$
$F$	$0.25H$
$E$	0.5 to 4 in
$G$	See Table 21-1

**Table 21-1. Examples of dimensions for DDRR**

Dimension	Band (MHz)							
	1.8	4	7.5	15	22	30	50	150
$G$ (in)	16	7	5	3	2.5	2	1.5	1
$C_1$ (pF)	150	100	75	35	15	12	10	6
$F$ (in)	12	6	6	1.5	1.5	1	1	0.5
$H$ (in)	48	24	11	6	4.75	3	1.5	1
$D$ (ft)	36	18	9	4.5	3.33	2.33	1.4	6
$E$ (in)	5	4	2	1	0.75	0.75	0.5	0.25

The construction details of the DDRR are so similar to those of the ring radiator that the same diagram can be used (see Fig. 21-3 again).

The normal attitude of the DDRR for communications is horizontal. However, for Jupiter radio reception, the antenna groundplane screen can be tilted to face Jupiter's point in the sky.

## Helical antennas

The helical antenna (Fig. 21-5) provides moderately wide bandwidth and circular polarization. Because of the polarization some people find the helical antenna to be

particularly well suited to radio astronomy reception. The antenna (of diameter  $D$ ) will have a circumference  $C$  of  $0.75\lambda$  to  $1.3\lambda$ . The pitch of the helix ( $S$ ) is the axial length of one turn, while the overall length  $L=NS$  (where  $N$  is the number of turns). The ratio  $S/C$  should be  $0.22\lambda$  to  $0.28\lambda$ . At least three turns are needed to produce axial-mode main lobe maxima.

The diameter or edge of the groundplane  $G$  should be on the order of  $0.8\lambda$  to  $1.1\lambda$  if circular and/or square, respectively. The offset between the groundplane and the first turn of the helix is  $0.12\lambda$ .

The approximate gain of the helical antenna is found from

$$\text{Gain} = 11.8 + 10 \log(C^2 NS) \quad \text{dBi} \quad \mathbf{[21.2]}$$

The pitch angle  $\phi$  and turn length  $\gamma$  for the helical antenna are given by

$$\phi = \tan^{-1}\left(\frac{\pi D}{S}\right) \quad \mathbf{[21.3]}$$

and

$$\gamma = \sqrt{(\pi D)^2 + S^2} \quad \mathbf{[21.4]}$$

The beamwidth of the helical antenna is

$$\Theta = \frac{K}{\sqrt{NS}} \quad \mathbf{[21.5]}$$

where

$\Theta$  is the beamwidth, in degrees

$N$  is the number of turns

$S$  is the pitch, in wavelengths

$C$  is the circumference, in wavelengths

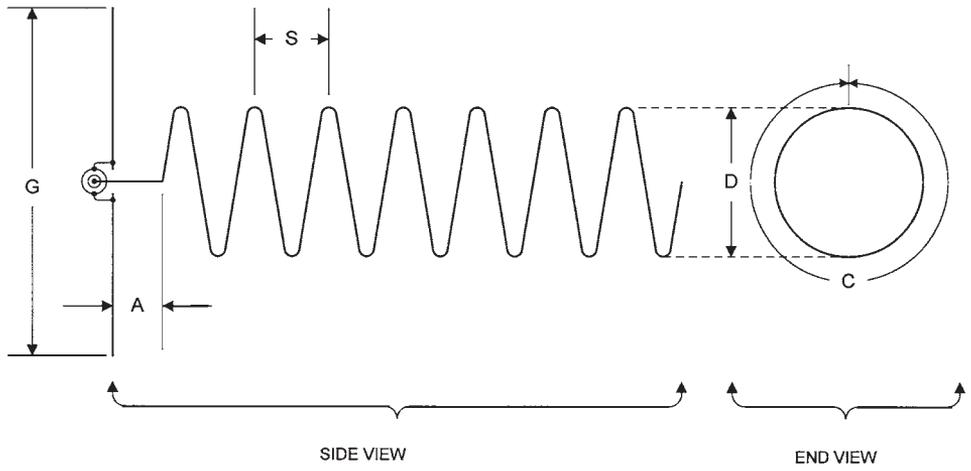
$K$  is 52 for the  $-3$ -dB beamwidth and 115 for the beamwidth to the first null in the pattern

The short section between the helix and the groundplane is terminated in a coaxial connector, allowing the antenna to be fed from the rear of the groundplane. The feedpoint impedance is approximately  $140 \Omega$ .

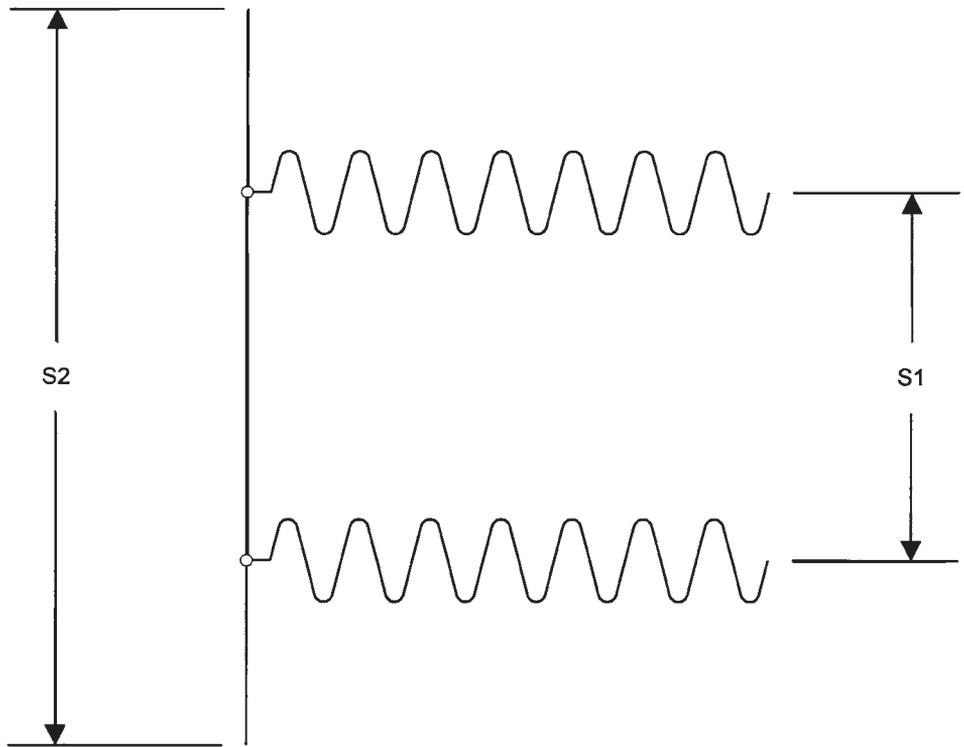
## Multiple helical antennas

Stacking helical antennas allows a radiation pattern that is much cleaner than the normal one-antenna radiation pattern. It also provides a good way to obtain high gain with only a few turns in each helix. If two helixes are stacked, then the gain will be the same as for an antenna that is twice the length of each element, while for four stacked antennas the gain is the same as for a single antenna 4 times as long. Figure 21-6 shows a side view of the stacked helixes.

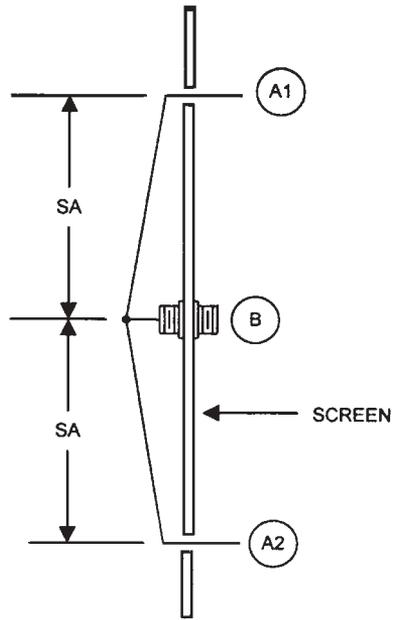
The feed system for stacked helixes is a little more complex than for a single helix. Figure 21-7A shows an end view of a set of four stacked helical antennas. Tapered lines (TL) are used to carry signal from each element and the coaxial connector (B). In this case, the coaxial connector is a feed-through “barrel” SO-239 device at the center of the groundplane (B). A side view of the tapered line system



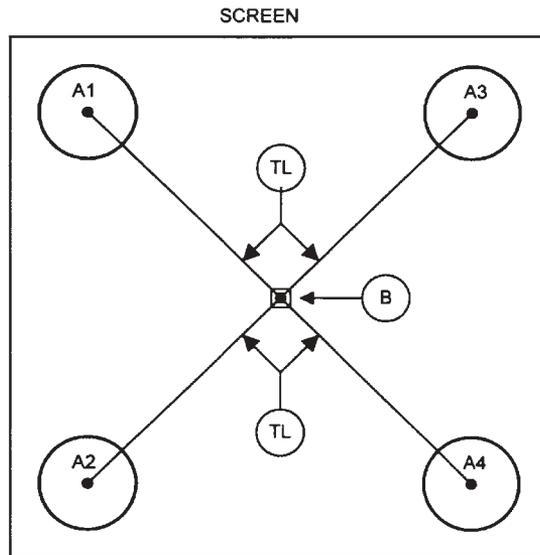
21-5 Helical antenna.



21-6 Dual/quad helical antenna.



B



A

**21-7** Frontal view of quad helical antenna.

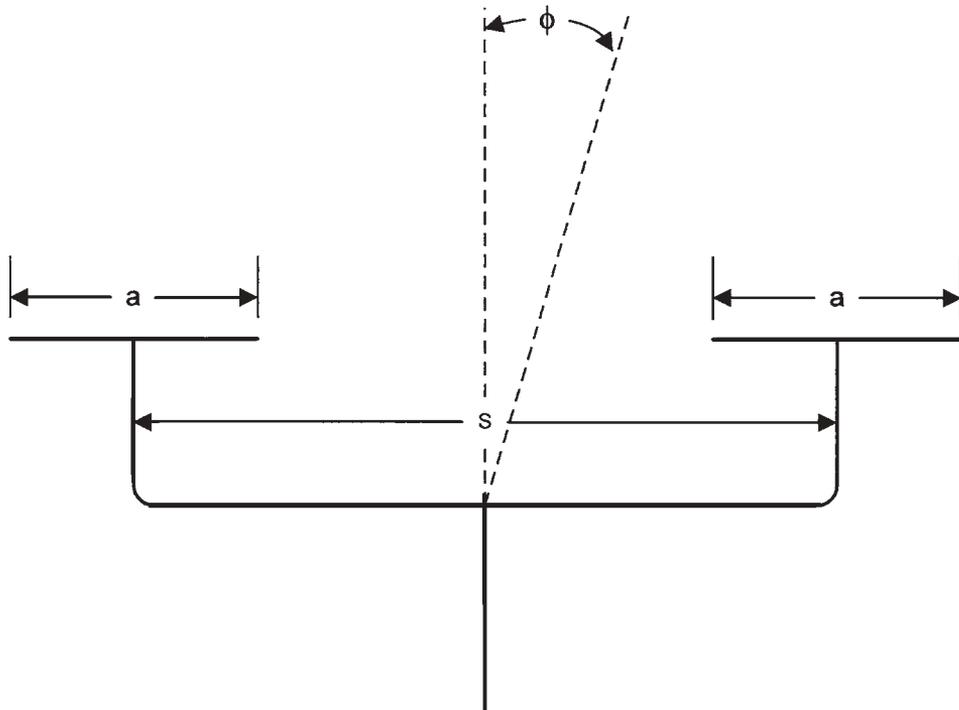
is shown in Fig. 21-7B. The length of the tapered lines is  $1.06\lambda$ , while the center-to-center spacing between the helical elements is  $1.5\lambda$ . The length of each side of the groundplane is  $2.5\lambda$ . In the case of Fig. 21-7, the antenna is fed from the front of the groundplane.

## Interferometer antennas

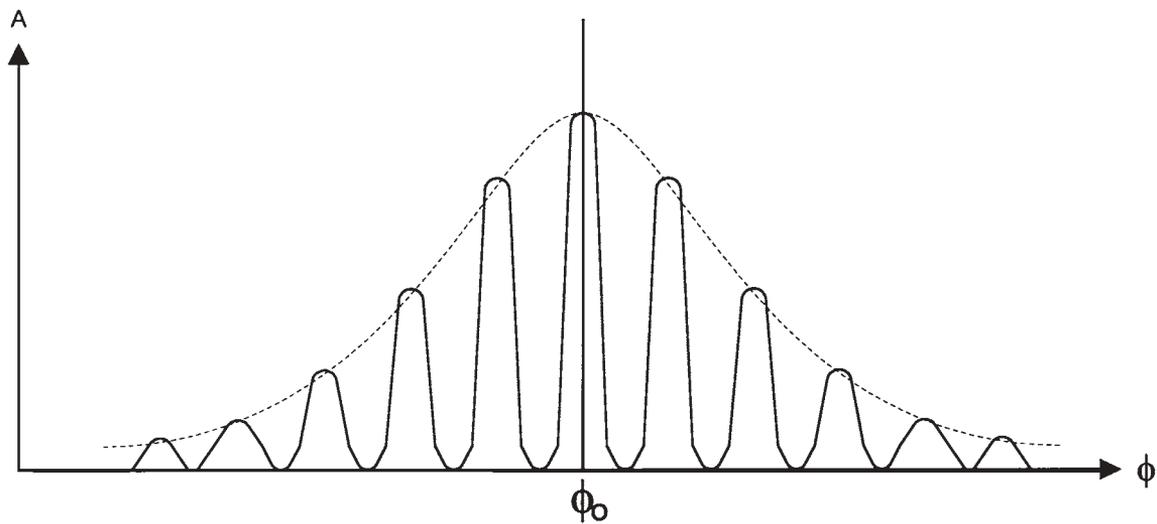
The resolution of an antenna is set by its dimensions  $a$  relative to the wavelength  $\lambda$  of the received signal. Better resolution can be achieved by increasing the size of the antenna, but that is not always the best solution. Figure 21-8 shows a *summation interferometer array*. Two antennas with aperture  $a$  are spaced  $S$  wavelengths apart. The radiation pattern is a fringe pattern (Fig. 21-9). This pattern consists of a series of maxima and nulls. The resolution angle  $\phi$  to the first null is

$$\phi = \frac{S+a}{57.3} \quad [21.6]$$

The interferometer can be improved with additional antennas in the array. Professional radio astronomers use very wide baseline antennas. With modern communications it is possible to link radio telescopes on different continents to make the widest possible baseline.



21-8 Interferometer array.



**21-9** Interferometer pattern.

## 22 CHAPTER

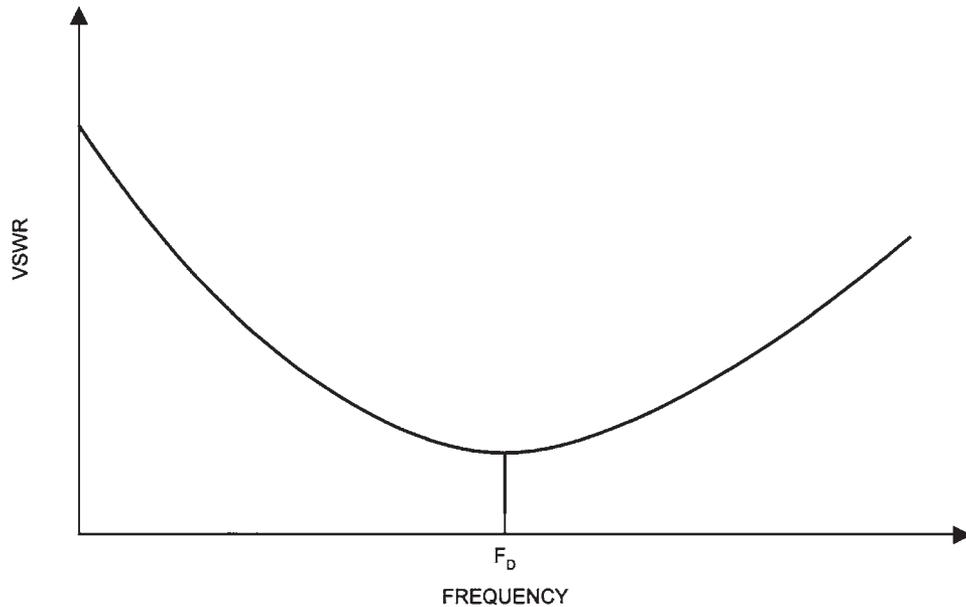
# Adjusting, installing, and troubleshooting antennas and transmission lines

ANTENNA AND TRANSMISSION LINE MEASUREMENTS SHOULD BE MADE WHEN THE ANTENNA is first installed and periodically thereafter. If some difference in operation is noted, the same measurements should be repeated. Many antenna measurements are difficult to make with any degree of accuracy. There are some things about antennas that can and should be measured, however, regardless of the difficulty. For example, VSWR and the resonant frequency of the antenna are readily accessible. It's also possible to measure the impedance of the antenna feedpoint. You can measure the VSWR either with a special VSWR meter (often built into transmitters or antenna tuning units), or by using an RF wattmeter.

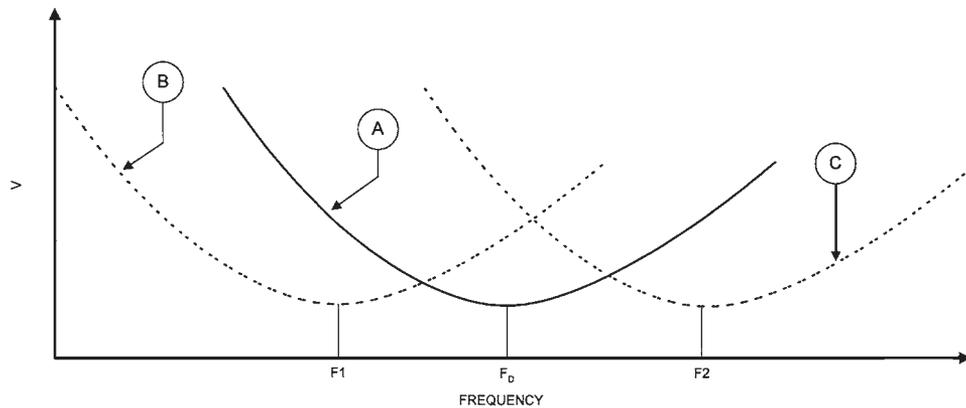
By frequency stepping through the band and testing the VSWR at various frequencies, one can draw a VSWR curve (Fig. 22-1) that shows how the antenna performs across the band. The resonant frequency is the point where the VSWR dips to a minimum (which may or may not be the much sought-after 1:1).

You can use the resonant frequency to figure out whether the antenna is too long (resonant frequency lower than the hoped-for design frequency), or too short (resonant frequency above the design frequency). Figure 22-2 shows all three situations. Curve A represents the desired or ideal curve that is centered on the desired frequency  $F_D$ . If the antenna is too long, however, the resonant frequency will be shifted downward towards  $F_1$ , and curve B will be observed. Similarly, if the antenna is too short, curve C will be found as the resonant frequency shifts upband to  $F_2$ .

But resonant frequency and VSWR curves are not the entire story because they don't tell us anything about the impedance presented by the antenna. One cannot get the VSWR to be 1:1 unless the antenna impedance and transmission line impedance are the same. For example, a dipole has a nominal textbook impedance of  $73 \Omega$ ,



**22-1** VSWR-vs-frequency plot.



**22-2** VSWR-vs-frequency plot for three frequencies.

so it makes a very good match to  $75\text{-}\Omega$  coaxial cable. But the actual impedance of a real dipole may vary from a few ohms to more than  $100\ \Omega$ . For example, if your antenna exhibits a feedpoint impedance of (say)  $25\ \Omega$ , using  $75\text{-}\Omega$  coaxial cable to feed it produces a  $\text{VSWR}=75/25=3:1$ . Not too great. Measuring the feedpoint impedance is therefore quite important to making the antenna work properly.

There are any number of instruments on the market that will aid in making antenna measurements. Some of them are quite reasonably priced (or can be built), while others are beyond the reach of all but the most ardent and well-endowed en-

thusiasts. In this chapter we will look at a newer breed of more universal instrument called the *SWR analyzer*.

## VSWR analyzers

One basic premise in this chapter is that the instruments used for measuring VSWR must be accessible to people who don't have a commercial or ham operator's license, as well as those who do (we must remember our SWL and scanner friends). Some of the instruments used by hams meet that requirement, but a relatively new breed of instrument called the SWR analyzer provides a lot of capability to the SWL, scanner operator, and ham radio operator alike. It uses a low-power RF signal generator and some clever circuitry to measure the VSWR of the antenna. One model also measures the feedpoint resistance.

The MFJ-259 VSWR analyzer (MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, MS, 39762) is shown in Fig. 22-3. This instrument combines a VSWR analyzer with a digital frequency counter, and operates over the range 1.8 to 170 MHz. A band switch is set to the desired band, and then the *tune* control is set to the desired frequency. The meter will then read the VSWR at the design frequency. Alternatively, you can adjust the tune control until the minimum VSWR is found. This frequency is the actual, versus the desired, resonant frequency of the antenna. The front panel of the MFJ-259 has two meters, *SWR* and *RESISTANCE*. The SWR meter is calibrated up to 3:1, with a little uncalibrated scale to indicate higher SWRs.

The resistance meter is calibrated from 0 to 500  $\Omega$ , which is consistent with the SWR range. Two controls on the front panel are *tune* and *frequency* (a band switch). The MFJ-259 has a digital frequency meter to measure the operating frequency of the internal oscillator. This frequency counter can also be used to measure the frequency of external signal sources. (*Do not* connect the counter to the output of a transmitter; the instrument will be destroyed.) The top end of the MFJ-259 has a number of controls and connectors. An SO-239 UHF-style coaxial connector is provided for the antenna connection. A BNC coaxial connector is provided to apply external signal to the frequency counter, while a push button *input* switch is available to switch the counter from internal to external signal sources. Another push button switch is used to set the gate timing of the counter (a red LED on the front panel blinks every time the gate is triggered). The tuning is from 1.8 MHz to 174 MHz, while the counter will measure up to 200 MHz.

The MFJ-259 will work from an external 12-Vdc source, or from an internal battery pack consisting of eight size-AA standard cells. MFJ recommends that either alkaline or rechargeable batteries, rather than ordinary zinc-carbon cells, be used in order to reduce the possibility of leakage that can damage the instrument (this is good practice in all battery-powered instruments).

Unlike many lesser SWR meters, this instrument is not fooled by antennas that have impedances consisting of both resistance and reactance elements. An example in the manual demonstrates an impedance of  $25 + j25 \Omega$  (i.e.,  $R$  is 25  $\Omega$  and reactance  $X$  is also 25  $\Omega$ ). When connected to a 50- $\Omega$  load one might be tempted to think the VSWR is 1:1, and some cheaper meters will so indicate. But the actual SWR is 2.6:1, which is what the MFJ-259 will read.



**22-3** MFJ-259 VSWR analyzer.

The resistance measurement assumes a resistive load (i.e., the measurement is made at the antenna's resonant frequency), and is referenced to  $50\ \Omega$ . The VSWR and resistance measurements should be consistent with each other. If the VSWR is 2:1, then the resistance should be either  $100\ \Omega$  ( $100/50 = 2:1$ ) or  $25\ \Omega$  ( $50/25 = 2:1$ ). If the resistance is not consistent with the VSWR reading, then you should assume that the impedance has a significant reactive component and take steps to tune it out.

In addition to antenna measurements, the MFJ-259 is equipped to measure a wide variety of other things. It will measure the velocity factor of transmission line,

help in tuning or adjusting matching stubs or matching networks, measure capacitance or inductance, and determine resonant frequency of  $LC$  networks.

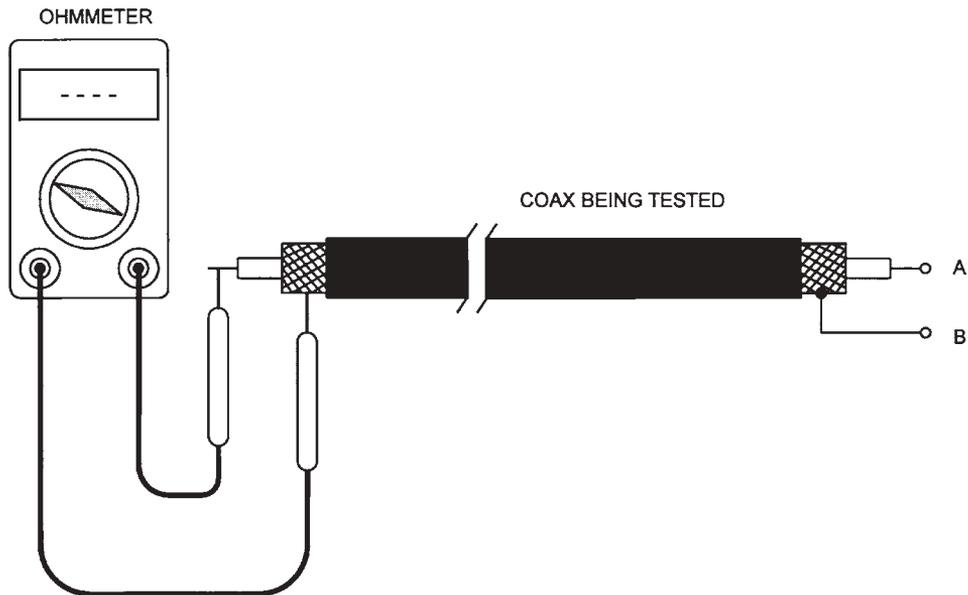
## Doping out coaxial cable

When you install an antenna, or do a bit of preventive maintenance, or find the antenna is not working properly, one thing to check is the transmission line. Two basic measurements are popular. Figure 22-4 shows how to make ohmic checks. The cable consists of an inner conductor and an outer conductor (shield). With terminals A and B open there should not be any resistive path across the input terminals (as shown). If a high resistance is seen, then there might be some contamination in the system, or the insulation has failed, permitting a current path. A low resistance indicates a short circuit. If the cable has been cut, or an object passed through, or the connector is messed up, then a short can result.

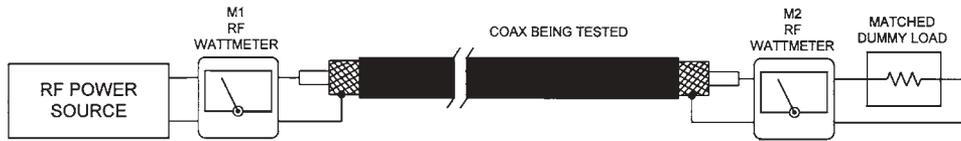
If terminals A-B are shorted together, a low resistance should be noted. If not, then it is likely that the center conductor is open. Of course, if connectors are on the line, either the shield or center conductor could be at fault.

The other issue is loss of the transmission line. Cable losses get higher as cables age, so low loss on installation does not guarantee low loss later on. Losses in new cables run from 0.2 dB/100 ft at 4 MHz. At 30 MHz the losses are 0.7 dB/100 ft to 2.6 dB/100 ft. At VHF/UHF frequencies the losses rise considerably.

Figure 22-5 shows a test setup for coaxial cable. It is necessary to know the length of the piece of coax under test. For ease of calculations a 100-ft section



22-4 Ohmmeter used to test coax.

**22-5** Attenuation test for coax.

should be used. Two RF power meters ( $M_1$  and  $M_2$ ) are used to measure the input power to the line ( $M_1$ ) and the delivered power ( $M_2$ ). The difference in power determines the loss. The loss is

$$\text{Loss}_{\text{dB}} = 10 \log \left( \frac{P_{M2}}{P_{M1}} \right) \quad [22.1]$$

where

Loss is the loss of the cable, in dB

$P_{M1}$  is the power reading on  $M_1$

$P_{M2}$  is the power reading on  $M_2$

If the cable length is 100 ft, then you already have the loss in dB/100 ft. But if length  $L$  is anything other than 100 ft, then you need to make the following calculation:

$$\text{Loss}_{\text{dB}/100\text{ft}} = \frac{\text{Loss}_{\text{dB}}}{L_{\text{ft}}} \times 100 \quad [22.2]$$

## 23 CHAPTER

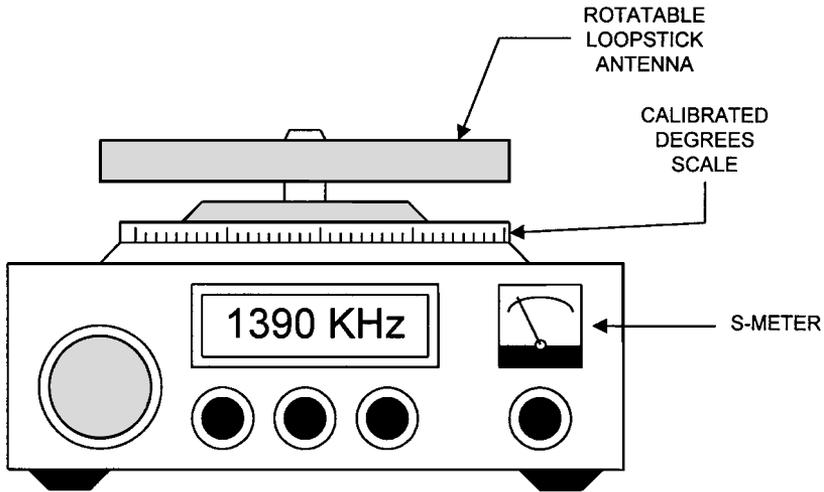
# Antennas for radio direction finding (RDF)

RADIO DIRECTION FINDING IS THE ART AND PRACTICE OF EITHER LOCATING YOURSELF OR A radio station by using a directional radio antenna and receiver. When the FCC wants to locate an illegal station that is transmitting, they will use radio direction finders to triangulate the position. If they find the bearing from two stations they will locate the station at the intersection. However, there is a fair degree of ambiguity in the measurement. As a result, radio direction finders typically use three or more (hence “*triangulate*”) sites. Each receiving site that can find the bearing to the station reduces the overall error.

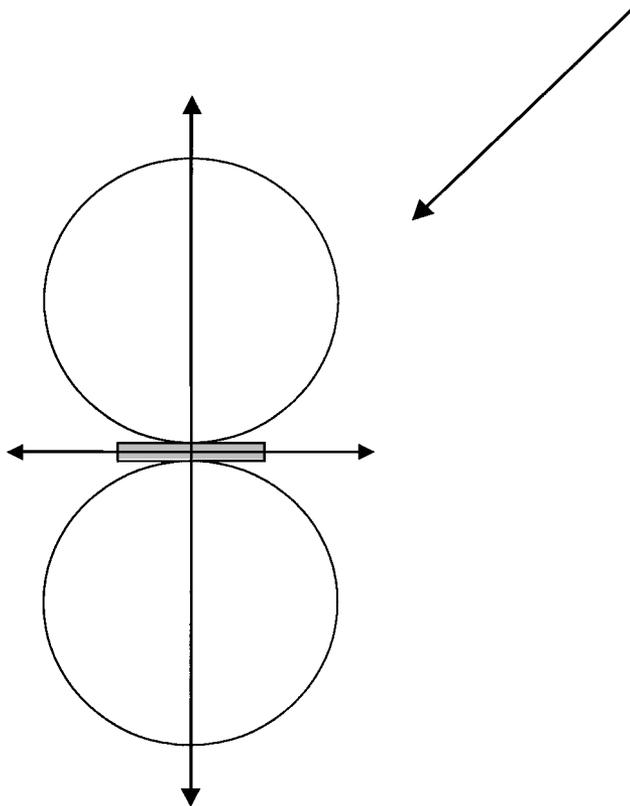
At one time aviators and seafarers relied on radio direction finding. It is said that the Japanese air fleet that attacked Pearl Harbor, Hawaii, on December 7, 1941 homed in on a Honolulu AM radio station. During the 1950s and early 1960s AM radios came with two little circled triangle marks at the 640-kHz and 1040-kHz points on the dial. These were the CONALRAD frequencies that you could tune to in case of a nuclear attack (right before you kissed your butt good-bye, I suspect). All other radio stations were off the air except the CONALRAD stations. The enemy was prevented from using these frequencies for RDF because the system used several stations that transmitted in a rapidly rotating pattern. No one station was on the air long enough to allow a “fix.” The result was a wavering sound to the CONALRAD station (which we heard during tests) that would confuse any enemy who tried to DX his way into our cities with nuclear weapons.

Radio direction finders based on the AM broadcast band (BCB) looked a bit like Fig. 23-1. A receiver with an S meter (which measures signal strength) is equipped with a rotatable ferrite loopstick antenna to form the RDF unit. A degree scale around the perimeter of the antenna base could be oriented toward north so that the bearing could be read.

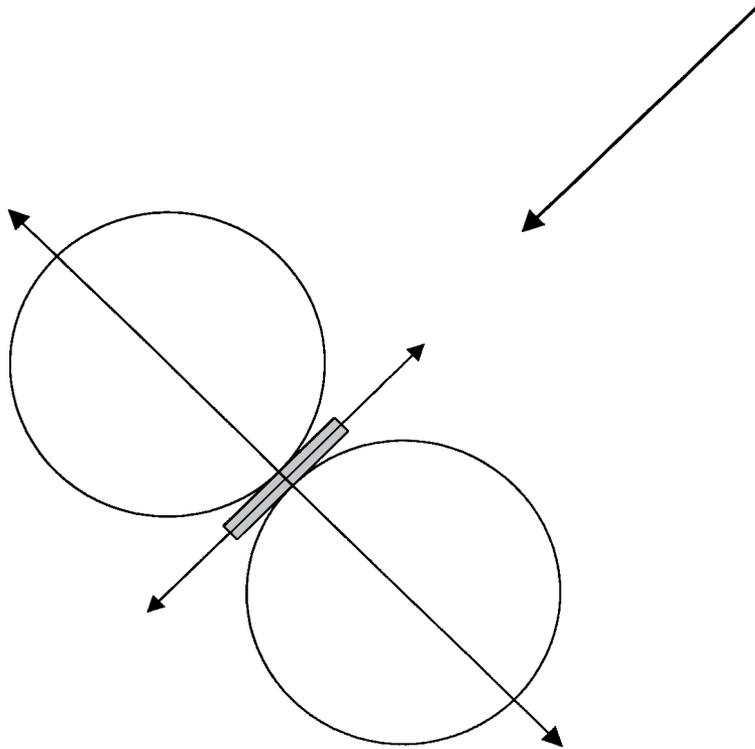
Loopstick antennas have a figure-8 reception pattern (Fig. 23-2A) with the maxima parallel to the loopstick rod and the minima off the ends of the rod. When the antenna is pointed at the signal, maximum reception strength is achieved. Unfortunately, the maxima are so broad that it is virtually impossible to find the true point on the compass dial where the signal peaks. The peak is too shallow for that



**23-1** Radio direction finding (RDF) receiver using a loopstick antenna.



**23-2A** Pattern of loopstick antenna: not oriented towards signal.

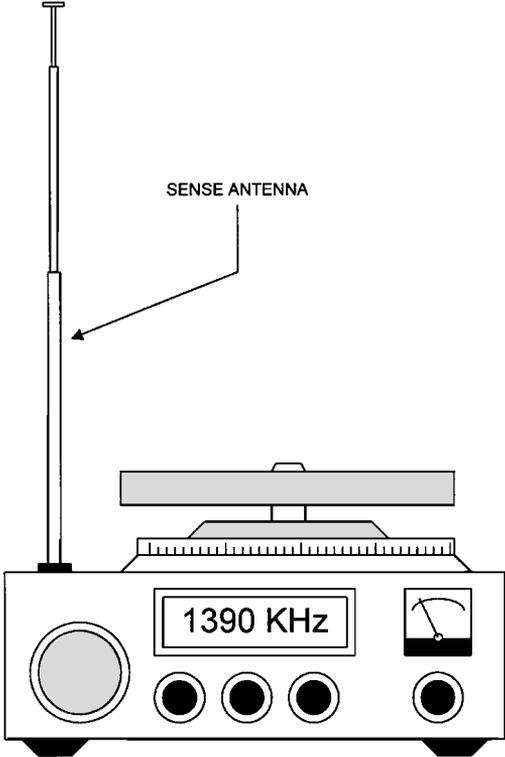


**23-2B** Pattern of loopstick antenna: oriented towards signal.

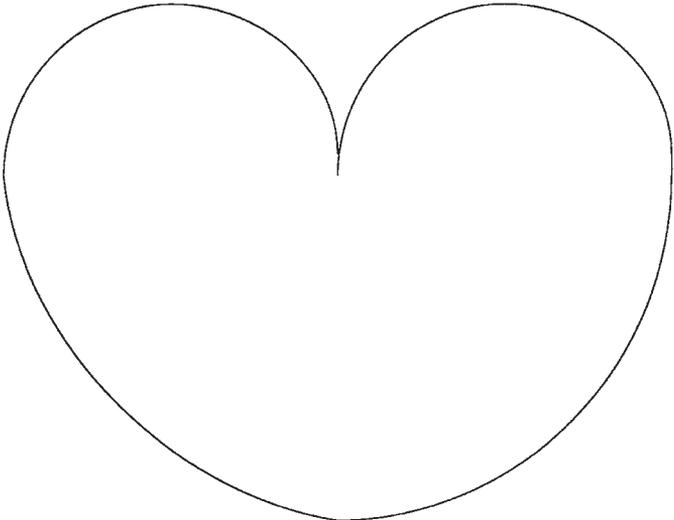
purpose. Fortunately, the minima are very sharp. You can get a good fix on the direction of the signal by pointing the minima toward the station. This point is found by rotating the antenna until the audio goes to zip or the S meter dips to a minimum (Fig. 23-2B).

The loopstick is a really neat way to do RDF—except for one little problem: The darn thing is bidirectional. There are two minima because, after all, the pattern is a figure 8. You will get exactly the same response from placing either minima in the direction of the station. As a result, the unassisted loopstick can only show you a line along which the radio station is located, but can't tell you which direction it is. Sometimes this doesn't matter. If you know the station is in a certain city, and that you are generally south of the city, and can distinguish the general direction from other clues, then the line of minima of the loopstick will refine that information. A compass helps, of course. Shortly we will take a look at an impromptu radio direction finder using a portable radio.

The solution to the ambiguity problem is to add a sense antenna to the loopstick (Fig. 23-3). The sense antenna is an omnidirectional vertical whip, and its signal is combined with that of the loopstick in an *RC* phasing circuit. When the two patterns are combined, the resultant pattern will resemble Fig. 23-4. This pattern is called a *cardioid* because of the heart shape it exhibits. This pattern has only one null, so it resolves the ambiguity of the loopstick used alone.



**23-3** Addition of sense antenna overcomes directional ambiguity of the loopstick.



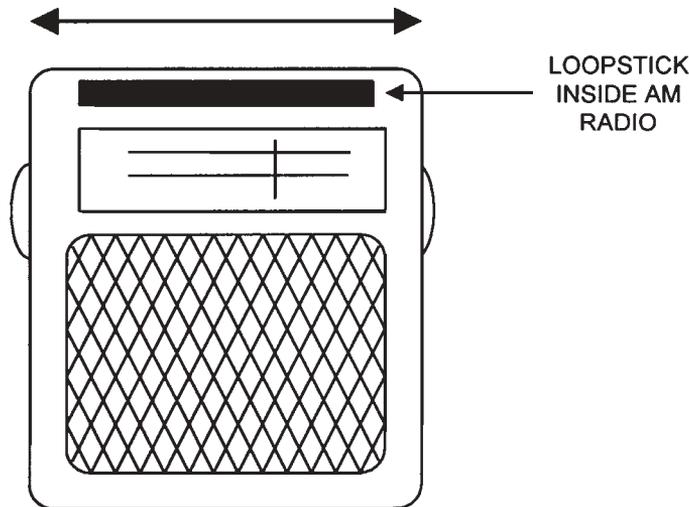
**23-4** Cardioid pattern of sense-plus-loopstick antennas.

## Field improvisation

Let's suppose you are out in the woods trekking around the habitat of lions, tigers, and bears (plus a rattlesnake or two for good measure). Normally you find your way with a compass, a Geological Survey 7.5-min topological map, and a Global Positioning System (GPS) receiver. Those little GPS marvels can give you real good latitude and longitude indication. But what happens if it breaks or a bear eats it? The answer to your direction finding problem might be the little portable AM BCB radio (Fig. 23-5) that you brought along for company.

Open the back of the radio and find the loopstick antenna. You will need to know which axis it lies along. In the radio shown in Fig. 23-5 the loopstick is along the top of the radio, from left to right. In other radios it is vertical, from top to bottom. Once you know the direction, you can tune in a known AM station and orient the radio until you find a null. Your compass can give you the bearing. If you know the approximate location of the station, then you can reverse the compass direction from it and mark the line on the topo map. Of course, it's still a bidirectional indication, so all you know is the line along which you are lost.

But then you tune into a different station in a different city (or at least wide enough from the line to the other station to make a difference) and take another reading. Your approximate location is where the two lines cross. Take a third, fourth, and fifth reading and you will home in pretty tight. If you were smart enough to plan ahead, you will have selected candidate stations in advance and located their latitude and longitude. Alternatively, you would have bought the topo map that covers their location as well as where you want to hike, and from those maps you can find the latitudes and longitudes of the distant stations.



**23-5** AM portable radio. Dark bar represents position and direction of internal loopstick.

## Regular loop antennas

Regular wire loop antennas (Fig. 23-6) are also used for radio direction finding. In fact, in some cases the regular loop is preferred over the loopstick. The regular loop antenna may be square (as shown), circular, or any other regular “*n*-gon” (e.g., hexagon), although for practical reasons the square is easier to build. The loop has pretty decent inductance even with only a few turns. One loop I built was 24 in square (A in Fig. 23-6) and had, if I recall correctly, about 10 turns of wire spaced over a 1-in width (B in Fig. 23-6). It resonated to the AM BCB with a standard 365-pF “broadcast” variable capacitor.

When you use a regular loop antenna be aware that the antenna has a figure-8 pattern like the loopstick, but it is oriented 90° out of phase with the loopstick antenna. In the regular loop the minima (nulls) are perpendicular to the plane of the loop, while the maxima are off the sides. In Fig. 23-6 the minima are in and out of the page, while the maxima are left and right (or top and bottom).

## Fox hunting

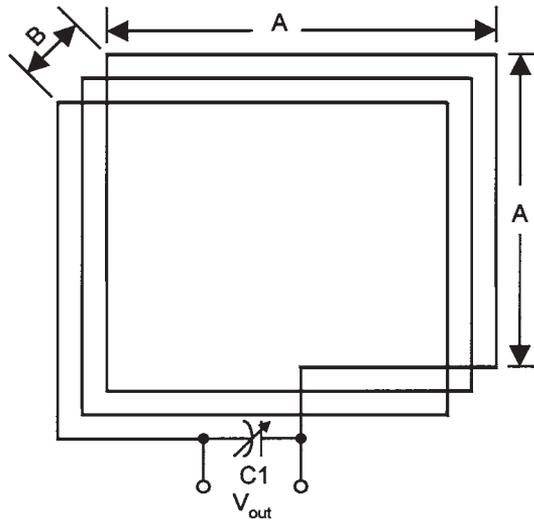
An activity popular with ham radio operators in the 1960s was “fox hunting.” A fiendishly clever ham (the “fox”) would go hide with a mobile or portable transmitter (usually on either 10 or 75 m). The “hunters” would then RDF the fox’s brief transmissions and try to locate the transmitter. If you could locate the antenna, you located the station (according to the most common set of rules). Sometimes the hunt got a little wild as hunters raced each other (a no-no) in the final stretch.

One friend of mine had an interesting experience. He drove a 1949 Plymouth that was painted hideous dark green with fire-engine-red hub caps. It looked just plain awful (especially since the paint looked like it was put on with a whisk broom!). He used a Gonset converter linked to the AM BCB receiver installed in the car. The antenna was a wire loop on the end of a broom handle, similar to Fig. 23-7. The antenna was mounted about where you would expect the left-side rear view mirror to be located. He could reach his hand out the window and rotate the antenna while listening for the minima on the receiver. Worked rather well, and he won some fox hunts.

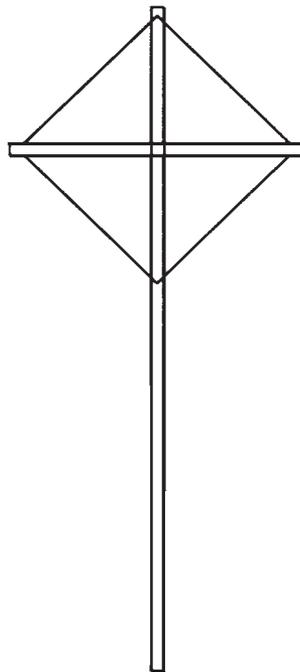
The fox hunts in his area were usually held on Saturday or Sunday morning, and after it was all over the whole crew would get to a restaurant for brunch or beer or something. One morning, however, my buddy with the hideous green car was racing down a residential street to be the first to the transmitter. Others were in the area, and the end game was in motion. Unfortunately, a little old lady called the police complaining about a “...nut racing up and down the street in a strange car waving a cross out the window.” Breakfast had to wait that morning, I bet.

## Shortwave and AM BCB “skip” RDF

Radio direction finding is most accurate over relatively short distances. If you can use the ground wave, then all the better (which is what you use during daylight hours for nearly all AM BCB stations). Skip rolls in on the AM BCB after local



23-6 Wire loop antenna.



23-7 "...waving a cross out the window" antenna.

sundown, so you can hear all manner of stations all up and down the dial. You can RDF distant stations. Some rather interesting sites have turned up when SWL DXers RDF'ed some of the infamous "numbers stations." Those stations transmit either CW or voice numbers groups, and are believed to be sending messages to spies around the world. RDFing by DXers has located some of these transmitters at, well, "interesting" sites.

Unfortunately, there are some problems with skip RDFing. When we look at propagation drawings of skip in textbooks we usually see a one plane view. The curvature of the earth shows, as does the transmitter and receiver site. A "pencil beam" radio signal travels at some angle up the ionosphere, where it is "reflected" (actually, it's a refraction phenomenon but looks like reflection to an observer on the surface) back to earth. We can tell from the drawing that the angle of incidence equals the angle of reflection, just like they told us in high school science classes. Oops! The real world is not so neat and crisp, however.

In the real world the wave might encounter a different ionization density along its path of travel and therefore be deflected from its original path. It might return to earth at a location offset from the direction it appears. If we look at the true bearing from the receiver site to the transmitter site, and then note the azimuthal angle of arrival of the signal, we note that something is amiss.

Actual reflection also causes some problems, especially when RDFing a station in the high end of the HF band or the VHF/UHF bands. Radio waves will reflect from geological features such as mountains, as well as man-made structures (e.g., buildings). If the reflection is strong enough it might appear to be the real signal, and cause a severe error in RDFing. Be wary of RDF results when the "skip is in."

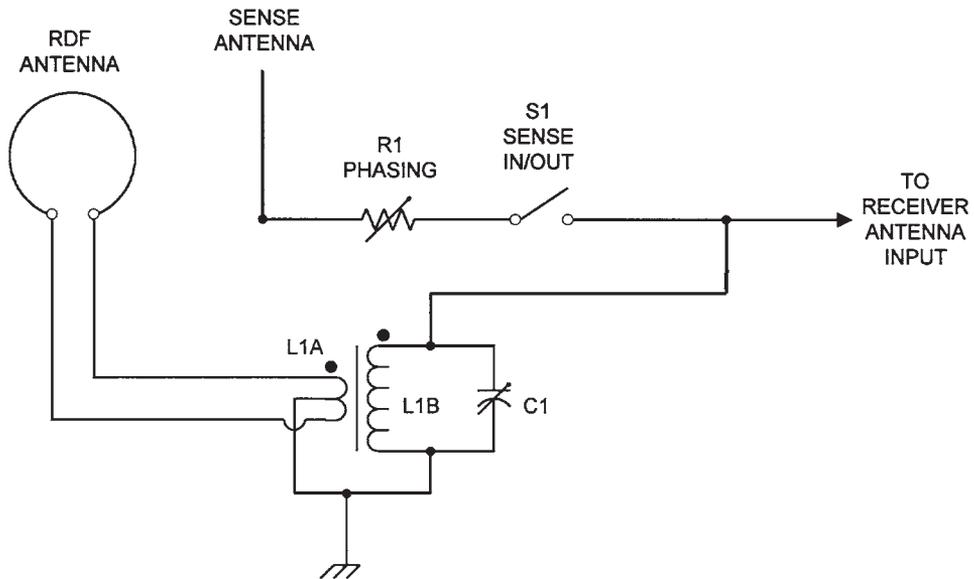
## Sense antenna circuit

Figure 23-8 shows a method for summing together the signals from an RDF antenna (such as a loop) and a sense antenna. The two terminals of the loop are connected to the primary of an RF transformer. This primary ( $L_{1A}$ ) is center-tapped and the center tap is grounded. The secondary of the transformer ( $L_{1B}$ ) is resonated by a variable capacitor  $C_1$ . The dots on the transformer coils indicate the  $90^\circ$  phase points.

The top of  $L_{1B}$  is connected to the sense circuit, and to the receiver antenna input. The phasing control is a potentiometer ( $R_1$ ). The value of this pot is usually 10 to 100 k $\Omega$ , with 25 k $\Omega$  being a commonly seen value. Switch  $S_1$  is used to take the sense antenna out of the circuit. The reason for this switch is that the nulls of the loop or loopstick are typically a lot deeper than the null on the cardioid pattern. The null is first located with the switch open. When the switch is closed you can tell by the receiver S meter whether or not the correct null was used. If not, then reverse the direction of the antenna and try again.

## Adcock antennas

The Adcock antenna has been around since 1919 when it was patented by F. Adcock. Figure 23-9 shows the basic Adcock RDF array. This antenna consists of two centered nonresonant (but identical) vertical radiators. Each side of each element is at least  $0.1\lambda$  long, but need not be resonant (which means the antenna can be used



23-8 Sense circuit for a loop/sense antenna RDF.

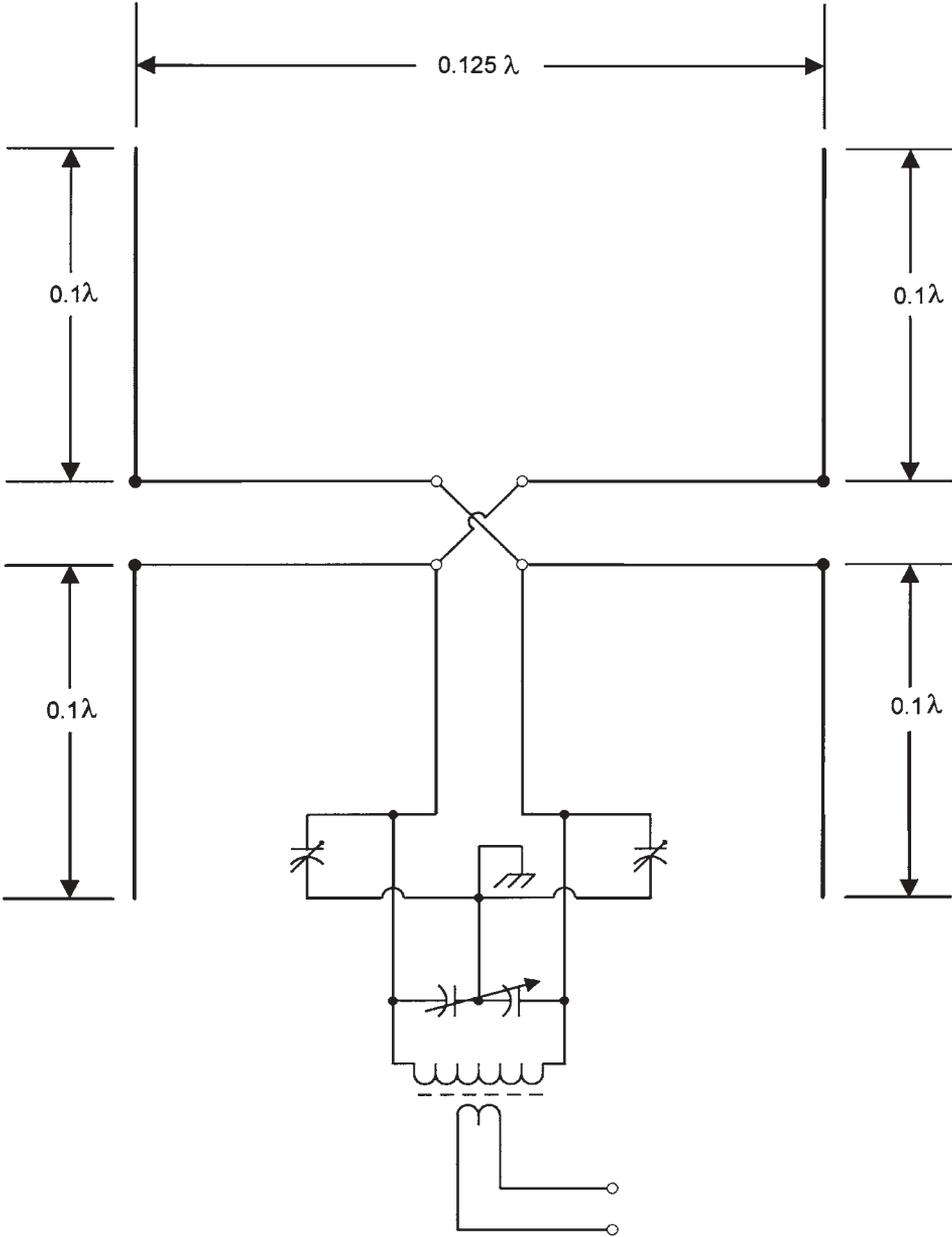
over a wide band). The elements are spaced from  $0.1\lambda$  to  $0.75\lambda$ , although the example shown here is spaced  $0.125\lambda$ .

The Adcock antenna is vertically polarized, so it will respond to the vertically polarized wave very much like loop or phased array antennas of similar size. The horizontally polarized wave, however, affects all elements the same, so the currents are essentially cancelled, resulting in no pattern. This characteristic makes the Adcock antenna suitable for high-frequency shortwave RDFing.

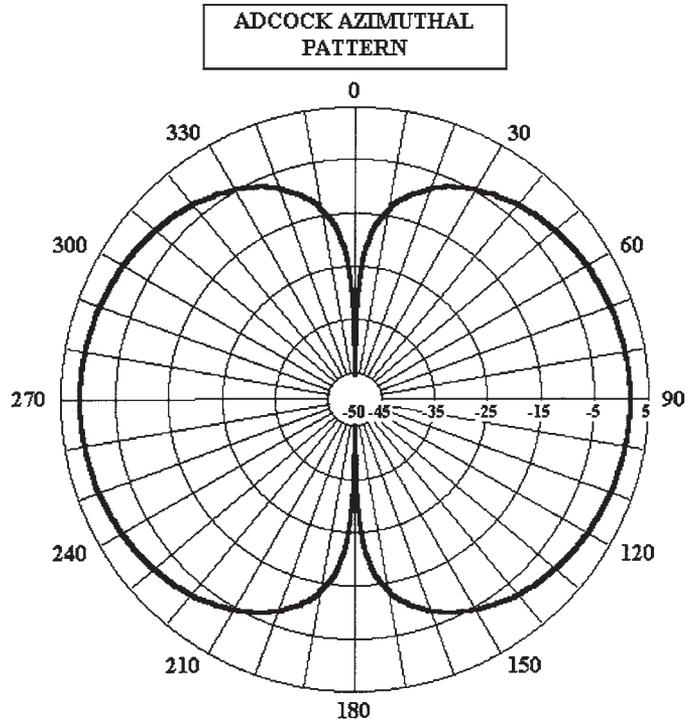
The pattern for an Adcock antenna is shown in Fig. 23-10. This pattern was generated using the NecWin Basic for Windows program (Chap. 17). The example antenna is a 10-MHz (30-m band) Adcock that uses 1.455-m elements (total 2.91 m on each side), spaced 4 m apart. The pattern is a traditional figure-8 with deep nulls at  $0^\circ$  and  $180^\circ$ . The antenna can be rotated to find a null in the same manner as a loop.

## Watson-Watt Adcock array

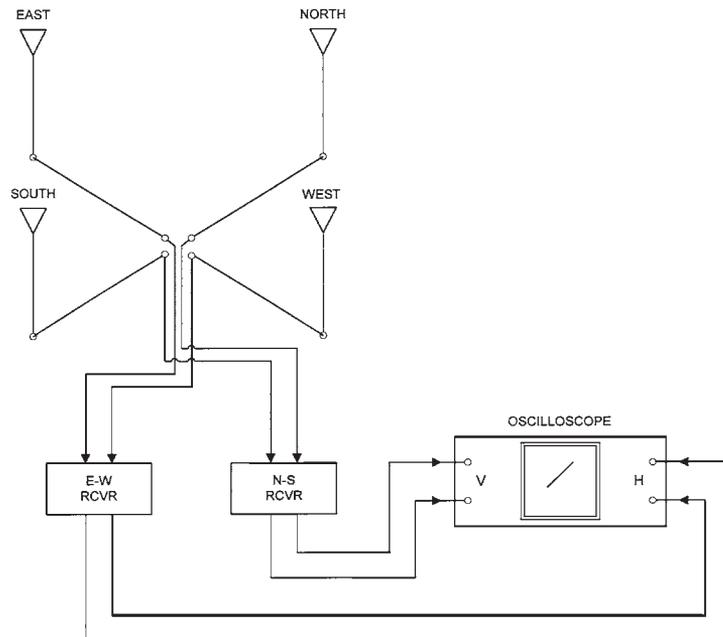
Figure 23-11 shows the Watson-Watt Adcock RDF array. It consists of two Adcock arrays arranged orthogonally to each other. It is common practice to arrange one Adcock in the east-west direction and the other in the north-south direction. These are fed to identical receivers that are controlled by a common local oscillator (LO). The outputs of the receivers are balanced, and are used to drive the vertical and horizontal plates of a cathode-ray oscilloscope (CRO). Figure 23-12 shows the patterns achieved by signals of various phases arriving at the Watson-Watt array. The patterns of Figs. 23-12A and 23-12B are made from signals  $180^\circ$  out of phase, while the signal of Fig. 23-12C has a  $90^\circ$  phase difference.



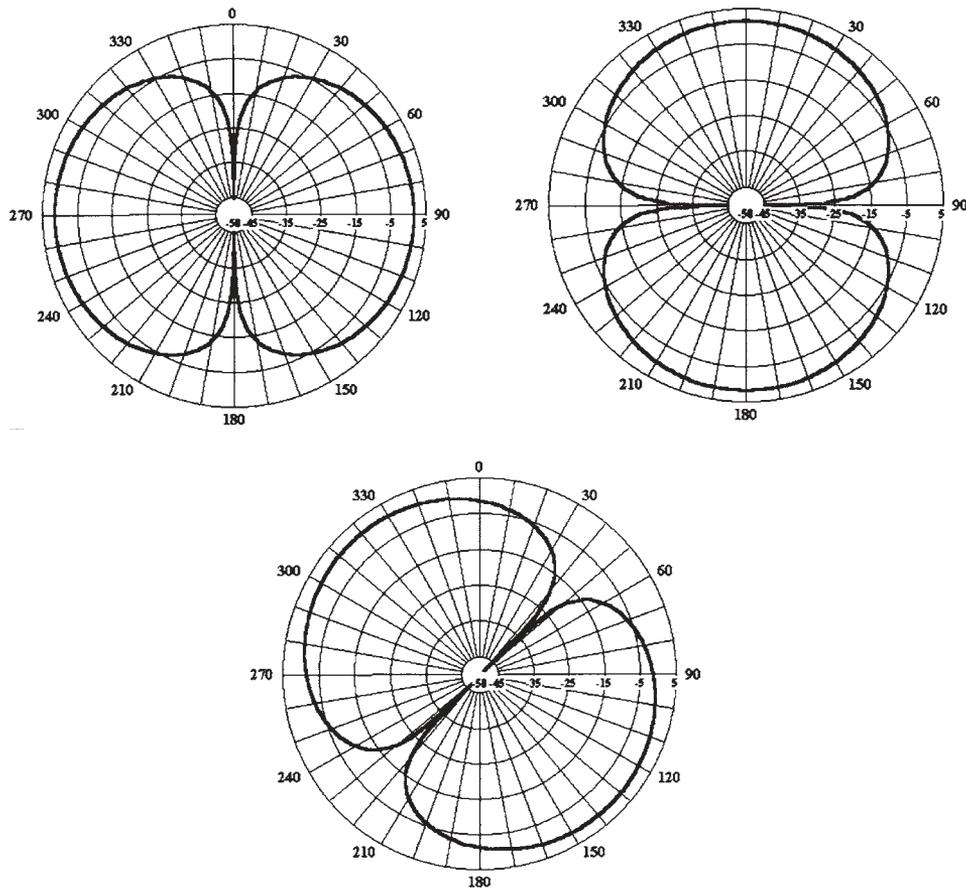
**23-9** Adcock array RDF antenna.



23-10 Adcock pattern.



23-11 Watson-Watt Adcock array.



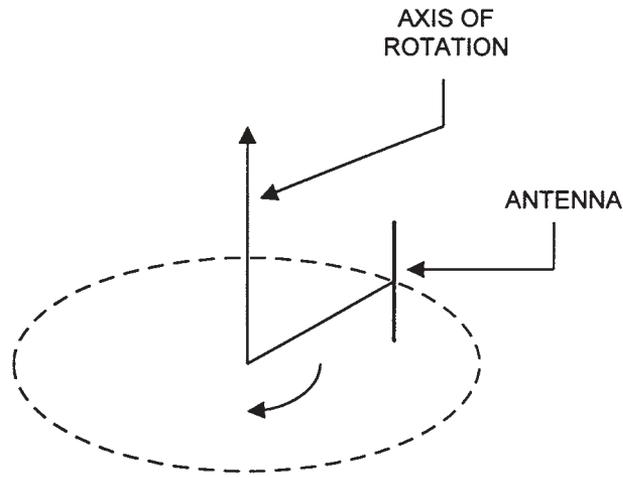
**23-12** Patterns for different signals on the Watson-Watt array.

## Doppler RDF antennas

Figure 23-13 shows the basic concept of a Doppler RDF antenna. The Doppler effect was discovered in the nineteenth century. A practical example of the Doppler effect is seen when a wailing ambulance siren is approaching you, and then passes you and heads away from you. The wailing pitch will rise as the sound source approaches, and then fall as the sound source recedes in the opposite direction.

In a radio system, when the antenna and signal source move with respect to each other, a Doppler shift is generated. This shift is proportional to the relative speed difference. The Doppler RDF antenna of Fig. 23-13 uses a rotating antenna. The signal approaches from a single direction, so there will be a predictable Doppler shift at any point on the circular path of the antenna. The magnitude of the phase shift is

$$S = \frac{R\omega F_c}{c} \quad [23.1]$$



23-13 Doppler antenna.

where

$S$  is the Doppler shift, in hertz

$R$  is the radius of rotation, in meters

$\omega$  is the angular velocity of the antenna, in radians per second

$F_c$  is the carrier frequency of the incoming signal, in hertz

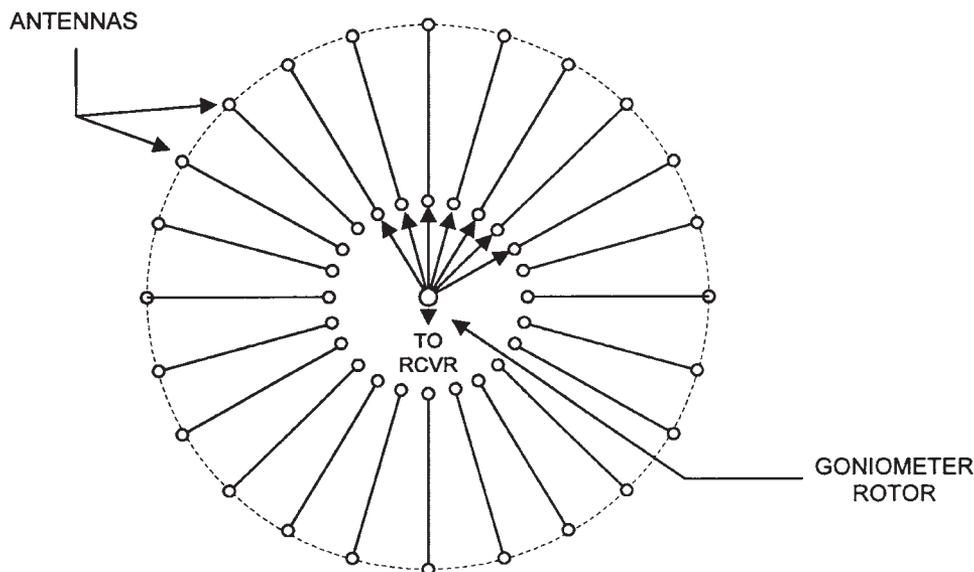
$c$  is the velocity of light ( $3 \times 10^8$  m/s)

In theory this antenna works nicely, but in practice there are problems. One of the big problems is getting a large enough Doppler shift to easily measure. Unfortunately, the rotational speed required of the antenna is very high—too high for practical use. However, the effect can be simulated by using a number of antennas, arranged in a circle, that are sequentially scanned. The result is a piecewise approximation of the effect seen when the antenna is rotated at high speed.

## Wullenweber array

One of the problems associated with small RDF antennas is that they have such a small aperture that relatively large distortions of their pattern result from even small anomalies. Follow that? What it means is that the pattern is all messed up by small defects. If you build a wide-aperture direction finder (WADF), however, you can average the signals from a large number of antenna elements distributed over a large-circumference circle. The Wullenweber array (Fig. 23-14) is such an antenna. It consists of a circle of vertical elements. In the HF band the circle can be 500 to 2000 ft in diameter.

A goniometer rotor spins inside the ring to produce an output that will indicate the direction of arrival of the signal as a function of the position of the goniometer. The theoretical resolution of the Wullenweber array is on the order of  $0.1^\circ$ , although practical resolutions of about  $2.8^\circ$  are commonly seen.



23-14 Wullenweber RDF array.

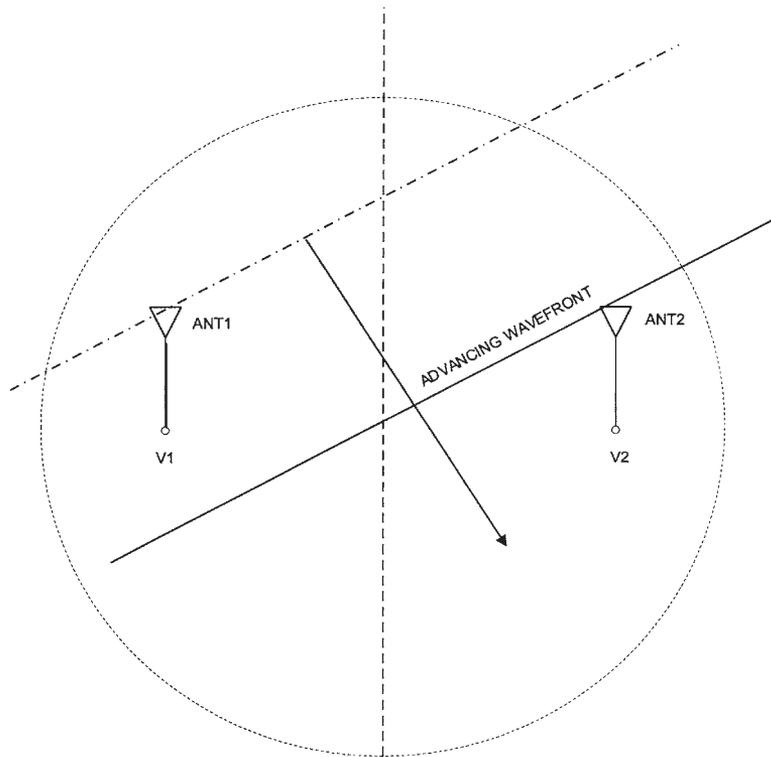
## Time difference of arrival (TDOA) array

If you erect two antennas with distance  $d$  apart, then arriving signals can be detected by examining the time-of-arrival difference. Figure 23-15A shows an example signal. If the advancing wavefront is parallel to the line between the antennas, then it will arrive at both antennas at the same time. The TDOA is zero in that case. But if the signal arrives at an angle (as in Fig. 23-15A), it will arrive at one antenna first. From the difference between the time of arrival at the two antennas we can discern the direction of arrival.

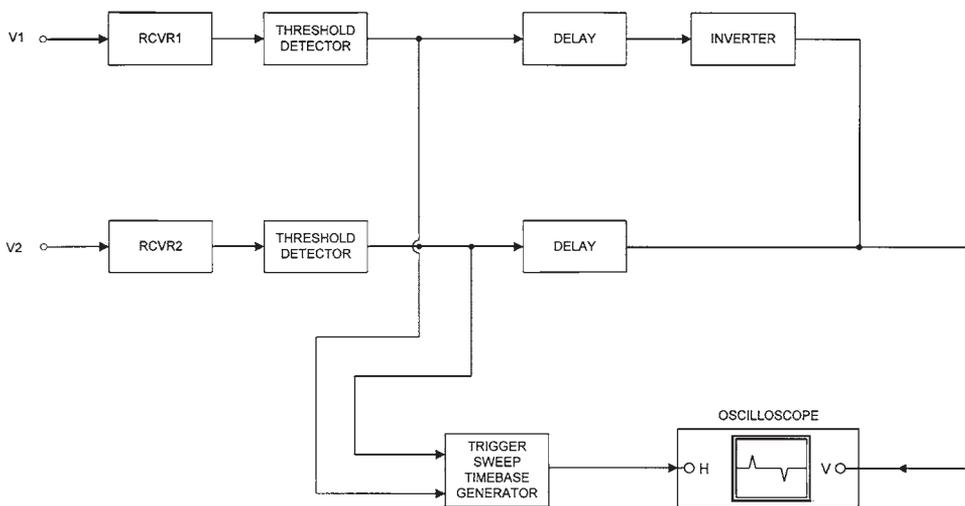
There is an ambiguity in the basic TDOA array in that the combined output will be the same for conjugate angles, i.e. at the same angle from opposite directions. This problem can be resolved by the system shown in Fig. 23-15B. The signals from ANT1 and ANT2 are designated  $V_1$  and  $V_2$ , respectively. These signals are detected by receivers (RCVR1 and RCVR2), and are then threshold detected in order to prevent signal-to-noise problems from interfering with the operation. The outputs of the threshold detectors are used to trigger a sawtooth generator that controls the horizontal sweep on an oscilloscope.

The two signals are then delayed, and one is inverted. The reason for inverting one signal is to allow the operator to distinguish them on the CRT screen. The reason why this is necessary is the ambiguity. If the antennas in Fig. 23-15A are arrayed east to west, then the line perpendicular to the line between them is north to south.

If we designate north as  $0^\circ$ , then the signal shown arrives at an angle of  $330^\circ$ . That means it will arrive at ANT1 before it arrives at ANT2. A signal arriving from a bearing of  $30^\circ$  will produce the same output signal, even though it arrives at ANT2 before ANT1. All signals of bearing  $0 \leq x < 180^\circ$  will arrive at ANT2 first, while all



**23-15A** Time difference of arrival (TDOA) RDF array.



**23-15B** Block diagram of typical TDOA circuit.

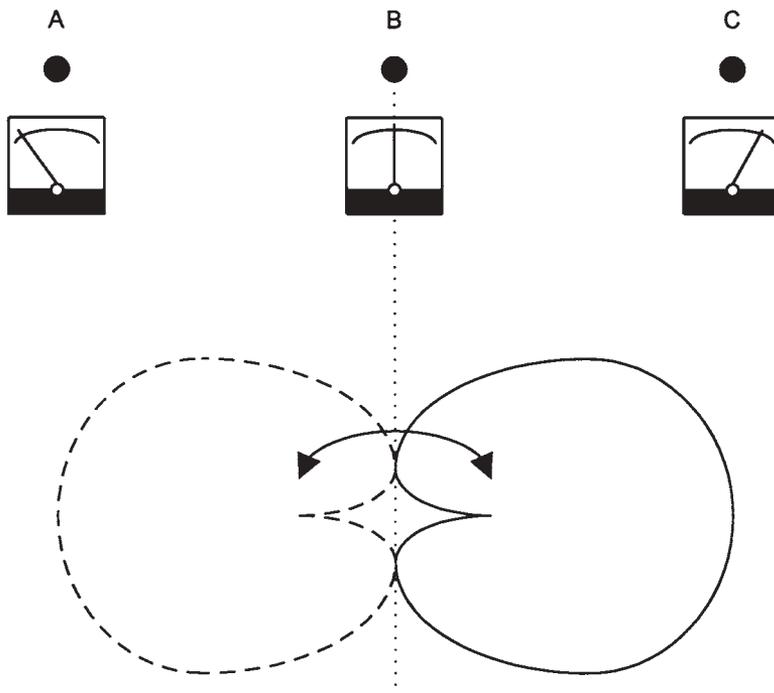
signals  $180 \leq x < 360^\circ$  will arrive at ANT1 first. Yet both will produce the same blip on the oscilloscope screen. The solution to discerning which of the two conjugate angles is intended is to invert the ANT1 signal. When this is done, the ANT1 signal falls below the baseline on the CRT screen, while the ANT2 signal is above the baseline. By noting the time difference between the pulses, and their relative position, we can determine the bearing of the arriving signal.

## Switched-pattern RDF antennas

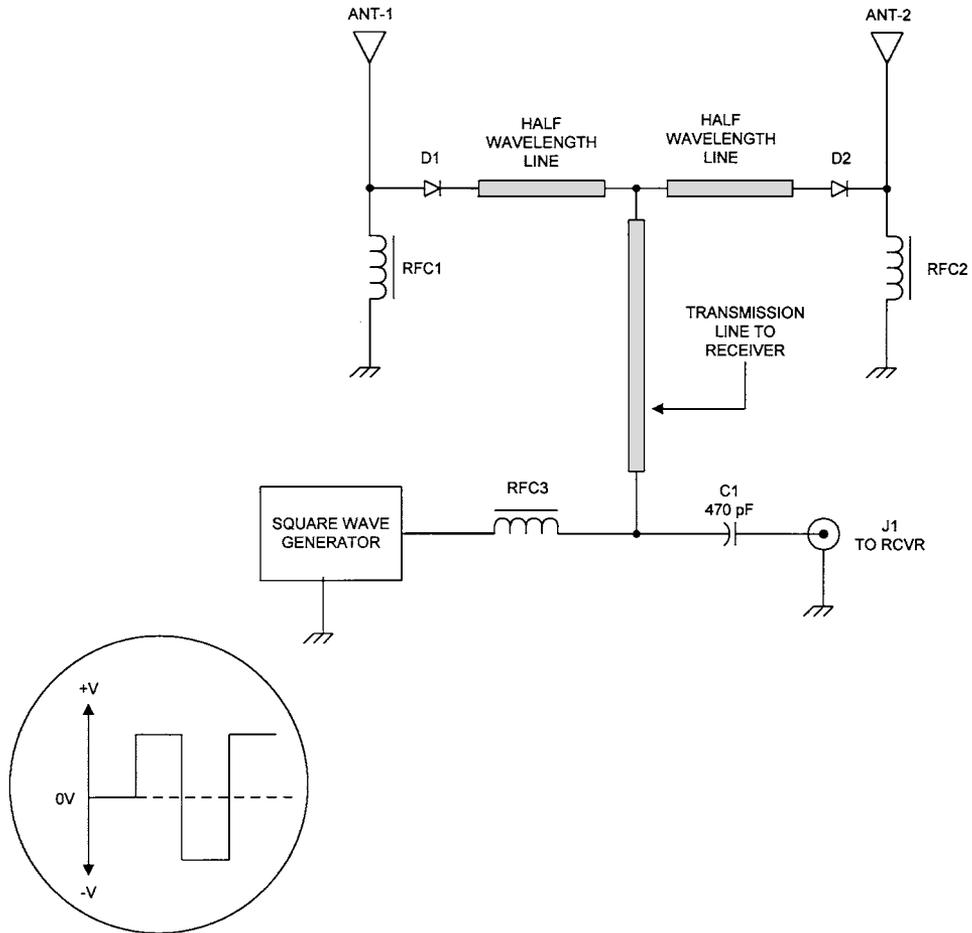
Suppose we have a unidirectional pattern such as the cardioid shown in Fig. 23-16. If we can rapidly switch the pattern back and forth between two directions that are  $180^\circ$  apart, then we can not only discern direction, but also we can tell whether an off-axis signal is left or right. This feature is especially useful for mobile and portable direction finding.

In Fig. 23-16 we have three different positions for a signal source. When the signal source is at point *A*, it will affect the pattern to the left more than the pattern to the right, so the meter will read left. If the signal source is at point *B*, on the other hand, the signal affects both pattern positions equally, so the meter reads zero. Finally, if the signal arrives from point *C*, it affects the right-hand pattern more than the left-hand pattern, so the meter reads right.

Figure 23-17 shows how such a system can be constructed. This system has been used by amateur radio operators with “rubber ducky” VHF antennas and a sin-



**23-16** Switched-pattern RDF system.



**23-17** Double-duddy RDF system based on the switched-pattern system.

gle receiver, where it is commonly called the *double-duddy direction finder* (DDDF). The antennas are spaced from  $0.25\lambda$  to  $1\lambda$  apart over a good groundplane (such as the roof of a car or truck). If no groundplane exists, then a sheet metal groundplane should be provided.

In Fig. 23-17 we see the antennas are fed from a common transmission to the receiver. In order to keep them electrically the same distance apart, a pair of identical half-wavelength sections of transmission line are used to couple to the antennas.

Switching is accomplished by using a bipolar square wave and PIN diodes. The bipolar square wave (see inset to Fig. 23-17) has a positive peak voltage and a negative peak voltage on opposite halves of the cycle. The PIN diodes ( $D_1$  and  $D_2$ ) are connected in opposite polarity to each other. Diode  $D_1$  will conduct on negative excursions of the square wave, while  $D_2$  conducts on positive excursions. The antenna connected to the conducting diode is the one that is connected to the receiver, while the other one is parasitic. The active antenna therefore switches back and forth between ANT1 and ANT2.

This antenna is coupled to the receiver through a small-value capacitor ( $C_1$ ) so that the square wave does not enter the receiver. This allows us to use the transmission line for both the RF signal and the switching signal. An RF choke (RFC<sub>3</sub>) is used to keep RF from the antenna from entering the square-wave generator.

The DDDF antenna produces a phase modulation of the incoming signal that has the same frequency as the square wave. This signal can be heard in the receiver output. When the signal's direction of arrival is perpendicular to the line between the two antennas, the phase difference is zero, so the audio tone disappears.

The pattern of the DDDF antenna is bidirectional, so there is the same ambiguity problem as exists with loop antennas. The ambiguity can be resolved by either of two methods. First, a reflector can be placed  $\lambda/4$  behind the antennas. This is attractive, but it tends to distort the antenna pattern a little bit. The other method is to rotate the antenna through  $90^\circ$  (or walk an L-shaped path).

## Conclusion

Radio direction finding can be very useful for locating RF noise sources, illegal stations, and other RF sources. It can also be used to locate yourself if you can get bearings on at least two stations. The lines of bearing will cross at a location approximately where you are standing—try it, you'll like it.

## 24 CHAPTER

# Impedance matching in antenna systems

ONE OF THE FIRST THINGS THAT YOU LEARN IN RADIO COMMUNICATIONS AND broadcasting is that antenna impedance must be matched to the transmission line impedance, and that the transmission line impedance must be matched to the output impedance of the transmitter. The reason for this requirement is that maximum power transfer, between a source and a load, always occurs when the system impedances are matched. In other words, more power is transmitted from the system when the load impedance (the antenna), the transmission line impedance, and the transmitter output impedance are all matched to each other.

Of course, the trivial case is where all three sections of our system have the same impedance. For example, we could have an antenna with a simple  $75\text{-}\Omega$  resistive feedpoint impedance (typical of a half-wave dipole in free space), and a transmitter with an output impedance that will match  $75\ \Omega$ . In that case, we need only connect a standard-impedance  $75\text{-}\Omega$  length of coaxial cable between the transmitter and the antenna. Job done! Or so it seems.

But there are other cases where the job is not so simple. In the case of our “standard” antenna, for example, the feedpoint impedance is rarely what the books say it should be. That ubiquitous dipole, for example, is nominally rated at  $73\ \Omega$ , but even the simplest antenna books tell us that value is an approximation of the theoretical free-space impedance. At locations closer to the earth’s surface, the impedance could vary over the approximate range of  $30$  to  $130\ \Omega$  and it might have a substantial reactive component.

But there is a way out of this situation. We can construct an impedance-matching system that will marry the source impedance to the load impedance. This chapter examines several matching systems that might prove useful in a number of antenna situations.

## Impedance-matching approaches

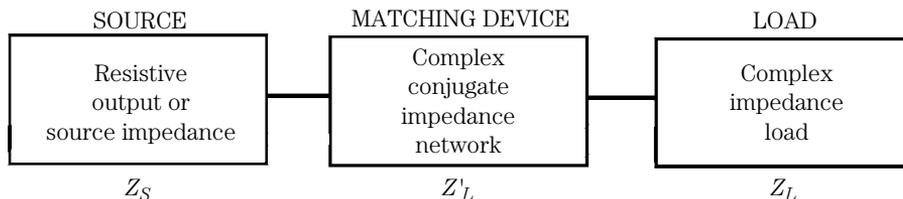
Antenna impedance can contain both reactive and resistive components. In most practical applications, we are searching for a purely resistive impedance ( $Z = R$ ), but that ideal is rarely achieved. A dipole antenna, for example, has a theoretical free-space impedance of  $73 \Omega$  at resonance. As the frequency applied to the dipole is varied away from resonance, however, a reactive component appears. When the frequency is greater than resonance, then the antenna tends to look like an inductive reactance, so the impedance is  $Z = R + jX$ . Similarly, when the frequency is less than the resonance frequency, the antenna looks like a capacitive reactance, so the impedance is  $Z = R - jX$ . Also, at distances closer to the earth's surface the resistive component may not be exactly  $73 \Omega$ , but may vary from about 30 to  $130 \Omega$ . Clearly, whatever impedance coaxial cable is selected to feed the dipole, it stands a good chance of being wrong.

The method used for matching a complex load impedance (such as an antenna) to a resistive source (the most frequently encountered situation in practical radio work) is to interpose a matching network between the load and the source (Fig. 24-1). The matching network must have an impedance that is the *complex conjugate* of the complex load impedance. For example, if the load impedance is  $R + jX$ , then the matching network must have an impedance of  $R - jX$ ; similarly, if the load is  $R - jX$  then the matching network must be  $R + jX$ . The sections to follow look at some of the more popular networks that accomplish this job.

### L-section network

The L-section network is one of the most used, or at least the most published, antenna matching networks in existence: it rivals even the pi network. A circuit for the L-section network is shown in Fig. 24-2A. The two resistors represent the source ( $R_1$ ) and load ( $R_2$ ) impedances. The elementary assumption of this network is that  $R_1 < R_2$ . The design equations are

$$R_1 < R_2 \text{ and } 1 < Q < 5 \quad [24.1]$$



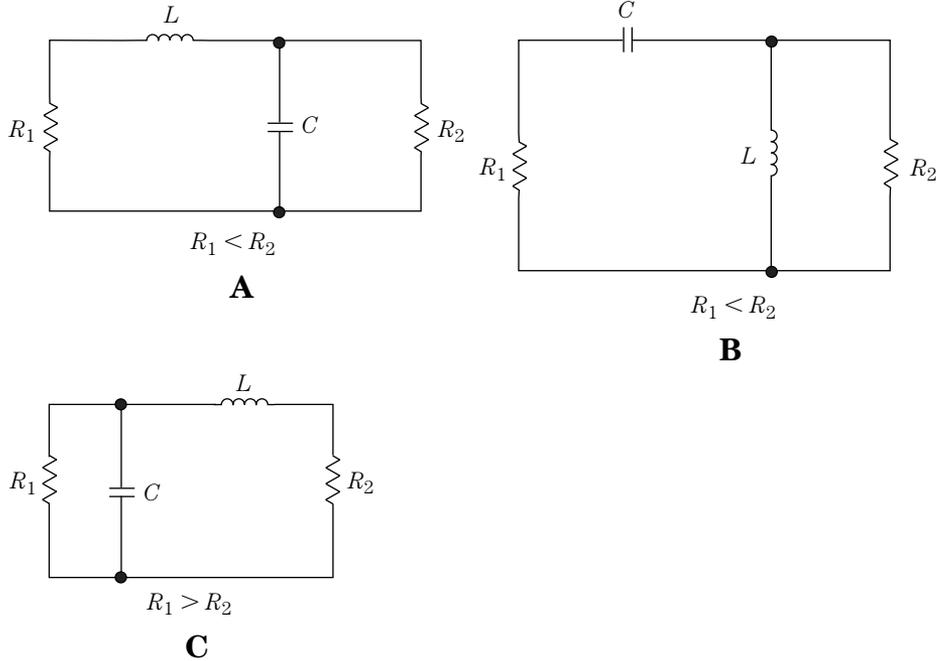
$$Z_S = R$$

If:

$$Z_L = R + jX, \text{ then } Z'_L = R - jX$$

$$Z_L = R - jX, \text{ then } Z'_L = R + jX$$

24-1 Transmitter, matcher, and antenna system.



**24-2** (A) L-section network; (B) reverse L section; (C) inverted L-section network.

$$X_L = 6.28FL = Q \times R_1 \quad [24.2]$$

$$X_C = \frac{1}{6.28FC} \quad [24.3]$$

$$Q = \sqrt{\frac{R_2}{R_1} - 1} \quad [24.4]$$

also,

$$Q = \frac{X_L}{R_1} = \frac{R_2}{X_C} \quad [24.5]$$

You will most often see this network published in conjunction with less-than-quarter-wavelength longwire antennas. One common fault of those books and articles is that they typically call for a “good ground” in order to make the antennas work properly. But they don’t tell you (1) what a “good ground” is and (2) how you can obtain it. Unfortunately, at most locations a good ground means burying a lot of copper conductor—something that most of us cannot afford. In addition, it is often the case that the person who is forced to use a longwire, instead of a better antenna, cannot construct a “good ground” under any circumstances because of

landlords (and/or logistics) problems. The very factors that prompt the use of a longwire antenna in the first place also prohibit any form of practically obtainable “good ground.” But there is a way out: *radials*. A good ground can be simulated with a counterpoise ground constructed of quarter-wavelength radials. These radials have a length in feet equal to  $246/F_{\text{MHz}}$ , and as few as two of them will work wonders. Another form of L-section network is shown in Fig. 24-2B. This circuit differs from the previous circuit in that the roles of the L and C components are reversed. As you might suspect, this role reversal brings about a reversal of the impedance relationships: in this circuit the assumption is that the driving source impedance  $R_1$  is larger than the load impedance  $R_2$  (i.e.,  $R_1 > R_2$ ). The equations are shown below:

$$R_2 > R_1 \quad [24.6]$$

$$X_L = R_2 \sqrt{\frac{R_1}{R_2 - R_1}} \quad [24.7]$$

$$X_C = \frac{(R_1 R_2)}{X_L} \quad [24.8]$$

Still another form of L-section network is shown in Fig. 24-2C. Again, we are assuming that driving source impedance  $R_1$  is larger than load impedance  $R_2$  (i.e.,  $R_1 > R_2$ ). In this circuit, the elements are arranged similar to Fig. 24-2A, with the exception that the capacitor is at the input rather than the output of the network. The equations governing this network are:

$$R_1 > R_2 \text{ and } 1 < Q < 5 \quad [24.9]$$

$$X_L = 6.28FL = \sqrt{(R_1 R_2) - (R_2)^2} \quad [24.10]$$

$$X_C = \frac{1}{6.28FC} = \frac{R_1 R_2}{X_L} \quad [24.11]$$

$$C = \frac{1}{6.28FX_C} \quad [24.12]$$

$$L = \frac{X_L}{6.28F} \quad [24.13]$$

Thus far, we have considered only matching networks that are based on inductor and capacitor circuits. But there is also a possibility of using transmission line segments as impedance-matching devices. Two basic forms are available to us: quarter-wave sections and the series matching section.

### Pi networks

The pi network shown in Fig. 24-3 is used to match a high source impedance to a low load impedance. These circuits are typically used in vacuum tube RF power amplifiers that need to match low antenna impedances. The name of the circuit comes from its resemblance to the Greek letter pi. The equations for the pi network are:

$$R_1 > R_2 \text{ and } 5 < Q < 15 \quad [24.14]$$

$$Q > \sqrt{\frac{R_1}{R_2}} - 1 \quad [24.15]$$

$$X_{C2} = \frac{R_2}{\sqrt{\left(\frac{R_2}{R_1}\right) (1 + Q^2) - 1}} \quad [24.16]$$

$$X_{C1} = \frac{R_1}{Q}$$

$$X_L = \frac{R_1 \{(Q + (R_2/X_{C2}))\}}{Q^2 + 1} \quad [24.17]$$

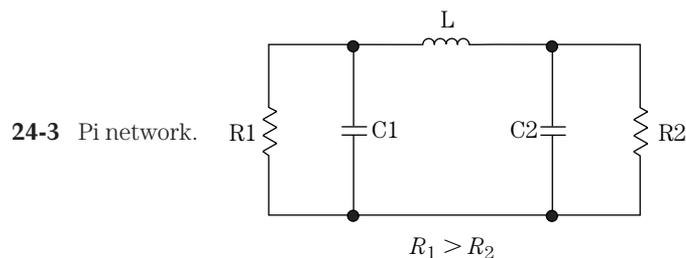
### Split-capacitor network

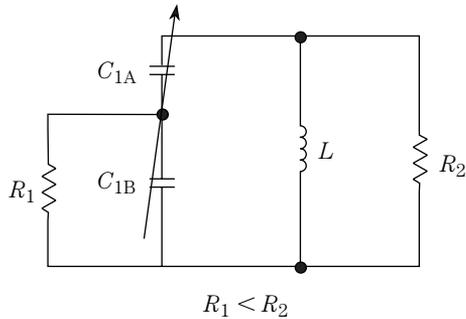
The split-capacitor network shown in Fig. 24-4 is used to transform a source impedance that is less than the load impedance. In addition to matching antennas, this circuit is also used for interstage impedance matching inside communications equipment. The equations for design are

$$R_1 < R_2 \quad [24.18]$$

$$Q > \sqrt{\frac{R_1}{R_2}} - 1 \quad [24.19]$$

$$X_L = \frac{R_2}{Q} \quad [24.20]$$





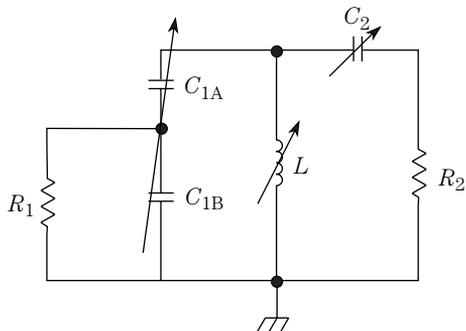
24-4 Split-capacitor network.

$$X_{C1B} = \sqrt{\frac{R_1 (Q_2 + 1)^2}{R_2}} - 1 \quad [24.21]$$

$$X_{C1A} = \frac{R_2 Q}{Q^2 + 1} \left( 1 - \frac{R_1}{Q X_{C1A}} \right) \quad [24.22]$$

## Transmatch circuit

One version of the *transmatch* is shown in Fig. 24-5. This circuit is basically a combination of the split-capacitor network and an output tuning capacitor ( $C_2$ ). For the HF bands, the capacitors are on the order of 150 pF per section for  $C_1$ , and 250 pF for  $C_2$ . The collar inductor should be 28  $\mu$ H. The transmatch is essentially a coax-to-coax impedance matcher, and is used to trim the mismatch from a line before it affects the transmitter.



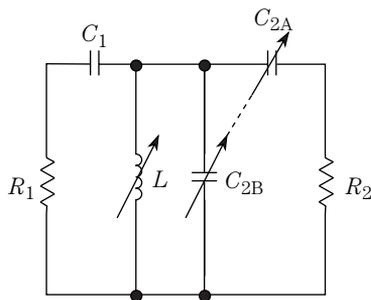
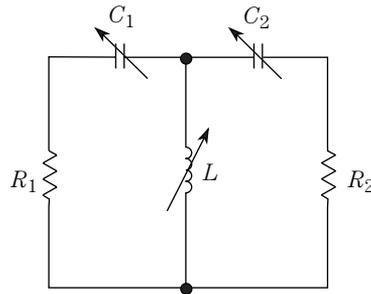
24-5 Split-capacitor transmatch network.

Perhaps the most common form of transmatch circuit is the tee network shown in Fig. 24-6. This network is lower in cost than some of the others, but suffers a problem. Although it does, in fact, match impedance (and thereby, in a naive sense, “tunes out” VSWR on coaxial lines), it also suffers a high pass characteristic. The network, therefore, does not reduce the harmonic output of the transmitter. The simple tee network does not serve one of the main purposes of the antenna tuner: harmonic reduction.

An alternative network, called the *SPC transmatch*, is shown in Fig. 24-7. This version of the circuit offers harmonic attenuation, as well as matching impedance.

Figure 24-8 shows commercially available antenna tuners based on this transmatch design. The unit shown in Fig. 24-8A is manufactured by MFJ Electronics, Inc. It contains the usual three tuning controls, here labeled *transmitter*, *antenna*, and *inductor*. Included in this instrument is an antenna selector switch that allows the operator to select a coax antenna through the tuner, to connect input to output (coax) without regard to the tuner, select a balanced antenna, or connect an internal dummy load. The instrument also contains a multifunction meter that can measure 200 W or 2000 W (full scale), in either forward or reverse directions. In addition, the meter operates as a VSWR meter.

24-6 Tee-network transmatch.



24-7 Improved transmatch offers harmonic attenuation.

Figures 24-8B through 24-8D show an imported tuner from the United Kingdom. This instrument, called the *Nevada* model, is a low-cost model, but contains the three basic controls. For proper operation, an external RF power meter, or VSWR meter, is required. The tuner is shown in Fig. 24-8B; also shown is a Heathkit transmatch antenna tuner. The rear panel of the Nevada instrument is shown in Fig. 24-8C. There are SO-239 coaxial connectors for input and unbalanced output, along with a pair of posts for the parallel line output. A three-post panel is used to select which antenna the RF goes to: unbalanced (coax) or parallel. The internal circuitry of the *Nevada* is shown in Fig. 24-8D. The roller inductor is in the center and allows the user to set the tuner to a wide range of impedances over the entire 3- to 30-MHz HF band.

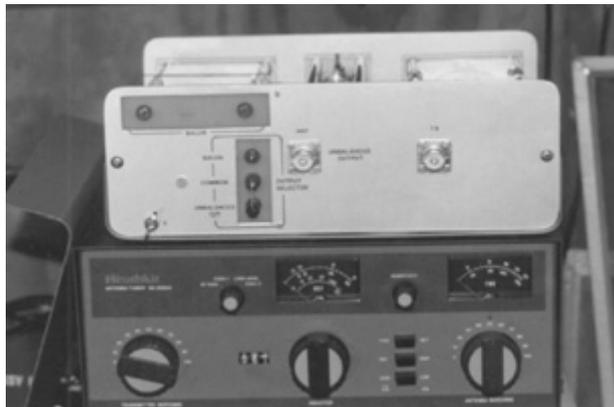
24-8A MFJ antenna tuner.



24-8B Author's Heathkit antenna tuner and the Nevada tuner from the United Kingdom.



24-8C Rear of Nevada tuner.





24-8D Inside of Nevada tuner.

### Coaxial cable BALUNS

A balun is a transformer that matches an unbalanced resistive source impedance (such as a coaxial cable), to a “balanced” load (such as a dipole antenna). With the circuit of Fig. 24-9, we can make a balun that will transform impedance at a 4:1 ratio, with  $R_2 = 4 \times R_1$ . The length of the balun section of coaxial cable is

$$L_{ft} = \frac{492V}{F_{\text{MHz}}} \quad [24.23]$$

where:

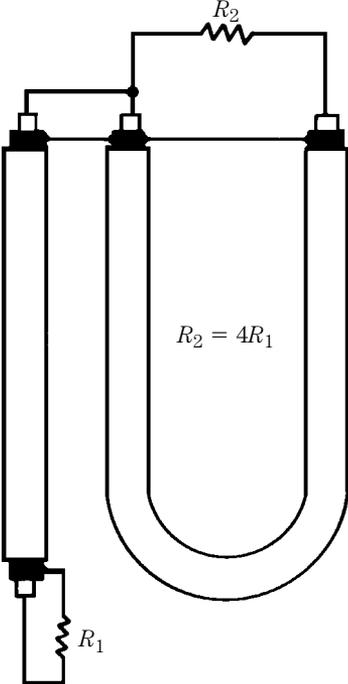
$L_{ft}$  is the length, in feet

$V$  is the velocity factor of the coaxial cable (a decimal fraction)

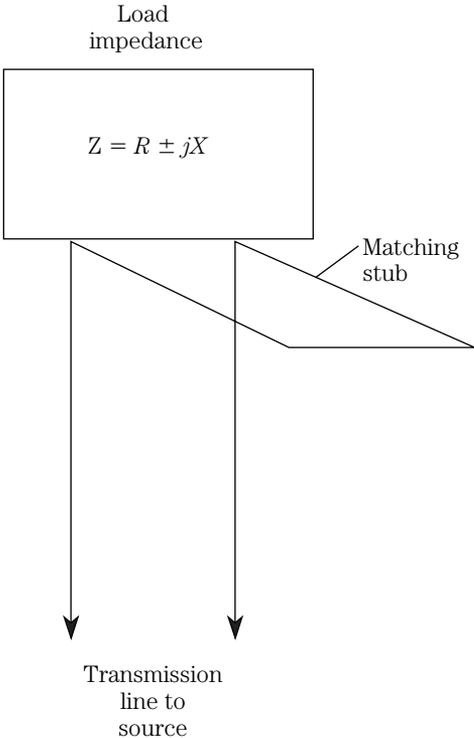
$F_{\text{MHz}}$  is the operating frequency, in megahertz

### Matching stubs

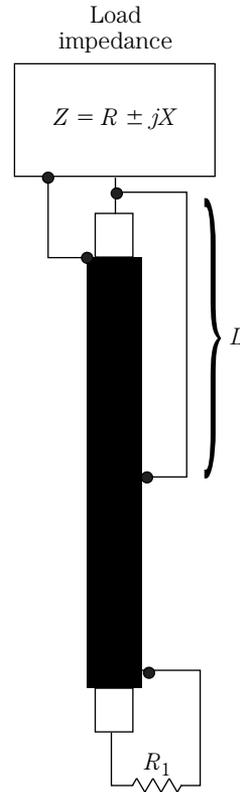
A shorted stub can be built to produce almost any value of reactance. This fact can be used to make an impedance-matching device that cancels the reactive portion of a complex impedance. If we have an impedance of  $Z = R + j30 \Omega$  we need to make a stub with a reactance of  $-j30 \Omega$  to match it. Two forms of matching stub are shown in Figs. 24-10A and 24-10B. These stubs are connected exactly at the feedpoint of the complex load impedance, although they are sometimes placed farther back on the line at a (perhaps) more convenient point. In that case, however, the reactance required will be transformed by the transmission line between the load and the stub.



24-9 Coaxial balun transformer.



24-10A Stub match scheme.



24-10B Coaxial stub matching.

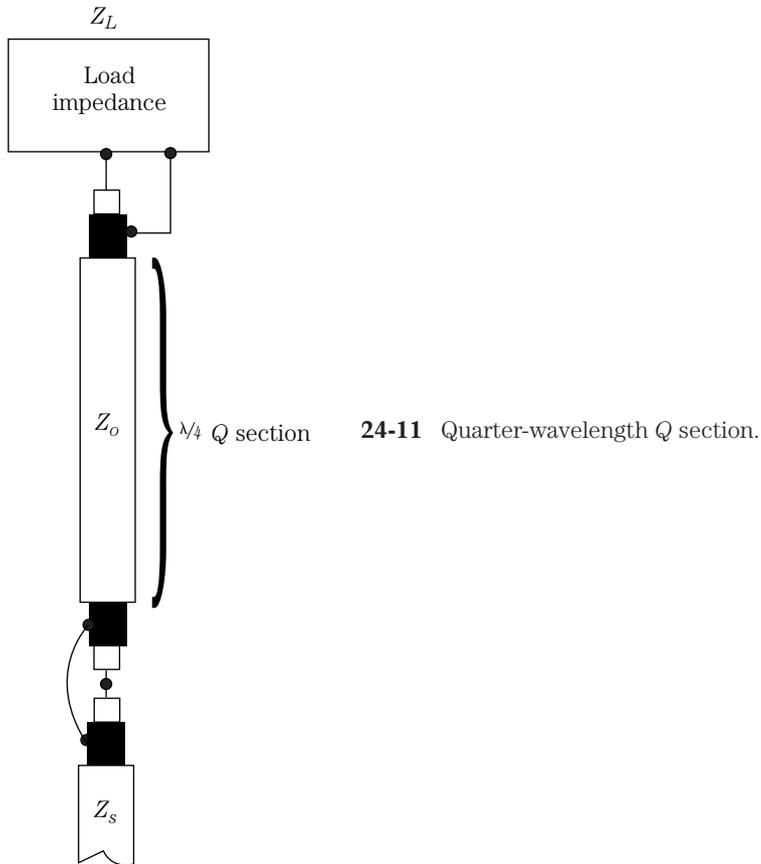
### Quarter-wave matching sections

Figure 24-11 shows the elementary quarter-wavelength transformer section connected between the transmission line and the antenna load. This transformer is also sometimes called a *Q section*. When designed correctly, this transmission line transformer is capable of matching the normal feedline impedance  $Z_s$  to the antenna feedpoint impedance  $Z_R$ . The key factor is to have available a piece of transmission line that has an impedance  $Z_o$  of

$$Z_o = \sqrt{Z_s Z_R} \quad [24.24]$$

Most texts show this circuit for use with coaxial cable. Although it is certainly possible, and even practical in some cases, for the most part there is a serious flaw in using coax for this project. It seems that the normal range of antenna feedpoint impedances, coupled with the rigidly fixed values of coaxial-cable surge impedance available on the market, combines to yield unavailable values of  $Z_o$ . Although there are certainly situations that yield to this requirement, many times the quarter-wave section is not usable on coaxial-cable antenna systems having standard impedance values.

On parallel transmission line systems, on the other hand, it is quite easy to achieve the correct impedance for the matching section. We use the equation above to find a value for  $Z_o$ , and then calculate the dimensions of the parallel feeders. Be-



cause we know the impedance, and can more often than not select the conductor diameter from available wire supplies, you can use the equation below to calculate conductor spacing:

$$S = D \times 10^{(Z/276)} \quad \text{[24.25]}$$

where

$S$  is the spacing

$D$  is the conductor diameter

$Z$  is the desired surge impedance.

From there you can calculate the length of the quarter-wave section from the familiar  $246/F_{\text{MHz}}$ .  $D$  and  $S$  are in the same units.

### Series matching section

The quarter-wavelength section, covered in the preceding section, suffers from drawbacks: it must be a quarter-wavelength and it must use a specified (often non-standard) value of impedance. The series matching section is a generalized case of

the same idea, and it permits us to build an impedance transformer that overcomes these faults. According to *The ARRL Antenna Book*, this form of transformer is capable of matching any load resistance between about 5 and 1200 Ω. In addition, the transformer section is not located at the antenna feedpoint.

Figure 24-12 shows the basic form of the series matching section. There are three lengths of coaxial cable:  $L_1$ ,  $L_2$ , and the line to the transmitter. Length  $L_1$  and the line to the transmitter (which is any convenient length) have the same characteristic impedance, usually 75 Ω. Section  $L_2$  has a different impedance from  $L_1$  and the line to the transmitter. Note that only standard, easily obtainable values of impedance are used here.

The design of this transformer consists of finding the correct lengths for  $L_1$  and  $L_2$ . You must know the characteristic impedance of the two lines (50 Ω and 75 Ω given as examples), and the complex antenna impedance. In the case where the antenna is non resonant, this impedance is of the form  $Z = R \pm jX$ , where  $R$  is the resistive portion,  $X$  is the reactive portion (inductive or capacitive), and  $j$  is the so-called imaginary operator (i.e., square root of minus one). If the antenna is resonant, then  $X = 0$ , and the impedance is simply  $R$ .

The first chore in designing the transformer is to *normalize* the impedances:

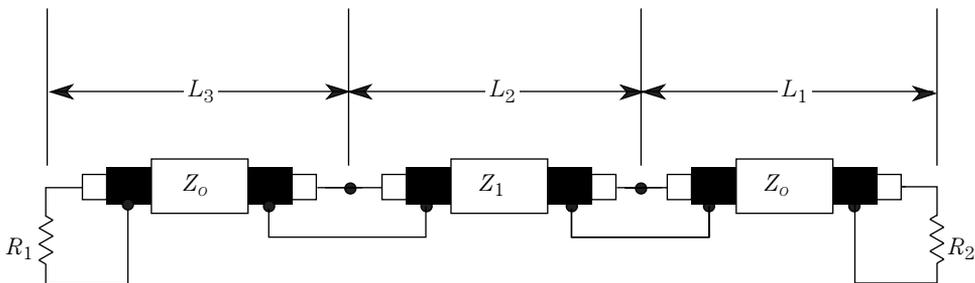
$$N = \frac{Z_{L1}}{Z_o} \tag{24.26}$$

$$R = \frac{R_L}{Z_o} \tag{24.27}$$

$$X = \frac{X_L}{Z_o} \tag{24.28}$$

The lengths are determined in electrical degrees, and from that determination we can find length in feet or meters. If we adopt ARRL notation and define  $A = \tan L_1$ , and  $B = \tan L_2$ , then the following equations can be written:

If:  $Z_L = R \pm jX_L$



24-12 Series matching section.

$$\tan L_2 = B = \frac{(r-1)^2 + X^2}{r\{N - (1/N)^2\} - (r-1)^2 - X^2} \quad [24.29]$$

$$\tan L_1 = A = \frac{\{N - (r/R)\} \times B + X}{r + (XNB) - 1} \quad [24.30]$$

where:

$$N = \frac{Z_1}{Z_0}$$

$$r = \frac{R_L}{Z_0}$$

$$Z = \frac{X_L}{Z_0}$$

Constraints:

$$Z_1 > Z_0 \sqrt{\text{VSWR}} \quad [24.31]$$

$$Z_1 < Z_0 \sqrt{\text{VSWR}} \quad [24.32]$$

If  $L_1 < 0$ , then add  $180^\circ$

If  $B < 0$ , then  $Z_1$  is too close to  $Z_0$

$Z_1$  not equal to  $Z_0$

$$Z_0 \sqrt{\text{VSWR}} < Z_1 < Z_0 \sqrt{\text{VSWR}}$$

Physical length in feet:

$$L_1 = \frac{L_1 \lambda}{360} \quad [24.33]$$

$$L_2 = \frac{L_2 \lambda}{360} \quad [24.34]$$

where

$$\lambda = \frac{984 \times \text{velocity factor}}{\text{frequency in megahertz}} \quad [24.35]$$

The physical length is determined from  $\arctan A$  and  $\arctan B$ , divided by 360 and multiplied by the wavelength along the line and the velocity factor.

Although the sign of  $B$  can be selected as either  $-$  or  $+$ , the use of  $+$  is preferred because a shorter section is obtained. In the event that the sign of  $A$  turns out negative, add  $180^\circ$  to the result.

There are constraints on the design of this transformer. For one thing, the impedance of the two sections  $L_1$  and  $L_2$  cannot be too close together. In general, the following relationships must be observed:

Either

$$Z_{L1} > Z_o \times \text{SWR} \quad [24.36]$$

or

$$\frac{Z_{L1} < Z_o}{\text{SWR}} \quad [24.37]$$

## Ferrite core inductors

The word *ferrite* refers to any of several examples of a class of materials that behave similarly to powdered iron compounds, and are used in radio equipment in the form of inductors and transformers. Although the original materials were of powdered iron, and indeed the name *ferrite* still implies iron, many modern materials are composed of other compounds. According to the Amidon Associates literature, ferrites with a permeability  $\mu$  of 800 to 5000 are generally of the manganese-zinc type of material, and cores with permeabilities of 20 to 800 are of nickel-zinc. The latter are useful in the 0.5- to 100-MHz range.

### Toroid cores

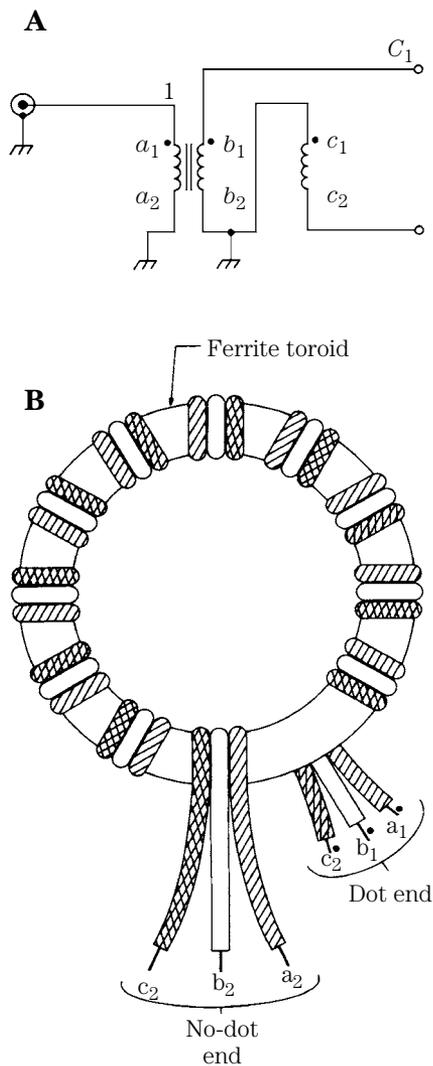
A *toroid* is a doughnut-shaped object, so one can reasonably expect a toroidal core to be an inductor (or transformer) form, made of a ferrite material, in the general shape of a doughnut. The type of core must be known for application purposes, and is given by the type number. The number will be of the form: FT –  $xx$  –  $nn$ , where FT means ferrite toroid and describes the shape,  $xx$  indicates the size, and  $nn$  indicates the material type. The F in FT is sometimes deleted in parts lists, and the core defined as a “T- $xx$ - $nn$ .”

A chart is provided by Amidon that gives the dimensions, and a description of the properties of the different types of material, along with a lot of other physical data. Some of these data are also available in the annual ARRL publication *The ARRL Handbook for the Radio Amateur* (the same material has appeared in earlier editions also).

Tables are derived in part from both Amidon and ARRL sources. From these tables, you can see the sizes and properties of various popular toroids. These tables, incidentally, are not exhaustive of either the variety of toroids available, or all of the properties of the toroids mentioned. Using the nomenclature mentioned, a T-50-2 core refers to a core that is useful from 1 to 30 MHz, has a permeability of 10, is painted red, and has the following dimensions: OD = 0.500 in (1.27 cm), ID = 0.281 in (0.714 cm) and a height (i.e., thickness) of 0.188 in (0.448 cm).

### Toroidal transformers

A magazine reader once asked a question of this author concerning the winding protocol for toroidal transformers as seen in textbooks and magazine articles. My correspondent included a partial circuit (Fig. 24-13A) as typical of the dilemma. The question was “How do you wind it?” and a couple of alternative methods were proposed. At first I thought it was a silly question because the answer was “obvious,”



24-13 Broadband RF transformer.

and then I realized that perhaps I was wrong, and to many people the answer was not at all that obvious.

The answer to the question is that all windings are wound together in a “multifilar” manner. Because there are three windings, in this case we are talking about “trifilar” windings. Figure 24-13B shows the trifilar winding method. For the sake of clarity, I have patterned all three wires differently so that you can follow them. This practice is also a good idea for practical situations. Since most small construction projects use no.26, no.28 or no.30 enameled wire to wind coils, I keep three colors of each size on hand, and wind each winding with a different color [Note: for transmitting antenna transformers use no.16, no.14, no.12, or no.10 wire]. Otherwise, label the ends with adhesive labels.

The dots in the schematic, and on the pictorial, are provided to identify one end of the coil windings. Thus, the dot and no-dot ends are different from each other, and it usually makes a difference to circuit operation which way the ends are connected into the circuit (the issue is signal phasing).

Figure 24-14 shows two accepted methods for winding a multifilar coil on a toroidal core. Figure 24-14A is the same method as in Fig. 24-13B, but on an actual toroid instead of a pictorial representation. The wires are laid down parallel to each other as shown previously. The method in Fig. 24-14B uses twisted wires. The three wires are “chucked up” in a drill and twisted together before being wound on the core. With one end of the three wires secured in the drill chuck, anchor the other end of the three wires in something that will hold it taut. Some people use a bench vise for this purpose. Turn on the drill at slow speed and allow the wires to twist together until the desired pitch is achieved.

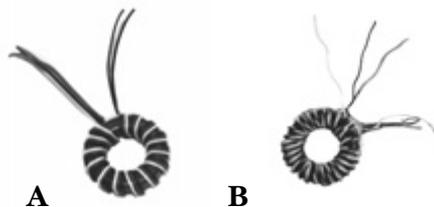
*Be very careful* when performing this operation. If you don't have a variable speed electric drill (so that it can be run at very low speed), then use an old-fashioned manual hand drill. If you use an electric drill, then wear eye protection. If the wire breaks, or gets loose from its mooring at the end opposite the drill, it will whip around wildly until the drill stops. That whipping wire will cause painful welts on the skin, and can easily damage eyes permanently.

Of the two methods for winding toroids, the method of Figs. 24-13B and 24-14A is preferred. When winding toroids, at least those of relatively few windings, pass the wire through the “doughnut hole” until the toroid is about in the middle of the length of wire. Then, loop the wire over the outside surface of the toroid, and pass it through the hole again. Repeat this process until the correct number of turns is wound onto the core. Be sure to press the wire against the toroid form, and keep it taut as you wind the coils.

Enameled wire is usually used for toroid transformers and inductors and that type of wire can lead to a problem. The enamel can chip and cause the copper conductor to contact the core. On larger cores, such as those for antenna-matching transformers (and baluns used at kilowatt power levels), the practical solution is to wrap the bare toroid core in a layer of fiberglass packing tape. Wrap the tape exactly as if it was wire, but overlap the turns slightly to ensure covering the entire circumference of the core.

On some projects, especially those in which the coils and transformers use very fine wire (e.g., no.30), you may experience a tendency for the wire windings to unravel after the winding is completed. This problem is also easily curable. At the ends of the windings, place a tiny dab of rubber cement or RTV silicone sealer.

**24-14** Winding a toroidal transformer  
A) parallel wound,  
B) twist wound.



**Mounting toroid cores** Now that you have a properly wound toroidal inductor or transformer, it is time to actually mount it in a circuit being constructed. There are three easy ways to do this job. First, you might be able to ignore it. If the wire is heavy enough, then just use the wire connections to the circuit board or terminal strip to support the component. But if this is not satisfactory, and in mobile equipment (or wherever else vibration is a factor) it won't be, then try laying the toroid flat on the board and cementing it in place with silicone seal or rubber cement. The third method is to drill a hole in the wiring board and use a screw and nut to secure the toroid. Do not use metallic hardware for mounting the toroid! Metallic fasteners will alter the inductance of the component and possibly render it unusable. Use nylon hardware for mounting the inductor, or transformer.

**How many turns?** Three factors must be taken into consideration when making toroid transformers or inductors: *toroid size*, *core material*, and *number of turns of wire*. The toroid size is selected as a function of power-handling capability, or for convenience of handling. The core material is selected according to the frequency range of the circuit. The only thing left to vary is the number of turns. The size and core material yields a figure called the  $A_L$  factor. The required value of inductance and the  $A_L$  factor are used in the following equation:

$$N = 100 \sqrt{\frac{L_{\mu\text{H}}}{A_L}} \quad \text{[24.38]}$$

where:

$N$  is the number of turns

$L_{\mu\text{H}}$  is the inductance in microhenrys

$A_L$  is the core factor, in microhenrys per 100 turns

**Example** Calculate the number of turns required to make a 5- $\mu\text{H}$  inductor on a T-50-6 core. The  $A_L$  factor is 40.

**Solution:**

$$\begin{aligned} N &= 100 \sqrt{\frac{L_{\mu\text{H}}}{A_L}} \\ &= 100 \sqrt{\frac{5_{\mu\text{H}}}{40}} \\ &= 100 \sqrt{0.125} \\ &= (100) (0.35) = 35 \end{aligned}$$

Don't take the equation value too seriously, however, because it is my experience that a wide tolerance exists on amateur-grade ferrite cores. Although it isn't too much of a problem when building transformers, it can be critical when making inductors for a tuned circuit. When you find that the tuned circuit takes considerably more (or less) capacitance than called for in the standard equation, and all of the stray capacitance is properly taken into consideration, then it may be that the actual  $A_L$  value of your particular core is different from the table value.

## Ferrite rods

Another form of ferrite core available on the market is the rod, shown in Fig. 24-15A. This type of core is used to make RF chokes, such as the RFC used in the vacuum tube filament lines of a linear amplifier power tube (Fig. 24-15B). They are also used for baluns by some people. The two windings are wound in a bifilar manner over the ferrite rod. Of course, the wires used must be heavy enough to carry the filament current of the tube. As was true in the toroidal transformer, I use two different wire colors in order to make it easy to identify windings.

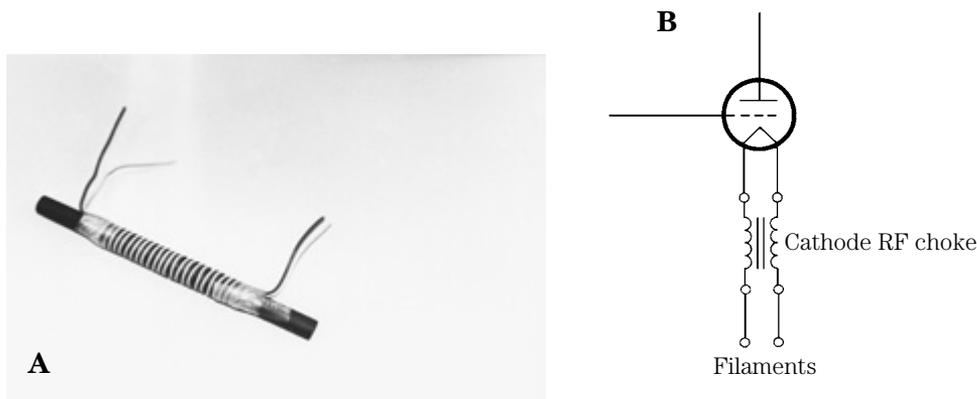
Ferrite rods are also used in receiving antennas. Although the amateur use is not extensive, there are places where a ferrite rod antenna (or “loopstick”) is used. For example, in radio direction finding antennas, it is common to see the ferrite loop. Also, some amateurs report that they use a loopstick receiving antenna when operating on crowded bands, such as 40 and 75 m. The small loopstick has an extremely directional characteristic, so it is capable of nulling out interfering signals. Of course, you would not want to use the loopstick for transmitting, so some means must be found for transferring the antennas over between the transmit and receive functions.

**Mounting ferrite rods** Ferrite rods can be mounted in several ways, two of which are analogous to the methods used on toroids. We can, for example, mount the rod using either its own wires for support or by using a dab of cement or silicone sealer to fasten it to the board. Although we cannot use simple nylon screws the way we can on toroids, we can use insulating cable clamps to secure the ends of the rod to the board.

## Toroid broadbanded impedance matching transformers

The toroidal transformer forms a broadbanded means for matching antenna impedance to the transmission line, or matching the transmission line to the transmitter. The other matching methods (shown thus far) are frequency-sensitive, and must be readjusted whenever the operating frequency is changed even a small amount. Although this problem is of no great concern to fixed-frequency radio stations, it is of critical importance to stations that operate on a variety of frequencies.

Figure 24-16A shows a trifilar transformer that provides a 1:1 impedance ratio, but it will transform an unbalanced transmission line (e.g., coaxial cable) to a

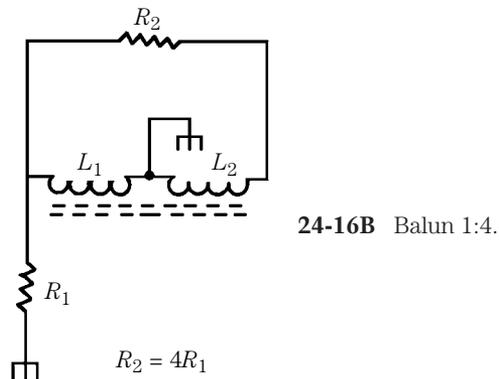
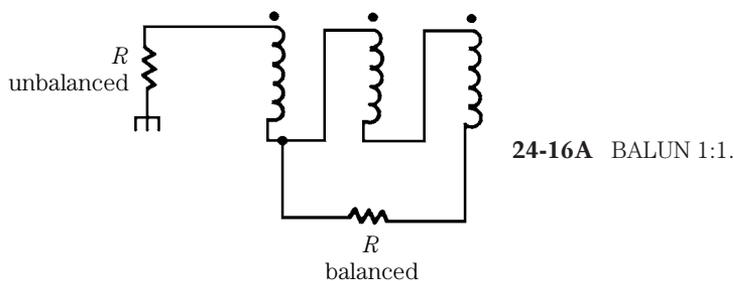


24-15 Ferrite rod inductor (A) construction, (B) use in amplifier.

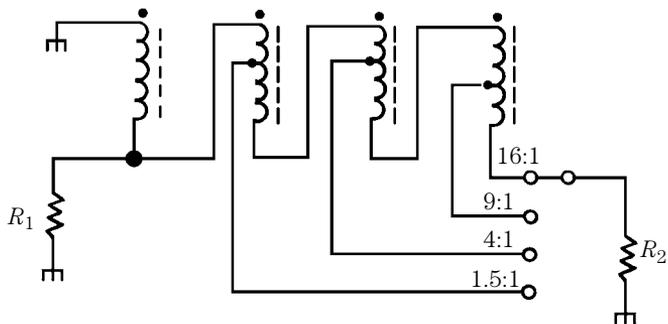
balanced signal required to feed a dipole antenna. Although it *provides no impedance transformation*, it does tend to balance the feed currents in the two halves of the antenna. This fact makes it possible to obtain a more accurate figure-8 dipole radiation pattern in the horizontal plane. Many station owners make it standard practice to use a balun at the antenna feedpoint.

The balun shown in Fig. 24-16B is designed to provide the unbalanced to balanced transformation, while also providing a 4:1 impedance ratio. Thus, a 300  $\Omega$  folded dipole feedpoint impedance will be transformed to 75  $\Omega$  unbalanced. This type of balun is often included inside antenna tuners, including all three models shown in Figs. 24-8A and 24-8B. A variable (or at least settable) broadbanded transformer is shown in Fig. 24-16C. In this case the output winding is tapped, and the operator selects the correct tap needed to provide the desired impedance ratio. The usual turns ratio criterion applies.

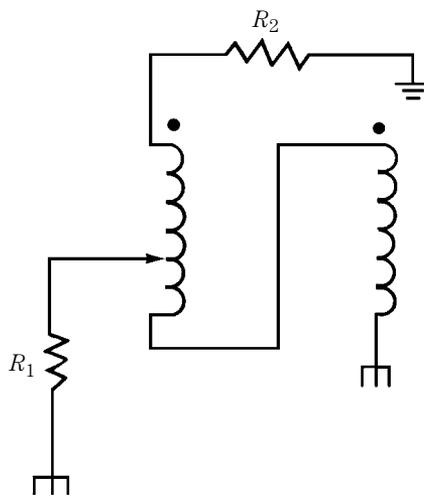
Another multiple-impedance transformer is shown in Fig. 24-16D. In this case, the operator can select impedance transformation ratios of 1.5:1, 4:1, 9:1 or 16:1. A commercial version of this type of transformer is shown in Fig. 24-16E. It is manufactured by Palomar Engineers, and is intended for vertical antenna feeding in the HF bands. It will, however, work well on antennas other than simple verticals. A top view of the vertical antenna feeding transformer is in Fig. 24-16F.



24-16C Tapped BALUN.



24-16D Multi-impedance BALUN.



24-16E Vertical antenna feeding transformer.



24-16F Top view of Fig. 24-16E.

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## 25 CHAPTER

# Mobile, emergency, portable, and marine antennas

MOBILE OPERATION OF RADIO COMMUNICATIONS EQUIPMENT DATES BACK TO ONLY A little later than “base station” operation. From its earliest times, radio buffs have attempted to place radio communications equipment in vehicles. Unfortunately, two-way radio was not terribly practical until the 1930s, when the earliest applications were amateur radio and police radio (which used frequencies in the 1.7- to 2.0-MHz region). Over the years, the land mobile and amateur radio mobile operation has moved progressively higher in frequency because of certain practical considerations. The higher the frequency, for example, the shorter the wavelength; and, therefore, the shorter a full-size antenna. On the 11-m Citizens Band, for example, a quarter-wavelength whip antenna is 102 in long, and for the 10-m amateur band only 96 in long. At VHF frequencies, antennas become even shorter. As a result, much mobile activity takes place in the VHF and UHF region.

The amateur 144-, 220-, and 440-MHz bands are popular because of several factors, not the least of which is the ease of making  $\frac{1}{4}$ - and  $\frac{5}{8}$ -wavelength antennas. Because low-cost commercial antennas are available for these frequencies, however, we will not examine such antennas in this chapter. Rather, we will concentrate on high-frequency antennas.

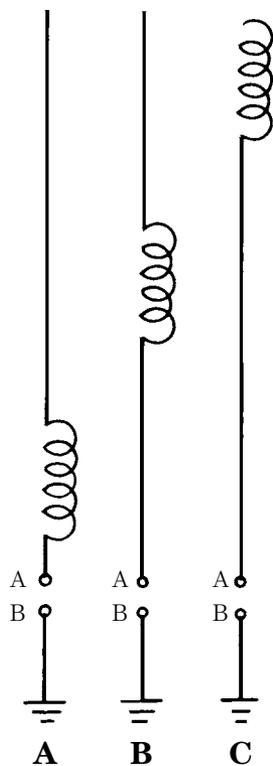
### Mobile HF antennas

High-frequency (HF) mobile operation requires substantially different antennas than VHF or UHF. Quarter-wavelength antennas are feasible only on the 11- and 10-m bands, with some argument in favor of 13 m as well. But by the time the frequency drops to the 21-MHz (15-m) band, the antenna size must be approximately 11 ft long, and that is too long for practical mobile operation. Because of practical considerations,

you must limit antenna size to 8 or 9 ft; in the amateur radio bands the 8-ft antenna is most popular because it is resonant on 10 m. At all lower frequencies, the 8-ft whip becomes capacitive, and therefore requires an equal inductive reactance to cancel the capacitive reactance of the antenna.

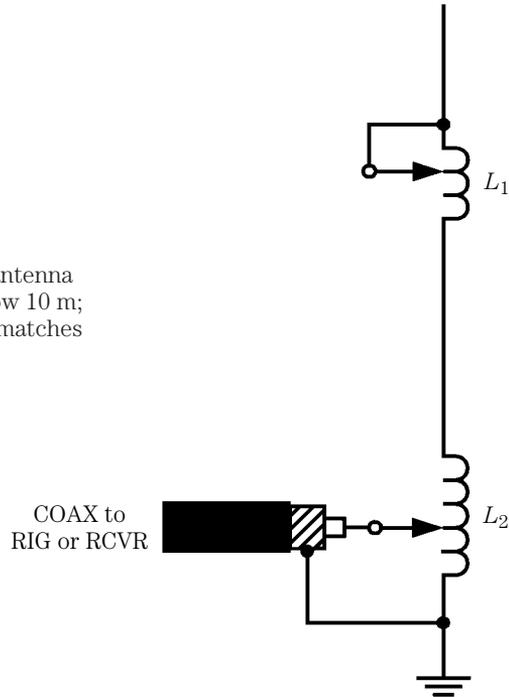
Figure 25-1 shows three basic configurations of coil-loaded HF antennas for frequencies lower than the natural resonant frequency of the antenna. In each case, the antennas are series fed with coaxial cable from the base; point *A* is connected to the coaxial cable center conductor, and point *B* is connected to the shield, and the car body (which serves as ground). The system shown in Fig. 25-1A is base-loaded. Although convenient, there is some evidence that the current distribution is less than optimum. The version shown in Fig. 25-1B is center-loaded, and results in an improved current distribution. This configuration is probably the most common among commercially available HF mobile antennas, although the coil is often located slightly above the center point. Finally, we see the top-loaded coil system in Fig. 25-1C. In all three cases, the point is to use the inductance of the coil to cancel the capacitive reactance of the antenna.

A modified version of the coil-loaded HF mobile antenna is shown in Fig. 25-2A. In this configuration the loading inductance is divided between two individual coils,  $L_1$  and  $L_2$ . Coil  $L_1$  is adjustable with respect to the antenna, while  $L_2$  is fixed; coil  $L_2$  is tapped, however, in order to match the impedance of the antenna to the charac-



25-1 Loaded mobile antennas.

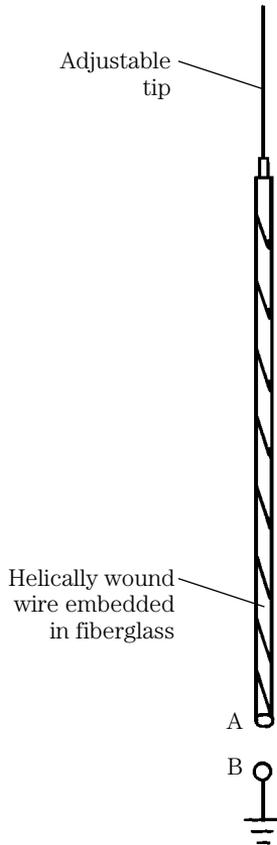
**25-2A** Practical mobile antenna for HF bands below 10 m;  $L_1$  resonates,  $L_2$  matches impedance.



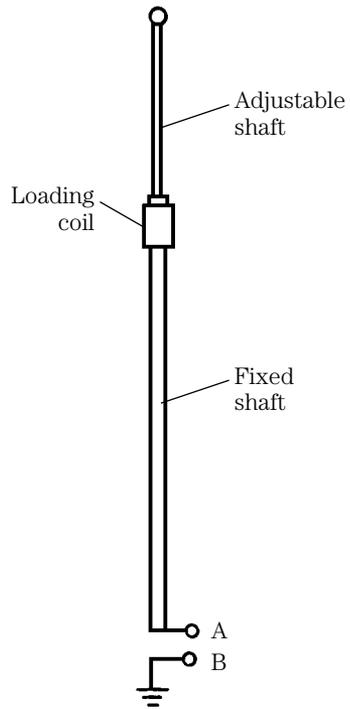
teristic impedance of the coaxial-cable transmission line. When tuning this antenna, two instruments are needed: a field strength meter and a VSWR meter. The field strength meter gives a relative indication of the amount of power radiated from the antenna, while the VSWR meter helps determine the state of the impedance match. A pair of relative field strength meter projects are discussed later in this chapter.

Variations on the theme are shown in Figs. 25-2B and 25-2C. A helical whip is shown in Fig. 25-2B. In this configuration, the inductor is distributed along the length of a fiberglass antenna rod. The conductor is a wire that is helically wound on the fiberglass shaft. An adjustable tip sets the antenna to resonance. The length of this tip can be set by a local field strength meter. Fixed versions of this type of antenna are very popular on the Citizens Band where they may be as short as 30 in, or as long as 48 in. There are also some amateur radio commercial antennas based on this concept.

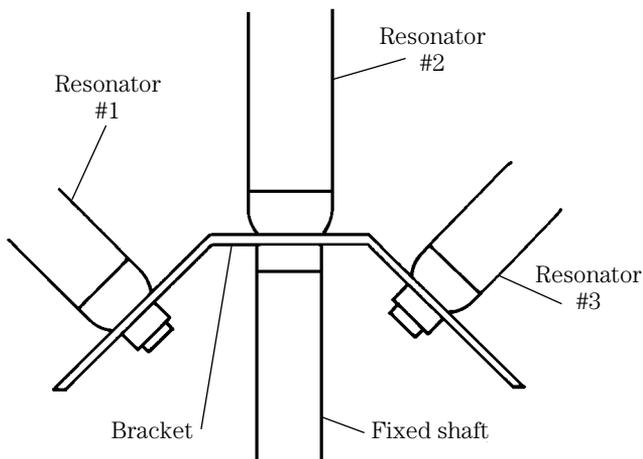
The other variant is shown in Fig. 25-2C. In this case, the lower end of the radiator consists of a metal tube topped with a loading coil. An adjustable shaft at the top end is used to tune the antenna to resonance. This form of antenna is popular among commercial makers of HF mobile antennas, such as the Hustler. The fixed shaft tends to be universal for all bands, while the coil and adjustable shaft form a separate “resonator” for each band. Multiband operation of this form of antenna can be accomplished by using a bracket such as shown in Fig. 25-2D. Although early versions of this scheme were homebrewed, several manufacturers currently make factory-built versions. The idea in Fig. 25-2D is to mount two or more resonators to a common fixed shaft.



**25-2B** Helically wound mobile antenna.



**25-2C** Top-loaded mobile antenna common in amateur radio.



**25-2D** Multiband top-loaded antenna.

A common problem with all coil-loaded mobile antennas is that they tend to be very high  $Q$  antennas. In other words, they are very sharply tuned. The VSWR tends to rise rapidly as the operating frequency departs from the frequency to which the antenna is tuned; as little as 25-kHz change of operating frequency will detune the antenna significantly. Although an antenna tuner at the output of the transmitter will reduce the VSWR to a point that allows the transmitter to operate, that type of tuner is merely a line flattener that does not fully address the problem. The actual problem is that the antenna is not resonant. The efficiency of the antenna drops off rapidly as the frequency changes. The only cure for this problem is to readjust the resonator's adjustable shaft as the band segment is changed (not merely the band, but the *band segment*). Unfortunately, this solution requires tools, and the operator must get out of the vehicle in order to do a good job of retuning. Another solution is to use a motor-driven variable inductor for the loading coil. Several manufacturers offer both base-loaded and center-loaded coils that are either motor-driven or relay-selectable to permit frequency changing from the operator's seat.

### Tuning HF mobile antennas

Although the procedures for tuning certain specific antennas might be different from those given below, the basic principles are the same, and can easily be adapted to any given situation. There are two situations to consider. First is the antenna in which there is either a single adjustable (or tap-selectable) loading coil, or a fixed loading coil and an adjustable shaft resonator. Second is the case ( Fig. 25-2A) where there is an adjustable or tapped coil in series with a tapped impedance-matching coil at the base.

*Case no. 1.* In this case, we assume that the antenna has a single adjustment, either a shaft resonator or adjustable coil. In this case we need to use a field strength meter (FSM) to measure the relative field strength of the radiated signal. The antenna resonator is adjusted until the radiated field strength is maximum. Of course, this procedure must be done in steps, keying the transmitter after each adjustment to see what happened. Alternatively, a VSWR meter can be used to set the resonator to a minimum VSWR.

*Case no. 2.* Here, we have a coil at the base that is fixed with respect to the antenna radiator element, but is tapped with respect to the coaxial cable from the transmitter. A second coil is also used in the antenna (see Fig. 25-2A), which can be either in the center or at the top. This coil is adjustable for setting resonance. Alternatively, this coil may be fixed, and an adjustable resonator shaft may be used to set operating frequency. In this type of antenna the upper coil, or resonator shaft, is adjusted to resonance by using a field strength meter. The lower coil is adjusted for minimum VSWR. Both adjustments are needed to complete the job.

### Field strength meters

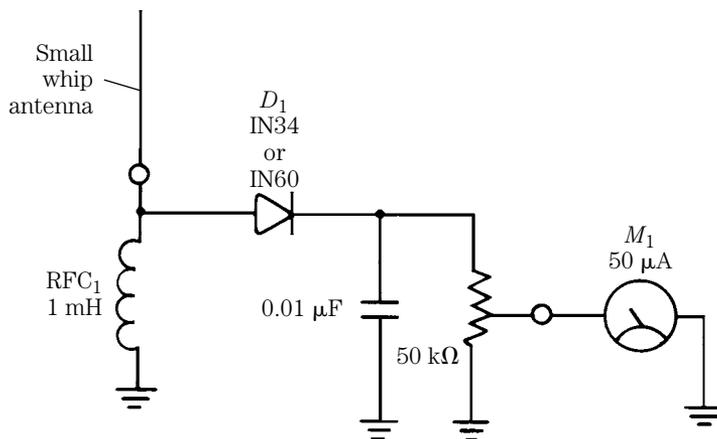
A *field strength meter* (FSM) is an instrument that measures the radiated field from an antenna. Commercial engineering-grade instruments are calibrated in terms of either watts/square centimeter or volts/meter, and are used for jobs such as broadcast station "proof-of-performance" tests and other professional jobs. For adjusting

antennas, however, a considerably different instrument is sufficient. This section describes two simple, passive (which means no dc power is required) field strength meters usable for adjusting HF radio antennas, including both CB and amateur radio antennas.

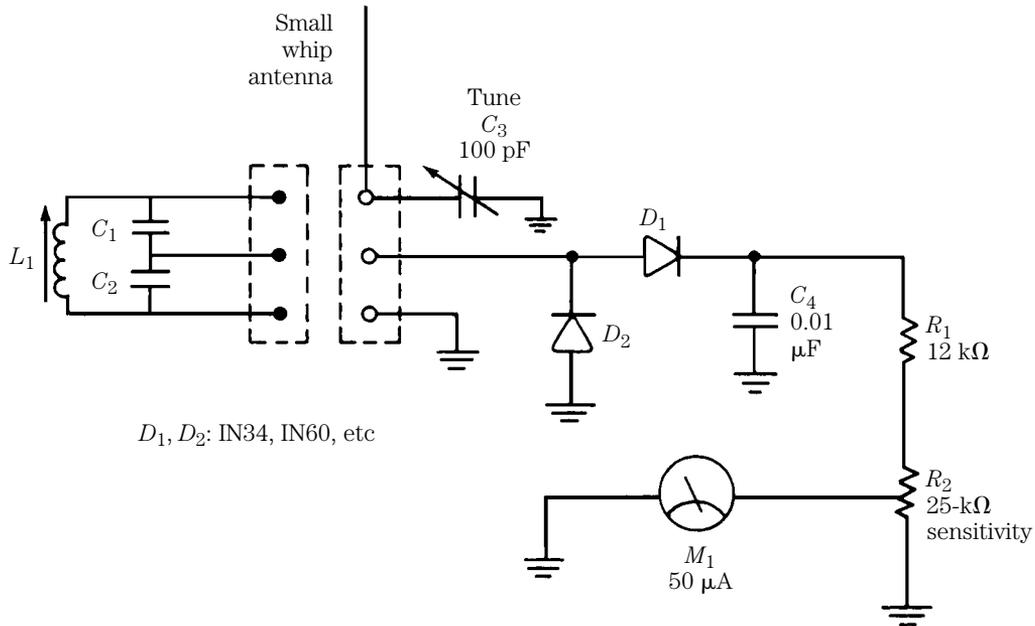
Two forms are shown in Fig. 25-3; both are basically variations on the old-fashioned “crystal set” theme. Figure 25-3A shows the simplest form of untuned FSM. In this circuit a small whip antenna (used for signal pickup) is connected to one end of a grounded RF choke (RFC). The RF voltage developed across the RFC is applied to a germanium diode detector; either IN34 or IN60 can be used. Silicon diodes are normally preferred in signal applications, but in this case we need the lower contact potential of germanium diodes in order to improve sensitivity ( $V_g$  is 0.2 to 0.3 V for Ge and 0.6 to 0.7 V for Si). A potentiometer is used both as the load for the diode and as a sensitivity control to set the meter reading to a convenient level.

The untuned version of the FSM is usable even at low powers, but suffers from a lack of sensitivity. Only a certain amount of signal can be developed across the RFC, so this limits the sensitivity. Also, the RFC does not make a good impedance match to the detector diode  $D_1$ . An improvement is possible by adding a tuned circuit and an impedance-matching scheme, as shown in Fig. 25-3B. In this case a variable capacitor  $C_3$  is used as a tune control. The tuning capacitor is parallel resonant with inductor  $L_1$ . A tapped capacitive voltage divider ( $C_1/C_2$ ) is used to provide impedance matching to the diode. Figure 25-3C gives values of capacitance and inductance for various bands.

Operation of the tunable FSM is simple and straightforward. Set the sensitivity control to approximately half-scale, and then key the transmitter; adjust the tuning control ( $C_3$ ) for maximum deflection of the meter pointer (readjustment of the sensitivity control may be needed). After these adjustments are made, the tunable FSM works just like any other FSM.



25-3A Simple field strength meter.



25-3B Tuned field strength meters.

Band (meters)	$L_1$ ( $\mu\text{H}$ )	$C_1$ (pF)	$C_2$ (pF)
75/80	27	25	100
40	10	15	68
30	4.4	15	68
20	2.25	10	68
17	1.7	10	56
15	1.3	10	47
13	0.8	10	39
10	0.5	10	39

25-3C Values chart for field strength meters.

## Antennas for emergency operations

Some time ago, I met an interesting character at a convention. As a medical doctor, he was a medical missionary working at a relief station in Sudan. Because of his unique business address, he was able to discuss mobile and portable antennas for communications from the boondocks. His *bona fides* for this knowledge includes the fact that he is licensed to operate on both the amateur radio bands, and as a landmobile (or point-to-point) station in the 6.2-MHz band. The desert where he travels is among the worst in the world. The path they euphemistically call a “road” is occasionally littered with camel corpses because of the harsh conditions. The doctor’s organization requires him to check in twice daily on either 6.2 MHz or

3.885 MHz (which some missionary hams in Africa use as an unofficial calling frequency). If he misses two check-ins in a row, then the search and rescue planes are sent up. As a result of his unique “house calls,” he does a lot of mobile and portable operating in the lower-HF region of the spectrum. His problem is this: How do you reliably get through the QRM and tropical QRN with only 200 W PEP and a standard loaded mobile antenna?

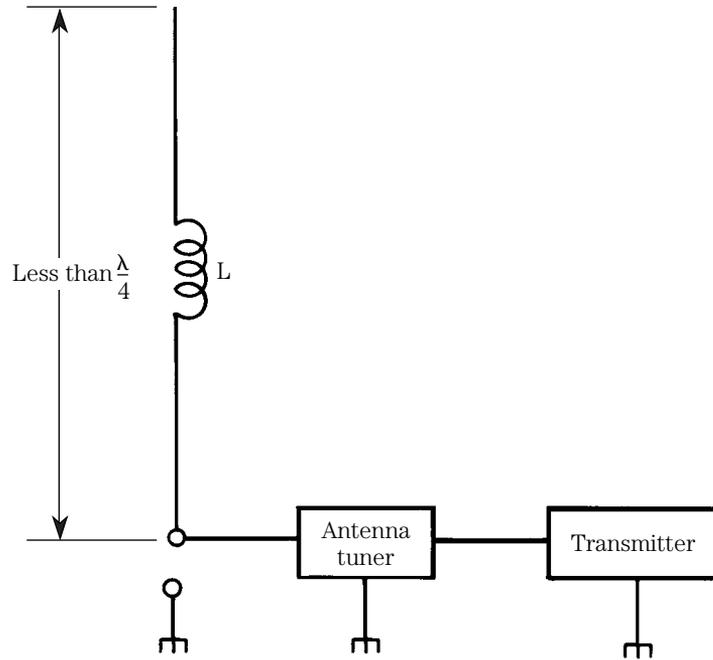
Another fellow I once met works in Alaska for a government agency. He faces many of the same problems as the doctor in Sudan, but at close to 100° colder. He frequently takes his 100-W mobile rig into the boondocks with him in a four-wheel drive vehicle. Again, with only 100 W into a poor-efficiency loaded mobile antenna, how does one reliably cut through the interference to be heard back at the homestead?

An earthquake, or hurricane, strikes your community. Antenna towers collapse, tribanders become tangled masses of aluminum tubing, dipoles are snarled globs of no.14 Copperweld, and the rig and linear amplifier are smashed under the rubble of one corner of your house. All that remains is the 100-W HF rig in your car. How do you reliably establish communications in “kilowatt alley” with a 100-W mobile driving a 75-W loaded whip? Of course, you always got through one way or another before, but now communications are not for fun—they are deadly serious. Somehow, the distant problems of a Sudanese missionary doctor and a KL7 government forester don’t seem too very far away.

For these operators, communications often means life or death for someone, perhaps themselves. Given the inefficiency of the loaded whips typically used as mobile antennas in the low-HF region, the generally low power levels used in available mobile rigs, and the crowded band conditions on the 80-, 75-, and 40-m bands, it becomes a matter of more than academic interest how you might increase the signal strength from your portable (or mobile) emergency station. Anything we can do, easily and cheaply, to improve the signal is like having money in the bank. Fortunately, there are several tricks of the trade that will help us out in a pinch.

Figure 25-4 shows a typical mobile antenna for a low-HF band. Because quarter-wavelength antennas on these frequencies are 30 to 70 ft high, full-size vertical whip antennas are not practical. In fact, at frequencies below 10 m, full-size whips are not generally used. Short antennas exhibit capacitive reactance, so we add a loading coil (inductor  $L$  in Fig. 25-4) to the radiator to make up the difference (its inductive reactance cancels the capacitive reactance of the antenna). The inductor in such a loaded antenna can be placed almost anywhere along the radiator, although base-, center-, and top-loaded designs predominate. The actual inductance needed varies somewhat with coil placement, as does antenna performance. The “resonators” used on commercial low-HF mobile antennas are loading coils encapsulated in a weathertight housing.

The mobile configuration is inefficient by its nature, and little can be done to improve matters. Of course, an antenna-matching device or tuner will help in optimizing the power transfer to the antenna, and should always be used in any event (especially with solid-state final amplifiers, which don’t tolerate VSWR as easily as do tube finals). In the portable configuration, however, we can both improve the performance of “mobile” antennas and look into antenna options not open to the mobile



25-4 Basic mobile antenna system often proves unsatisfactory in marginal cases.

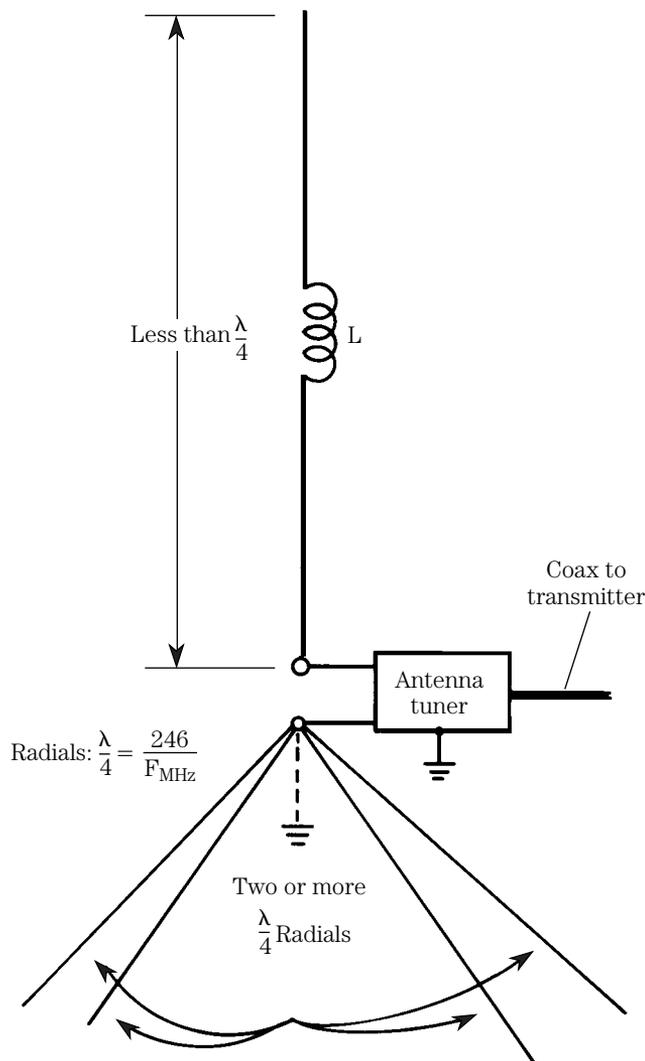
operator. We can make a basic assumption for these improvements: The operator needing emergency or urgent communications is not driving down the interstate somewhere, but is stopped in the back country and needs to call out.

In cases of emergencies on most highways, we are in range of some repeater, so we would use a VHF band (probably 2 m) to contact police or other emergency services through a repeater autopatch. In fact, with the wide availability of repeaters around the country it behooves any amateur, backpacking or four-wheeling into remote areas, to be familiar with repeater locations and frequencies. From hilltops, especially, it is often possible to hit repeaters from a surprisingly long distance. I have seen hand-held 2-m rigs trigger mountain top repeaters from the mountains of southwestern Virginia, and presume that other areas of the country are as well off repeaterwise. This chapter deals with HF rigs, especially those operating in the lower end of the HF spectrum, in situations where a temporary antenna must be erected.

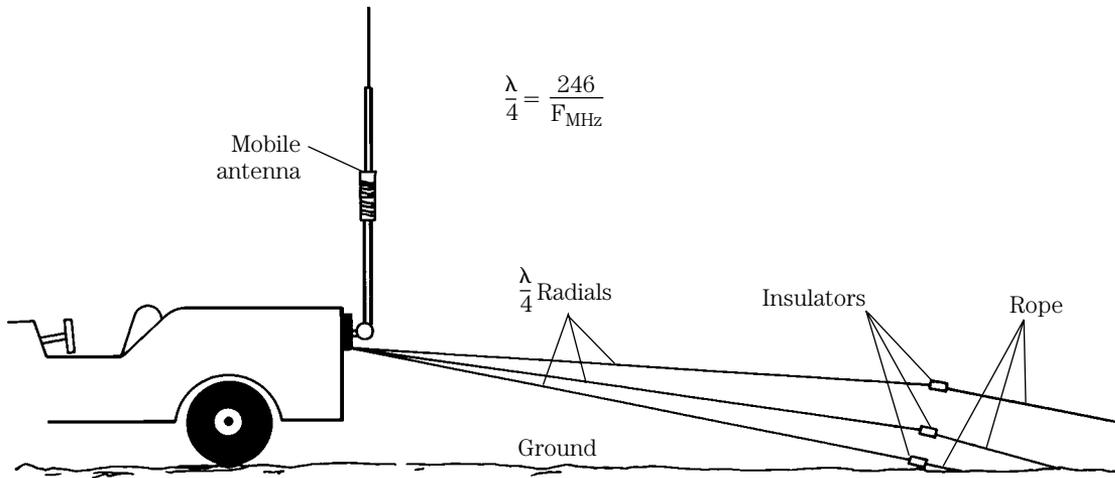
Most amateurs, who have low-HF mobile rigs, will testify that getting out is a pain in the mike button. Antenna efficiencies are simply too poor. One factor in this dismal equation is that the vehicle body makes a very poor antenna ground system. One could, I suppose, carry a supply of 6- or 8-ft copper-clad steel ground rods, and a 9-lb sledge hammer to drive them into the earth wherever a portable station is needed. Somehow, that solution doesn't seem too appealing at this point! Besides, have you ever tried to remove an effective 8-ft ground rod from the ground? Forget it, it's too much pain for too little gain. A better solution is to provide a counterpoise

groundplane, as shown in Fig. 25-5. In Fig. 25-5A we see the electrical situation and connection, while in Fig. 25-5B we see the mechanical scheme for a specific situation: A groundplane consists of two or more (even *one* helps) quarter-wavelength radials connected to the antenna ground point (i.e., where the coaxial cable shield connects to the vehicle body). The radials are made of no. 14 wire, so are relatively easy to stow.

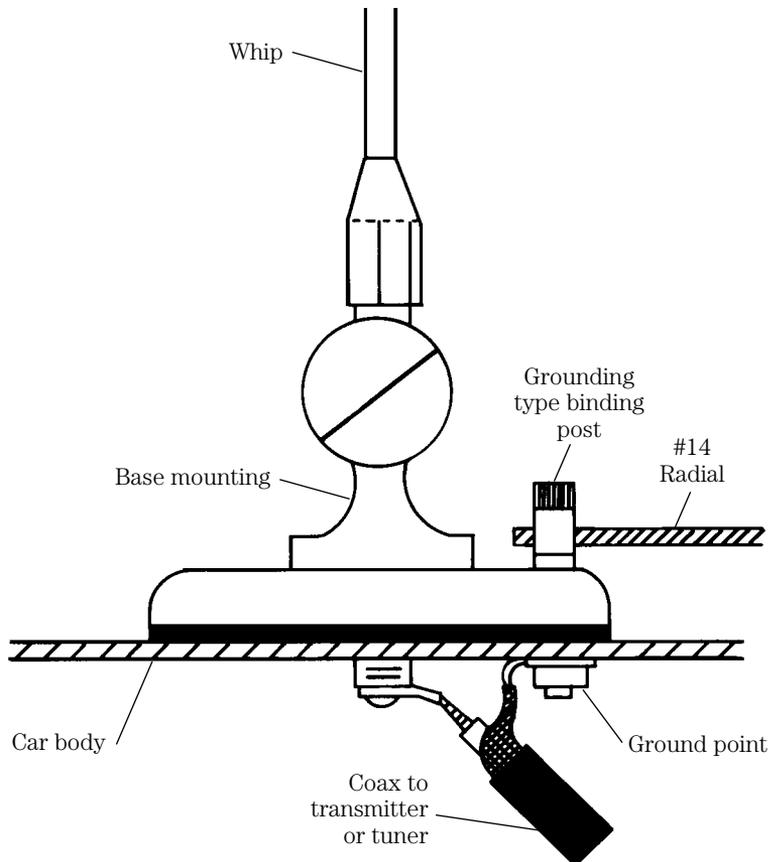
Figure 25-5B shows a workable system that will improve the performance of a mobile rig in stationary situations. The mobile antenna uses the normal base mount attached to the rear quarter panel of the car adjacent to the trunk lid. An all-metal grounding-type binding post is installed through an extra hole drilled in the base



**25-5A** Two or more radials improve efficiency of mobile antenna used in a portable communications setup.



25-5B Multiple radials are even better.



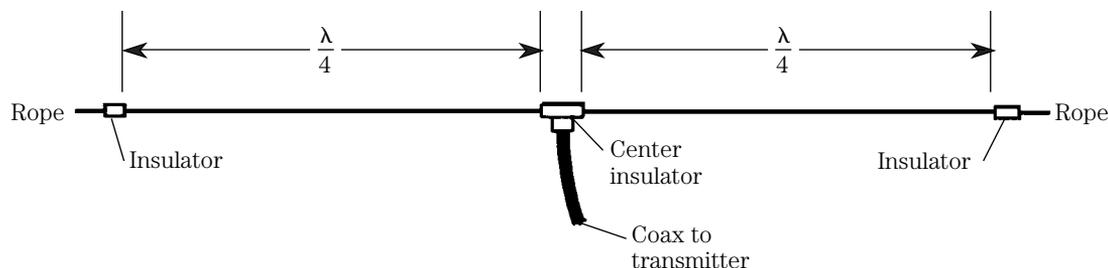
25-5C Connection of radials to mobile antenna.

insulator (see Fig. 25-5C). Radials for portable operation are attached at this point. Although the binding post is small, it easily accommodates two no. 14 radials. The owner used the vehicle for camping, and the radials were used to improve antenna efficiency. I don't believe the dramatic accounts enthusiastically reported by the owner of the car, but my own Field Day experience (and the testimony of the Sudanese missionary doctor) lead me to expect considerable improvement over the unaided loaded whip.

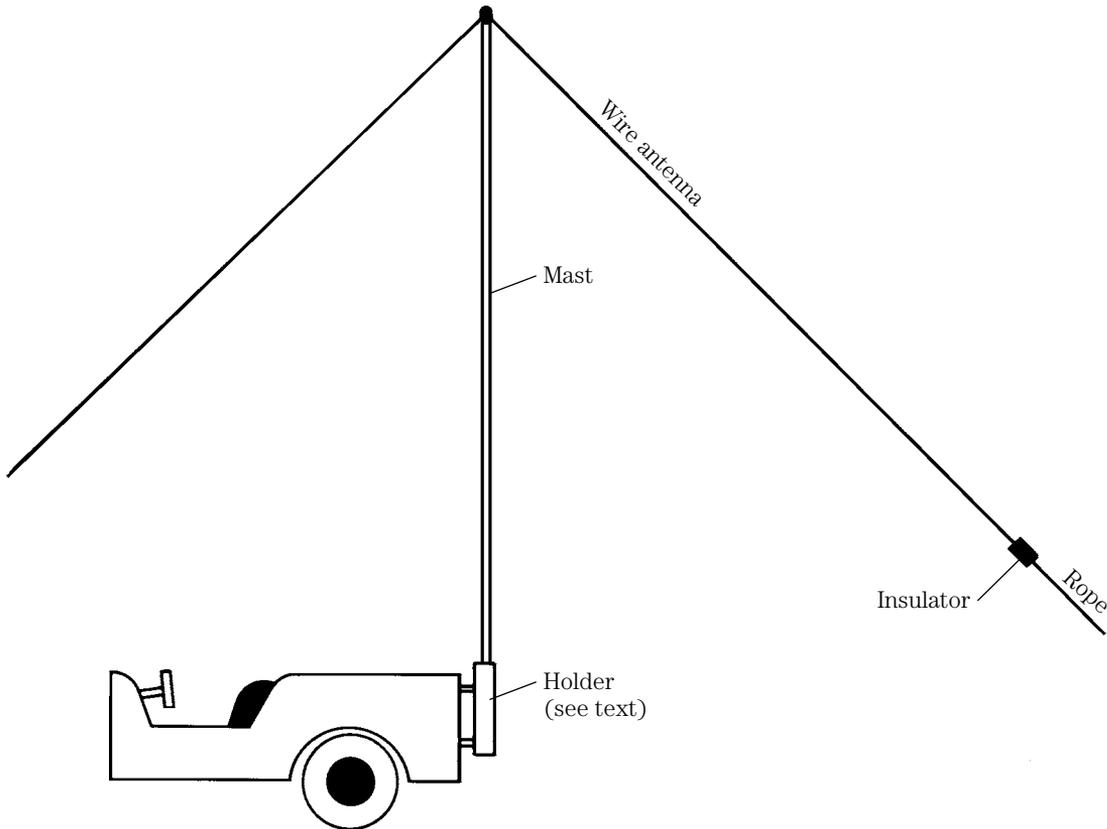
A potential solution to the inefficiency problem is to replace the mobile antenna with a more efficient, but stowable, antenna that can be brought out and erected when needed. One candidate is a surplus military HF whip antenna. Intended for jeeps and communications trucks, these antenna/tuner combinations are collapsible and are as efficient as any on the market. Although my Alaskan friend could use such a surplus find, my Sudanese friend could not. He told me that Americans in the boonies, even (perhaps especially) missionaries, are always suspected by pin-headed local authorities of being agents of the dreaded "Cuban Invasion Authority." Showing up "out there" with obviously military radio equipment serves only to seal that belief!

The common dipole is often looked down upon by the owners of massive array antennas, but those old-fashioned antennas are capable of turning in some impressive results. The dipole is made by connecting two quarter-wavelength pieces of wire to a coaxial-cable transmission line. One length is connected to the center conductor, and the other is connected to the shield of the coax. In a pinch, zip (i.e., lamp) cord and twisted pairs of hookup wire will do for a transmission line. Figure 25-6 shows the common dipole and the normal equation for determining *approximate* length. The actual length is found by trimming length until the VSWR drops to its lowest point, a nicety that might not seem altogether important in an emergency. The ends of the dipole must be supported on trees, masts, or some other elevated structure. Unfortunately, mounting points aren't easily found. My Sudanese friend doesn't see many trees!

Figure 25-7 shows another alternative antenna that works well for portable operation. Based on the dipole, this inverted-vee dipole does not require two end supports but rather a single center support. The length of each leg is 6 percent longer than for a nominal dipole. Because the application is both emergency-related and temporary in nature, we can get away with construction methods that would be unthinkable in more permanent installations.



25-6 Basic dipole.



25-7 Portable inverted-vee antenna.

Three problems present themselves. First, the antenna must be portable for backpackers, or stowable in the case of those people who motor into the back country. Size and weight are major constraints in both cases, even though the vehicular case is a lot easier to work with than the backpacker. Second, what materials and means of construction are needed for the mast? Third, how is the mast supported? Because the answer to the third question depends in part on the answer selected for the second, we will deal with it first. The first question is avoided at this point because I have selected the vehicular case.

One alternative is to use a telescoping TV antenna mast to support your antenna. We can either mount a mobile whip and its associated radials at the top, or install an inverted-vee per Fig. 25-7. These masts collapse to 6 or 8 ft in length, but can be slipped up to heights of 18, 25, 30, 40, or 50 ft, depending upon the type selected. Keep in mind, when shopping for these masts, that the larger models are considerably heavier than shorter models, and they require two or more people to install them. Erecting a 40- or 50-ft telescoping mast is not a single-person job, even though you might know someone who has done it. I've done it, and won't do it again! Even the 30-ft models are a bit hairy to install alone.

Another mast that can be used for the inverted vee is made from PVC plastic plumbing pipe. If it can be carried on or in the vehicle, then lengths up to 10 ft are available. Longer lengths are made at the site of erection by joining together 10-ft or shorter sections with couplings (also available at plumbing supply outlets). Be careful of using PVC pipe that is too small, however. PVC pipe is relatively thin walled, and it is therefore flexible. Sizes below 1.5-in diameter will not easily stand alone without guying. While a single 10-ft section might be self-supporting, two or more sections together will not support themselves and the weight of the antenna. Guying can be accomplished with ropes, or on a temporary basis, heavy twine.

Still another alternative is to carry steel TV antenna masts. Available in 5- and 10-ft lengths, these masts are flared on one end and crimped on the other, so that they can be joined together to form longer lengths. In addition, the same sources that sell these masts also sell the guy wire rings that help support the mast when installed. There is also a variety of rooftop mounting devices that aid also in ground mounting the antenna.

I used the TV mast solution on Field Day expeditions for many years. The inverted-vee antenna is, of course, a natural for this solution. Once, however, we used a Hustler mobile antenna mounted at the top of a 20-ft mast, with four radials, and it worked surprisingly well with the 30-W limit for extra points credit then available for Field Day contestants.

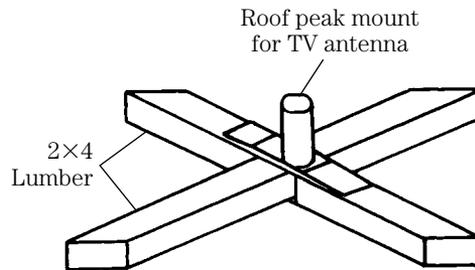
There are, then, several alternatives for masts that slip up, snap together, or are otherwise easily unstowed and installed. Let's now turn our attention to base mounting schemes that can be used to support the mast. Of course, in a life or death pinch, you could always just pile rocks around the base, or hold the darn thing up as you transmit. But a little forethought could eliminate those problems easily.

One of the first solutions I'll deal with was seen on the Outer Banks of North Carolina, and was provided to me by a CB operator. Surf fishers on the Outer Banks use four-wheel drive vehicles to get out on the beach (to the surf, where the big sea bass lurk). Welded to either the front or rear bumper attachments are steel tubes (see Fig. 25-7) used for mounting the very, very long surf casting rods they use to catch fish. The CB operator had a 20-ft mast consisting of two 10-ft TV mast sections mounted in one of the rod holders. At the upper end of the mast was his 11-m groundplane antenna. The same method of mounting would also support similar amateur antennas, inverted-vee dipoles, or VHF/UHF antennas.

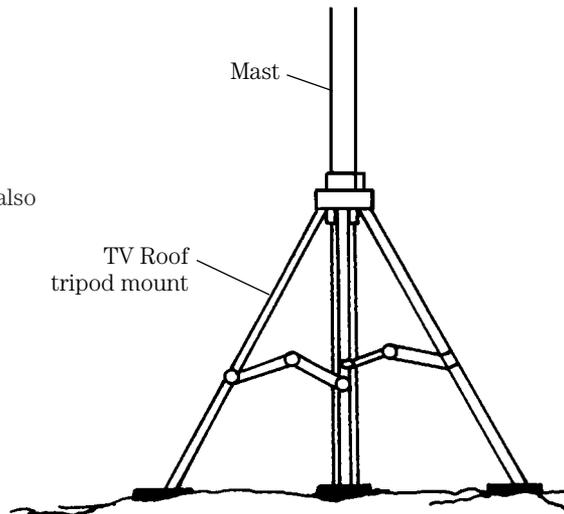
Given that the antenna installation will be temporary, lasting only a few hours or a couple of days in the worst case, we need not worry about long-term integrity, or the practicality of the installation. Mounting the mast to the back of a four-wheeler or pickup truck with a pair of U bolts is not terribly practical if you must move the vehicle, but works nicely if you plan to camp (or are stranded) for a few days.

For lightweight masts, up to about 25 ft, the base support could be an X-shaped base made of  $2 \times 4$ -in lumber (Fig. 25-8A), or even a Christmas tree holder. Alternatively, a TV antenna rooftop tripod mount (Fig. 25-8B) is easily adopted for use on the ground. None of these three alternatives can be depended upon for self-supporting installations, and must be guyed even if used for only a short period. Again, because of the temporary nature of the installation, aluminum or wooden tent pegs can be used to anchor the guy wires. Although they are insufficient for long-term installations, they work fine for the short run.

25-8A Wooden antenna stand.



25-8B TV-antenna tripod also works in portable situation for HF antenna masts.

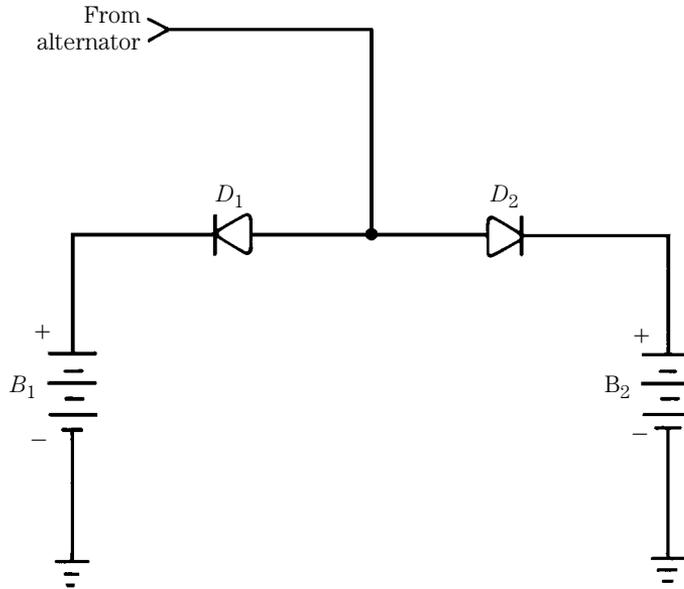


## Another problem

Field Day is a reasonably good training ground for people who anticipate operating radio equipment under primitive conditions. If you pay attention to the sometimes humorous foul-ups seen while getting on the air from atop High Peak Mountain in Goose Bump National Park, then you might avoid some of the same mistakes when foul-ups mean more than a quick drive to the nearest town. You will learn, for example, which simple tools are a must for constructing and repairing simple antennas (including the 12-Vdc soldering iron that runs off the vehicle battery). You will also learn a bit about electricity needed to run the rig.

Two alternatives present themselves. First, you could use 12 Vdc from the vehicle electrical system, which makes sense with the current crop of 12 Vdc HF rigs on the market. Or, you could operate from 120 Vac generated by a light plant generator.

When boondocking in a four-wheeler (or other vehicle), it is wise to use a dual battery system such as shown in Fig. 25-9. Two separate 12-Vdc auto batteries (preferably high ampere-hour capacity) are connected essentially in parallel with the alternative charging system. Diodes  $D_1$  and  $D_2$  are rated at 100 A, 50 V peak inverse voltage (PVI), and are used to isolate the two batteries from each other. Diode assemblies such as this can be built, or purchased from van conversion and recreational



25-9 Diode pack allows charging two batteries from system charger.

vehicle shops. Van and RV owners' use is the same as yours: It's a darn shame to run down the battery needed to start the vehicle just by operating your rig. You not only cannot start the vehicle in such a case, but cannot muster enough power to call for the wilderness equivalent of "Triple A."

One experienced boondocker I met in southern Arizona totes either a 450-W Kawasaki or 500-W Honda mini light plant generator in the back of his Bronco. These models are surprisingly lightweight and quiet operating. One of the generators has a 12-Vdc, 8-A outlet that can be used to charge a dead battery. The other has only the 110-Vac outlet, but can be used for battery charging if a small charger is provided. Simple 10-A battery chargers can usually be bought at auto parts or accessories stores, for about \$60.

Operating radio communications equipment successfully under primitive conditions depends upon two major factors: available electrical power and a proper, efficient, antenna system. Although dealing in detail with the means of obtaining power in remote locations is beyond the scope of this book, we have provided some ideas from which you can start planning your own "survival" radio system.

## Marine radio antennas

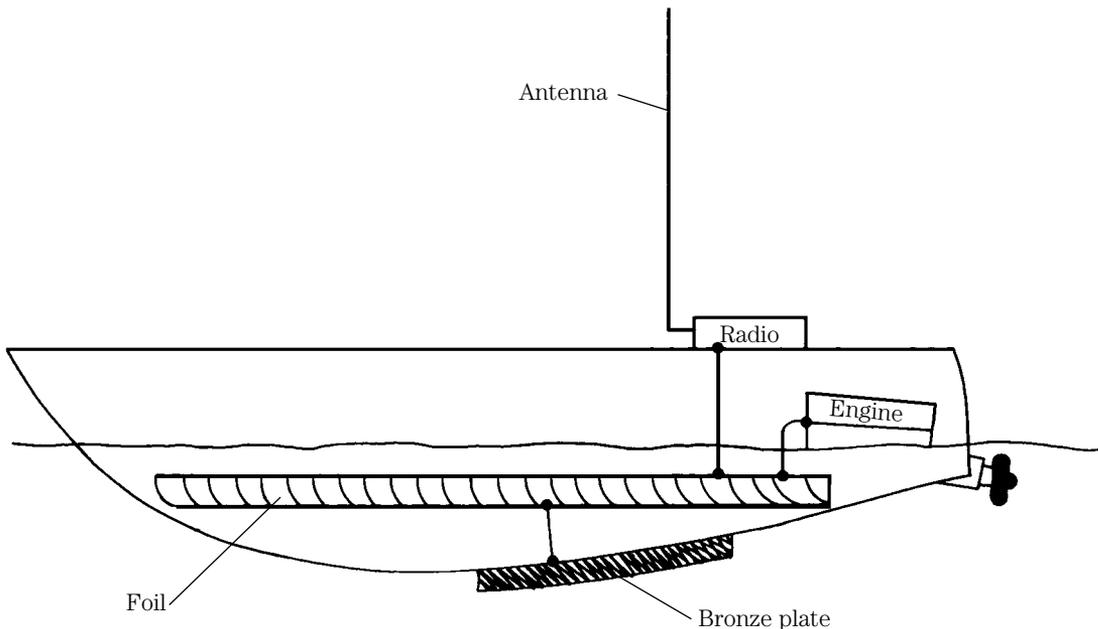
Radio communications considerably lessened the dangers inherent in sea travel. So much so, in fact, that the maritime industry took to the new "wireless telegraphy" earlier than any other segment of society. In the early days of radio, a number of exciting rescues occurred because of wireless. Even the infamous *S.S. Titanic* sinking might have cost less human life if the wireless operator aboard a nearby ship had

been on duty. Today, shore stations and ships maintain 24-hour surveillance, and vessels can be equipped with autoalarm devices to wake up those who are not on duty. Even small pleasure boats are now equipped with radio communications, and these are monitored around the clock by the U.S. Coast Guard.

The small shipboard operator can have a selection of either HF single-sideband communications or VHF-FM communications. The general rule is that a station must have the VHF-FM, and can be licensed for the HF SSB mode only if VHF-FM is also aboard. The VHF-FM radio is used in coastal waters, inland waters, and harbors. These radios are equipped with a high/low power switch that permits the use of low power (1 W typically) in the harbor, but higher power when under way. The HF SSB radios are more powerful (100 W typically), and are used for offshore long-distance communications beyond the line-of-radio-sight capability of the VHF-FM band.

The problems of antennas on boats are the same as for shore installations, but are aggravated by certain factors. Space, for one thing, is less on a boat so most HF antennas must be compensation types. Also, grounds are harder to come by on a fiberglass or wooden boat, so external grounds must be provided.

A typical powerboat example is shown in Fig. 25-10. The radio is connected to a whip antenna through a transmission line (and a tuner on HF), while also being grounded to an externally provided groundplate. Over the years, radio grounds have taken a lot of different forms. For example, the ground might be copper or aluminum foil cemented to the boat hull. Alternatively, it might be a bronze plate or hollow bronze tube along the center line of the boat, or along the sides just below the water line. The ground will also be connected to the engine. Careful attention must be paid

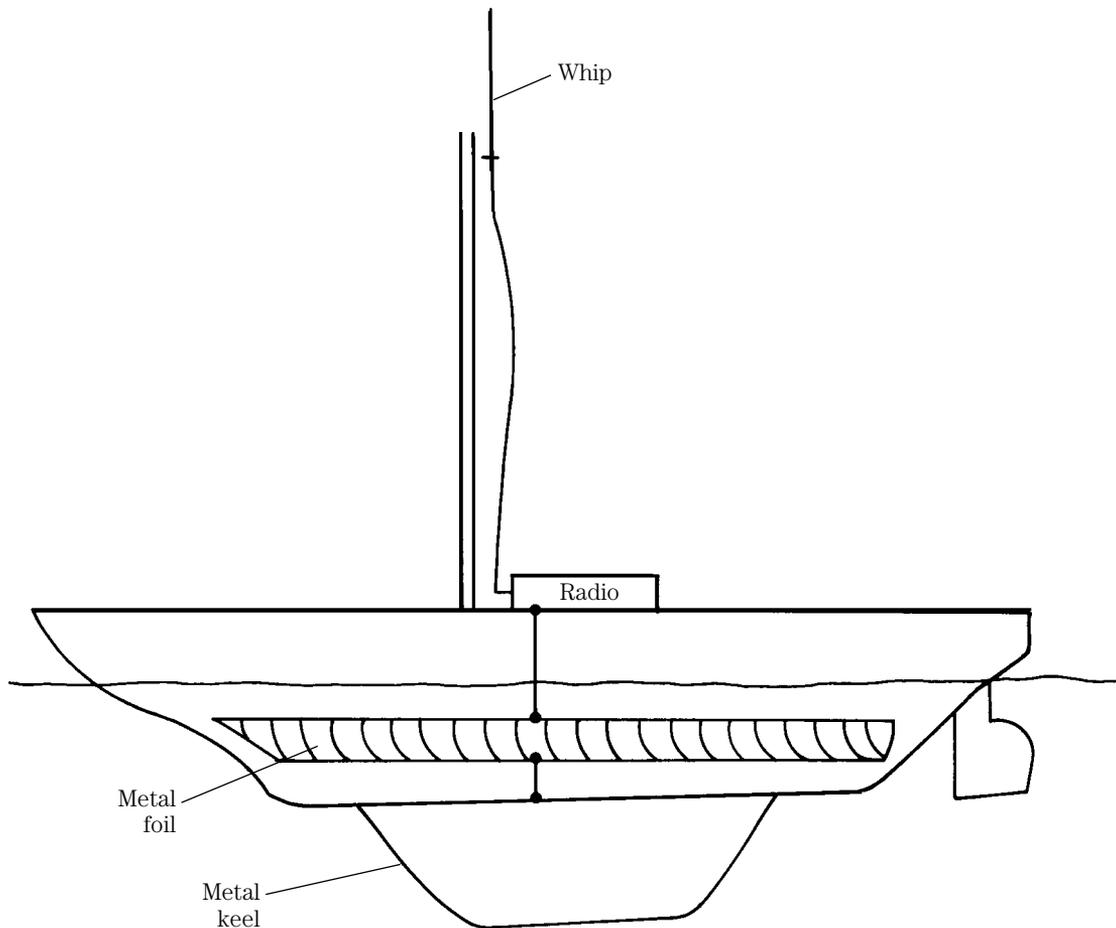


25-10 Motorboat grounding system.

to the electrical system of the boat when creating external grounds to prevent electrolytic corrosion from inadvertent current flows.

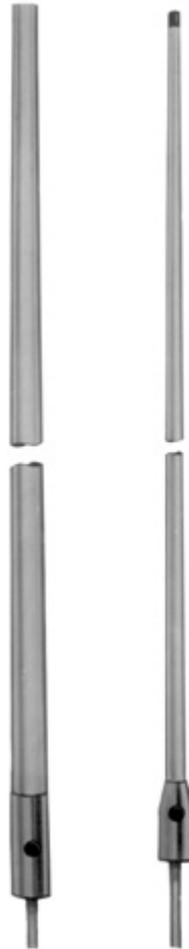
On sailboats the whip antenna (especially VHF-FM) might be mast-mounted as shown in Fig. 25-11. The same grounding scheme applies with the addition of a metal keel or metal foil over a nonmetallic keel.

The whips used for boat radios tend to be longer than land mobile antennas for the same frequency. The VHF-FM whip (Fig. 25-12) can be several quarter-wavelengths and take advantage of gain characteristics thereby obtained. Whips for the HF bands tend to be 10 to 30 ft in length, and often look like trolling rods on power boats. You will also see Citizens Band radios on board boats, and the CB antennas are also whips. It is not a good idea to rely solely on CB for boat communications, because it is a lot less



25-11 Sailboat grounding.

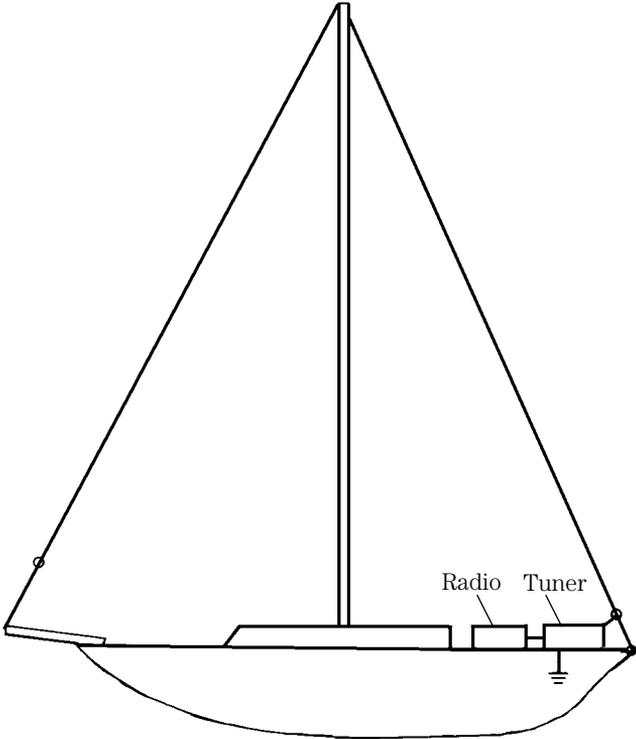
**25-12** VHF-FM boat antenna.  
Courtesy of Antenna Specialist Company



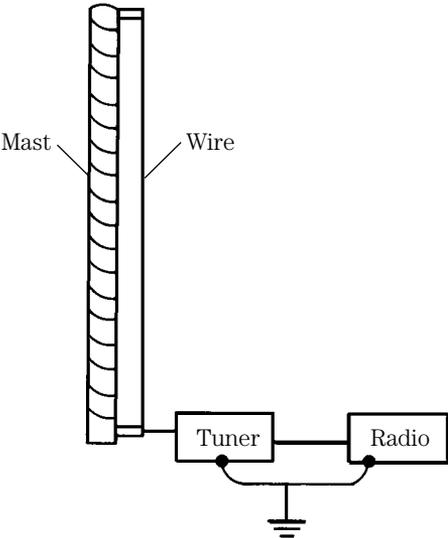
likely to be useful in emergencies. Also, some boaters use amateur radio sets, often illegally, for emergency communications in lieu of a proper HF-SSB unit.

Longwire antennas also find use in marine service. Figure 25-13 shows two installations. The antenna in Fig. 25-13A shows a wire stretched between the stern and bow by way of the mast. The antenna is end-fed from an antenna tuner or “line flattener.” The longwire shown in Fig. 25-13B is similar in concept, but runs from the bottom to the top of the mast. Again, a tuner is needed to match the antenna to the radio transmission line. Notice that the antennas used in this manner are actually not “longwires” in the truly rigorous sense of the term, but rather “random length” antennas.

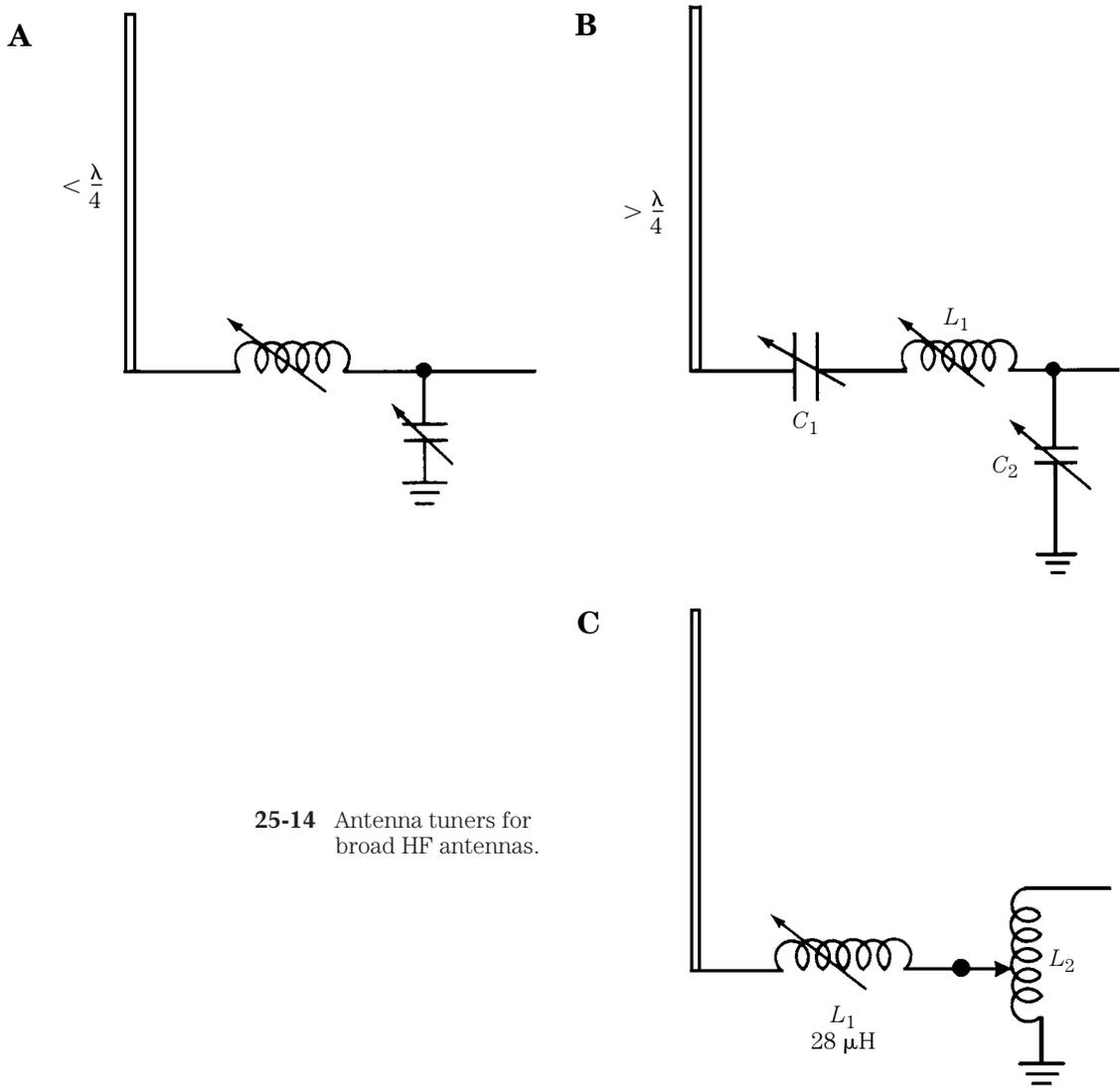
The format of the tuner can be any of several designs, shown in Fig. 25-14. The reversed L-section coupler shown in Fig. 25-14A is used when the antenna radiator element is less than a quarter-wavelength. Similarly, when the antenna is greater than a quarter-wavelength, the circuit of Fig. 25-14B is the tuner of choice. This circuit is a modified L-section coupler that uses two variable capacitors and the inductor.



**25-13A** Sailboat HF antenna.



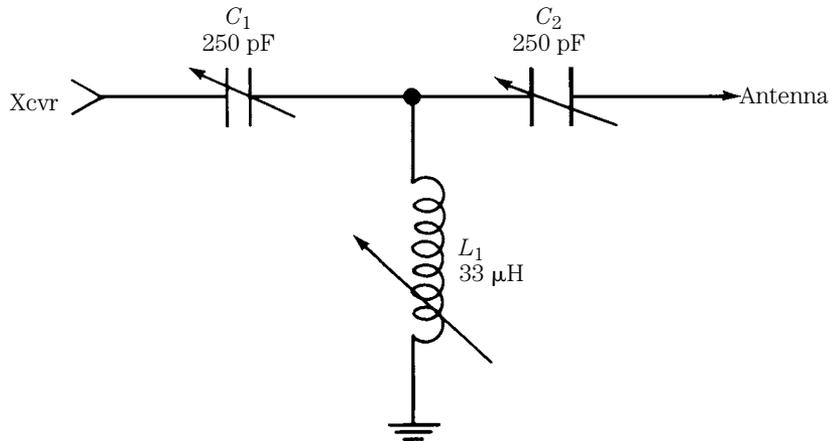
**25-13B** Mast antenna.



25-14 Antenna tuners for broad HF antennas.

Finally, we see in Fig. 25-14C the coupler used on many radios for random-length antennas. Two variable inductors are used;  $L_1$  is used to resonate the antenna, and  $L_2$  is used to match the impedance looking back to the transmitter to the system impedance. In some designs, the inductors are not actually variable, but rather use switch-selected taps on the coils. The correct coil taps are selected when the operator selects a channel. This approach is less frequently encountered today, when frequency synthesizers give the owner a selection of channels to use. In those cases, either the antenna must be tuned every time the frequency is changed, or an automatic (or motor-driven) preselected tuner is used.

The line flattener (Fig. 25-15) is a standard transmatch antenna tuner that flattens coaxial-cable transmission line for VSWR. This type of tuner is especially useful for transmitters with solid-state finals that are not “happy” with high VSWR. Some of those designs incorporate shutdown circuits that reduce (and then cut off) power as the VSWR increases. The line flattener basically tunes out the VSWR at the transmitter. It does nothing to tune the antenna, but only makes the transmitter operable.



25-15 Line flattener tuner.

## 26 CHAPTER

# Antennas for low-frequency operation

LOW-FREQUENCY OPERATION (E.G., 160- AND 75/80-METER BAND) POSES CERTAIN difficulties for the antenna. Of course, the first thing that springs to mind is the large size of those antennas. A half-wavelength dipole is between 117 and 133 ft long on 75/80 m, and on 160 m it is about twice that length. On my own suburban lot, I cannot erect a half-wave 75/80-m band antenna and stay within the property lines.

A similar situation is seen with vertical antennas. Although a 40-m vertical (33 ft high) is not an unreasonable mechanical job, the 66 ft 75/80-m vertical (never mind the 120-ft 160-m vertical) is a nightmare. In addition, the local authorities might not require any special inspections or permits (check!) on the 33-ft antenna, yet impose rigid and very exacting requirements on the higher structure. On suburban or urban lots, a typical 40-m antenna might well be able to fall over and still not cross the property line—or come in close proximity to power lines. A longer antenna, however, almost inevitably suffers one—or the other—problem when it falls.

## Grounding

Still another problem involves grounding. The ground system of a higher-frequency vertical antenna can be installed using  $\lambda/4$  radials buried a few inches below the surface. But that same idea becomes problematic when the radials are 66 or 125 ft long. Burying half a dozen 125-ft radials might remind your neighbors of the Galloping Gung-Ho Gopher—and not to their amusement!

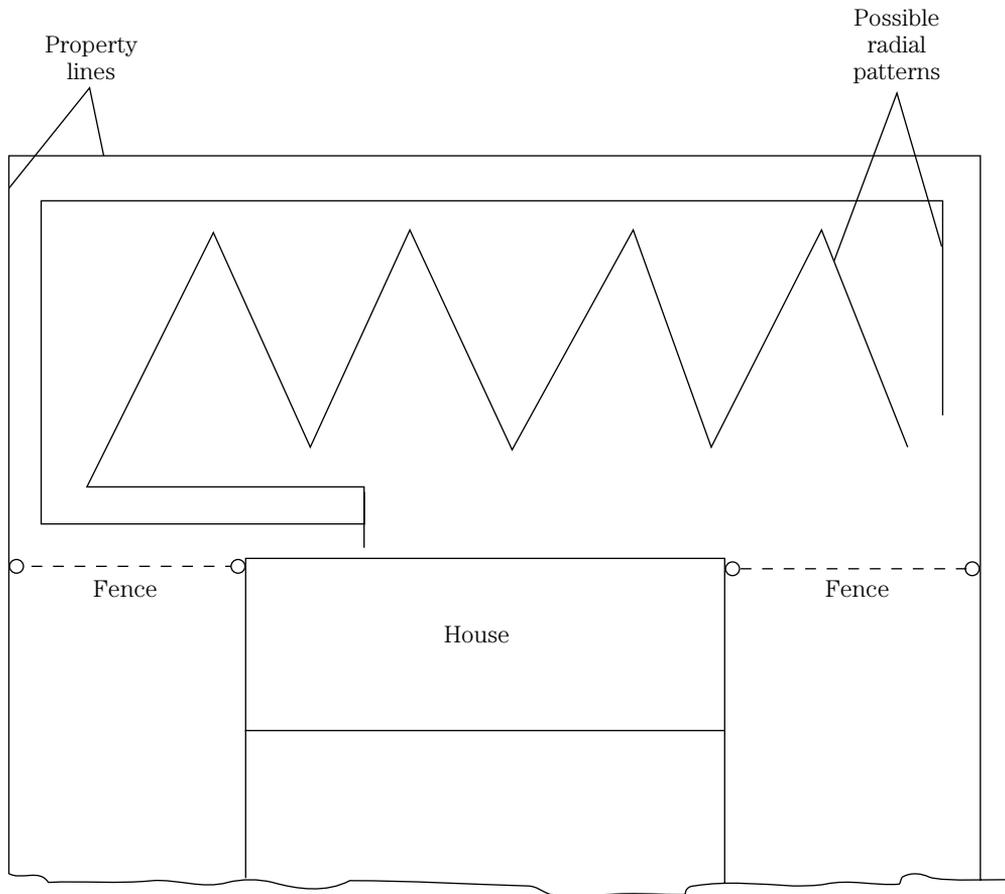
There are, however, some solutions to the problem—without having you buy a farm in the flatter regions of the midwest. For example, the length problem can be solved by using one of the bent dipoles covered in Chap. 10. No, they are not

as effective as an antenna that is correctly installed. But they will serve to get you on the air. Some of them work surprisingly well.

The ground radial problem can be solved by bending the radial system around your property (Fig. 26-1). You will never have to cross your neighbor's line. And, as for the gopher track appearance, it is not necessary to work sloppily, and you can install radials so that nary an eye, practiced or otherwise, can see their location.

## Shortened vertical antennas

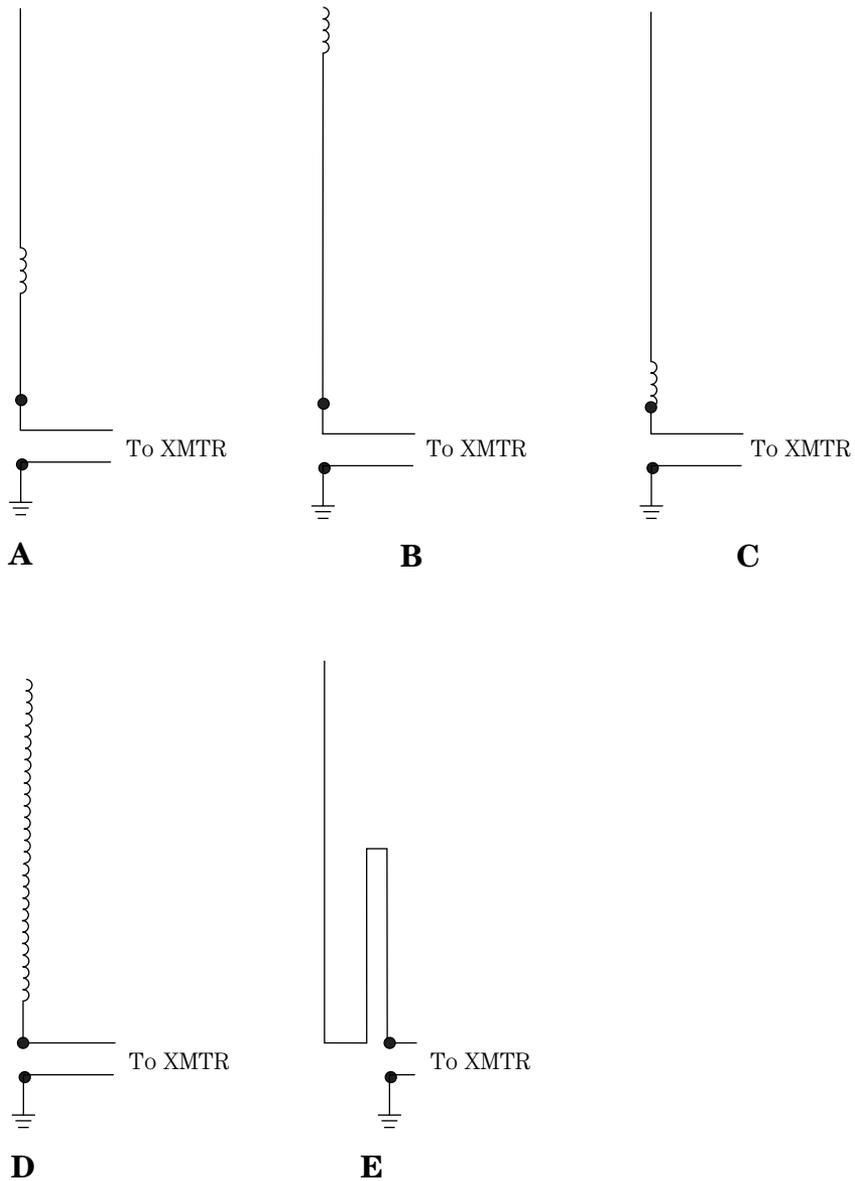
The biggest problem for most low-frequency DXers, as you have seen, is the excessive size of antennas for those frequencies; it is “not for nothing” that those AM broadcast band (<1.6 MHz) towers are usually hundreds of feet tall. But there are ways to shorten an antenna—not for free, because the TANSTAAFL\* principle still



**26-1** Methods for laying out low-frequency radials in a cramped space.

\*TANSTAAFL = There ain't no such thing as a free lunch!

applies—to a point where it becomes mechanically possible. Let me reiterate once again that these compensation antennas will not work as well as a properly installed full-sized antenna, but they will serve to get you on the air on frequencies where it is otherwise utterly impossible. Several different compensation configurations are popular, and these are shown in Fig. 26-2.



**26-2** Inductance loading of vertical antennas: (A) center; (B) top; (C) base; (D) linear or length-loaded; (E) linear hairpin-loaded.

The basic foundation for these antennas is a very short vertical antenna. The “standard” vertical antenna is a quarter-wavelength ( $\lambda/4$ ) (i.e.,  $90^\circ$  electrical length), and it is unbalanced with respect to ground.

Recall that a vertical antenna that is too short for its operating frequency (i.e., less than  $\lambda/4$ ) will exhibit capacitive reactance. In order to resonate that antenna, it is necessary to cancel the capacitive reactance with an equivalent inductive reactance, such as  $|X_L| = |X_C|$ . By placing an inductance in series with the antenna radiator element, therefore, we can effectively “lengthen” it electrically. Of course, what is really happening is that the effects of the lower operating frequency are being accommodated (i.e., “cancelled out”) of a too-short antenna.

An antenna that is reactance-compensated for a different frequency is said to be *loaded*, or in the case of a very low frequency antenna, *inductively loaded*. Three basic forms of loading are popular: *discrete loading*, *continuous loading*, and *linear loading*.

Discrete loading means that there is a discrete, or lumped, inductance in series with the antenna radiator (Figs. 26-2A through 26-2C). These antennas are so constructed that a loading coil is placed at the center (Fig. 26-2A), top (Fig. 26-2B), or bottom (Fig. 26-2C) of the radiator element.

You will recognize these configurations as being the same as those found on mobile antennas. Indeed, low-band mobile antennas can be used in both mobile and fixed installations. Note, however, that although it is convenient to use mobile antennas for fixed locations (because they are easily available in “store bought” form); they are less efficient than other versions of the same concept. The reason is that the mobile antenna, for low frequencies, tends to be based on the standard 96- to 102-in whip antenna used by amateur operators on 10 m or Citizens Band operators on 11 m. In fixed locations, on the other hand, longer radiator elements (which are more efficient) are more easily handled. For example, a 16- to 30-ft high aluminum radiator element can easily be constructed of readily available materials. The 16-ft element can be bought in the form of one or two lengths of 1- to 1.5-in aluminum tubing at do-it-yourself outlets.

A problem seen with these antennas is that they tend to be rather high  $Q$ , so the bandwidth is necessarily narrow. An antenna might work in the center of a band, but present a high VSWR at the ends of the band, and thus be unusable. This problem is solved by making the inductor variable so that slightly different inductance values can be selected at different frequencies across the band. For the base-loaded version (Fig. 26-2C), this is particularly easy: a rotary inductor (perhaps motor driven for remote operation) can be used. For the other configurations, a tapped fixed inductor can be used instead. Each tap represents a different inductance value. Either clip connectors, or relay connections, can be used to select which tap is used.

A continuously loaded antenna has the inductance distributed along the entire length of the radiator (Fig. 26-2D). Typical of these antennas is the helically wound verticals in which about a half-wavelength of insulated wire is wound over an insulating form (such as a length of PVC pipe or a wooden dowel); the turns of the coil are spread out over the entire length of the insulated support.

*Linear loading* (Fig. 26-2E) is an arrangement whereby a section of the antenna is folded back on itself like a stub. Antennas of this sort have been successfully

built from the same type of aluminum tubing as regular verticals. For 75/80-m (3.75-MHz) operation a length of 30 ft for the radiator represents  $41^\circ$ , while a normal  $\lambda/4$  vertical is  $90^\circ$ . The difference between  $90^\circ$  and  $41^\circ$  of electrical length is made up by the “hairpin” structure at the base.

## Shortened horizontal antennas

The same strategies that worked for vertical antennas also work for horizontal antennas, although in the horizontal case we are simulating a half wavelength ( $180^\circ$ ) balanced antenna—rather than a  $\lambda/4$  unbalanced antenna. Figure 26-3 shows several different shortened, low frequency antennas that are based on the same methods as the verticals shown previously.

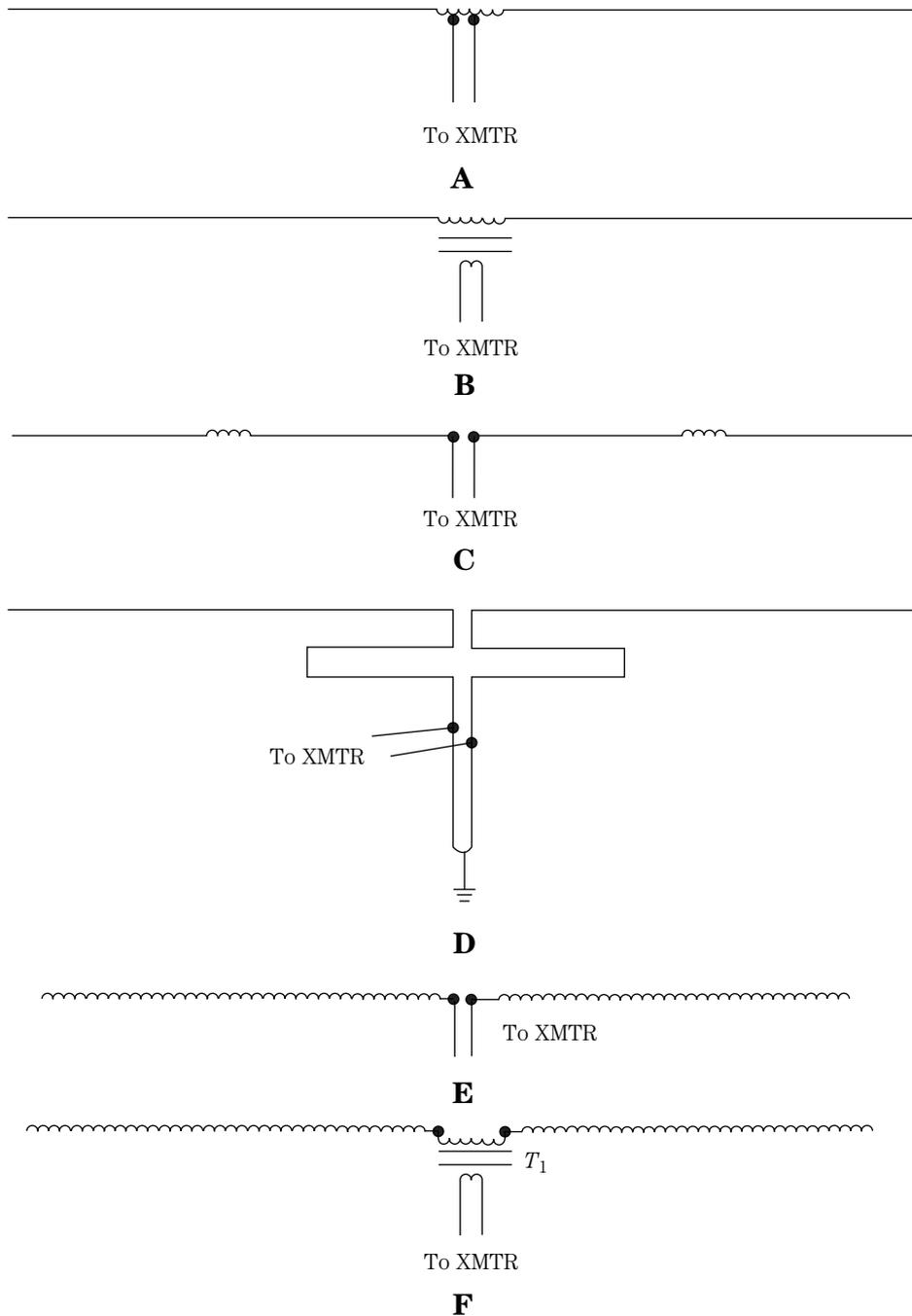
Figures 26-3A through 26-3C show discretely loaded dipole antennas. Figures 26-3A and 26-3B are center-loaded versions, and Fig. 26-3C is a center (of each element) -loaded version. In each case, it is assumed that the radiator elements are the same physical length. In the example of Fig. 26-3A, the coil is tapped to provide a match to either 52- or 75- $\Omega$  transmission line. In some variants, the shield of the coaxial-cable transmission line is connected directly to the junction of the coil and one radiator element, and the coax center conductor is connected to the tap that best matches the impedance of the line.

The impedance-matching problem is solved a little differently in Fig. 26-3B. In this type of antenna, the coil is used to center-load the dipole, but the transmission line is connected to a link wound on the same form as the loading inductor. The turns ratio between the loading inductor and the coupling link determines the impedance match.

A “hairpin” linear loading scheme is shown in Fig. 26-3D. This design is basically the same as for the vertical case, but it is balanced out of respect for the design of the dipole. This type of design is used for both the driven and parasitic elements on some commercial 40-m beam antennas. In addition, some “add-on” 40-m dipoles (designed for beam antennas that were intended for operation at frequencies in the 20-m and higher bands) use this method. Such antennas are essentially rotatable dipoles.

The continuously loaded dipole of Fig. 26-3E is constructed like the helical vertical (i.e., about a half-wavelength of insulated wire is wound over the entire length of an insulated rod or pipe of some sort). The winding can be either broken at its center point to accommodate the feedline, or link-coupled to the transmitter, or receiver, as shown in Fig. 26-3F. Of the two feed methods, the most popular appears to be the type that breaks the winding into two pieces.

The form of continuously loaded dipole in Fig. 26-3E has a combination of the two methods of feed. The distributed loading coil is broken into two sections, as is often done on continuously loaded dipoles. But in this case, a portion of the overall inductance is made by using a discrete inductor that is part of a toroidal transformer  $T_1$ . The length of the wire used for the continuous loading coil is somewhat shorter (find by experimentation) than otherwise would be the case because of the inductance of the transformer secondary. Transformer  $T_1$  can be wound using the same sort of toroidal ferrite or powdered iron core used for Balun transformers.



**26-3** Inductance loading of a dipole: (A) center loading with impedance tapped coil; (B) center loading with transformer coupling; (C) loading at center of elements; (D) linear loading (hairpin); (E) linear or length loading; (F) transformer coupling.

## Loaded tower designs

Many amateur radio stations and other services use a rotatable beam antenna on top of a tower. Typical towers are from 30 to 120 ft in height. In my locality, towers to 30 ft, or less than 3 ft above the roof line, whichever is taller, can be erected without a permit as long as they are attached to the house and will not fall onto a power line, or across the property line, in a catastrophic failure. Higher towers must be erected under a mechanical permit, and properly inspected by the county. These towers typically support a two-element (or more, three is popular) Yagi/Uda beam antenna, or a quad, or some other highly directional antenna. But they can also be treated as a vertical antenna under the right circumstances.

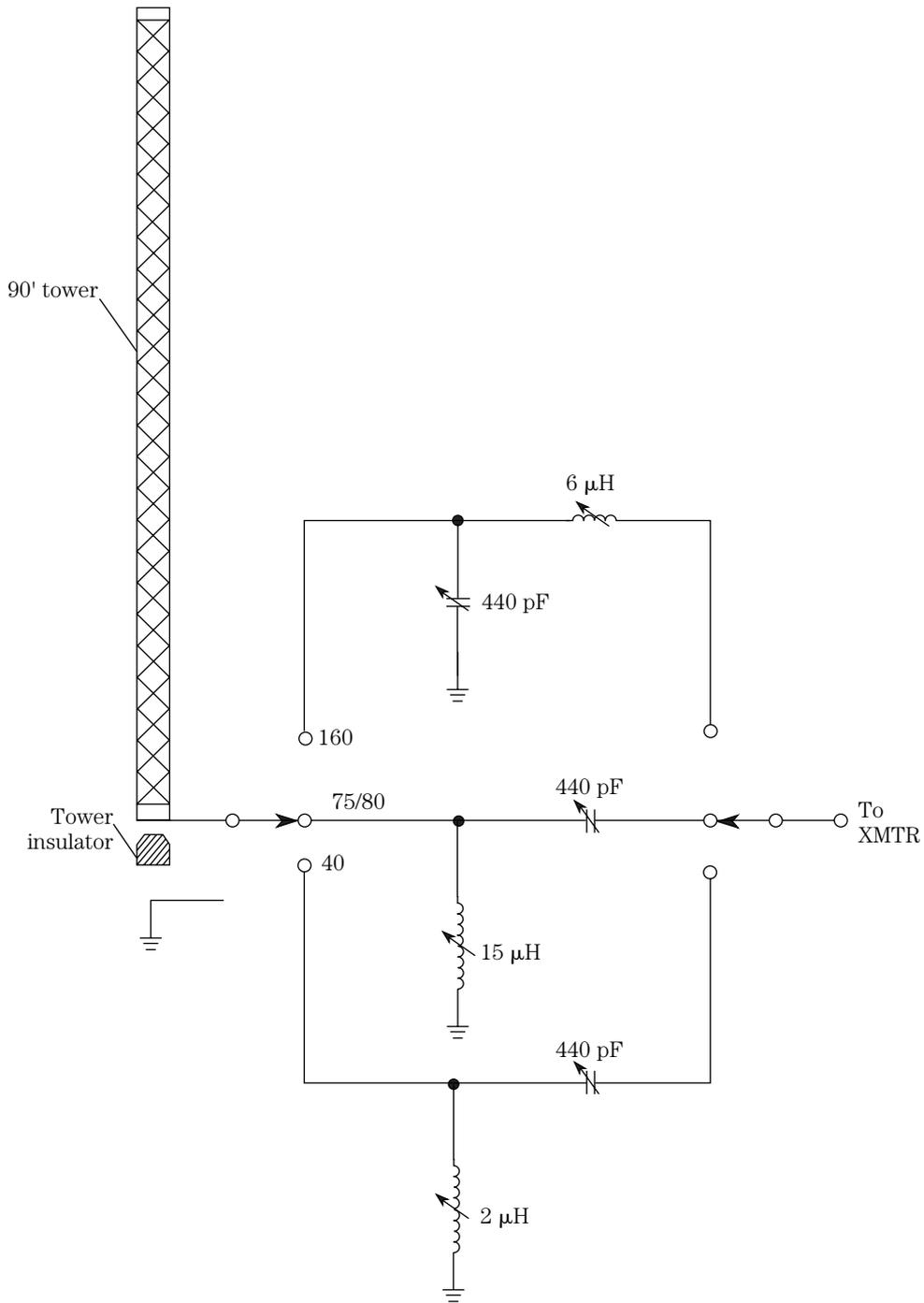
If the tower is close to 66 ft high (which is a popular height for amateur radio towers), then the tower (already  $\lambda/4$ ) can be used as a resonant vertical on 75/80 m. The same tower can also be used on lower frequencies if proper antenna tuning unit (ATU) components are provided.

Figure 26-4 shows a situation in which an 80- to 110-ft tower is insulated from ground, and is considered a random-length vertically polarized Marconi antenna. Like other such antennas, it is fed by a simple L-section, or reverse L-section, ATU. Note the approximate values of the inductors and capacitors in each configuration, and also their relationships. On the 40- and 75/80-m versions, the tower is too long, so a series capacitance and shunt inductance are used for the ATU (i.e., Reverse L-section coupler). On the other hand, the tower is too short for 160 m, so the inductor and capacitor are reversed.

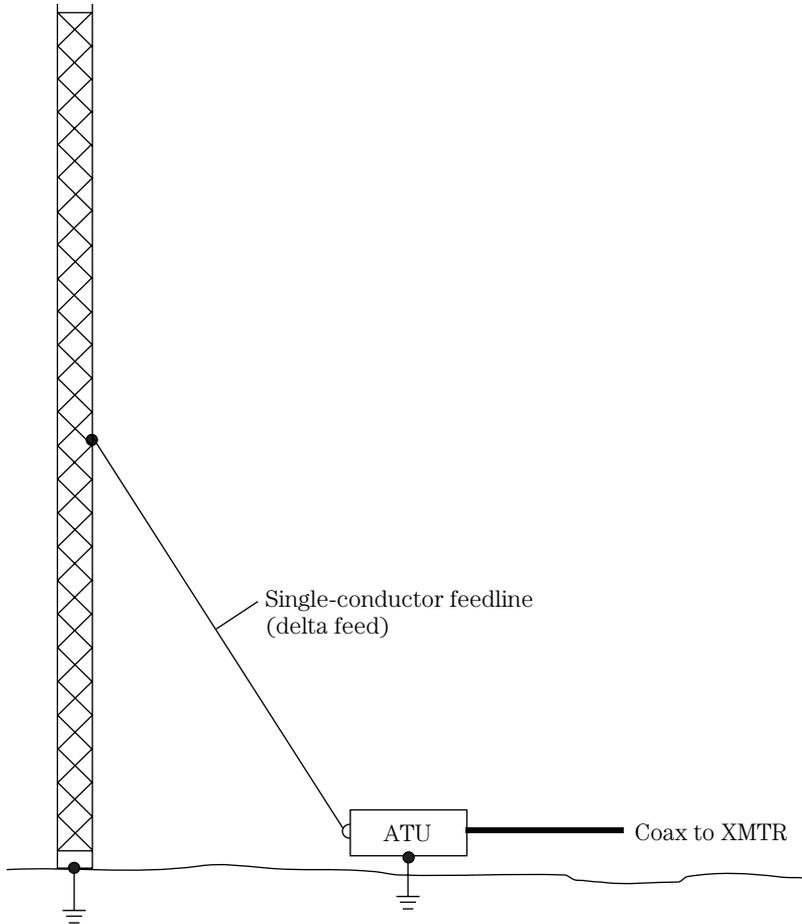
There is only one small problem with the design of Fig. 26-4: Compared with grounded towers, insulated towers are expensive to buy and install. In addition, lightning protection is probably better in the grounded tower. So how do we work a grounded tower? See Fig. 26-5. In this antenna, the tower is grounded at its base. If the tower is mounted to a concrete pedestal (the usual arrangement), then a separate ground rod and ground wire adjacent to the pedestal must be provided. Concrete is not a good insulator, but it is also not a good conductor. A *delta feed* system is used with this antenna. That is, a single wire from the ATU is connected to a point on the antenna where an impedance match can be achieved. It is important that the ATU be spaced away from the tower base (as shown), and the wire must be run straight to the feedpoint on the antenna. This system is an example of several possible *shunt feeding* systems.

## Random-length Marconi

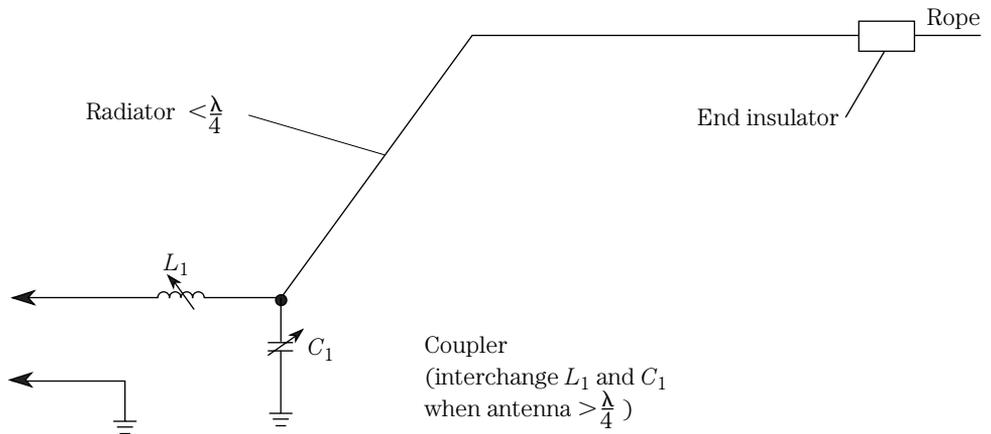
One of the old standbys for all bands is the random-length Marconi antenna. Such antennas consist of a length of wire, typically (but not always) less than  $\lambda/4$ , and fed at one end with coaxial cable and an L-section coupler (Fig. 26-6). The antenna radiator element can be angled in any direction as needed, but it works best if the radiator is either as horizontal, or vertical, as possible (for pure polarization). The usual situation, however, is to run the wire at an inclined angle, or with about equal portions horizontal and vertical (see the inverted L) which yields complex polarization.



**26-4** Method for impedance matching a tower antenna.



26-5 Delta matching a grounded vertical tower antenna.



26-6 Random-length wire antenna.

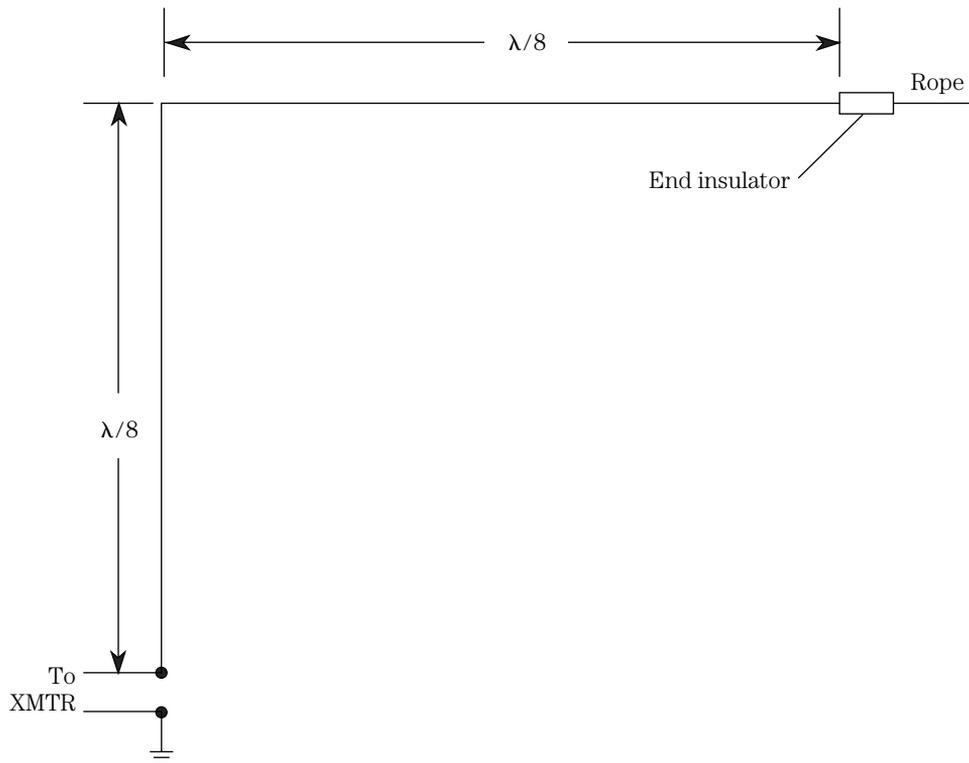
The L-section coupler shown in Fig. 26-6 is set up for the case where the antenna radiator element is less than  $\lambda/4$  (i.e., a series inductor and a shunt capacitor). If the antenna is longer than  $\lambda/4$  on some frequency, then reverse the positions of the capacitor and inductor.

### Inverted-L antennas

Another popular antenna for low frequencies is the  $\lambda/4$  *inverted-L* (Fig. 26-7). In this type of antenna, two sections are erected at a  $90^\circ$  angle with respect to each other; one vertical and the other horizontal. One way to think of this antenna is bent  $\lambda/4$  vertical, although some people liken it to a top-loaded vertical. The feedline can be 52- or 75- $\Omega$  coaxial cable.

It is generally the case that the sections of the inverted-L are equal in length ( $\lambda/8$  each), but that is not strictly necessary. As the vertical section becomes longer (with overall length remaining at  $\lambda/4$ ), the angle of radiation depresses.

One popular method of construction is to use a tower for the vertical section, and a run of wire for the horizontal section. If you already have a 60-ft tower to accommodate the beam antenna used on higher frequencies, then it is relatively easy to build an inverted-L antenna for 160 m.



26-7 L-section quarter-wavelength antenna.

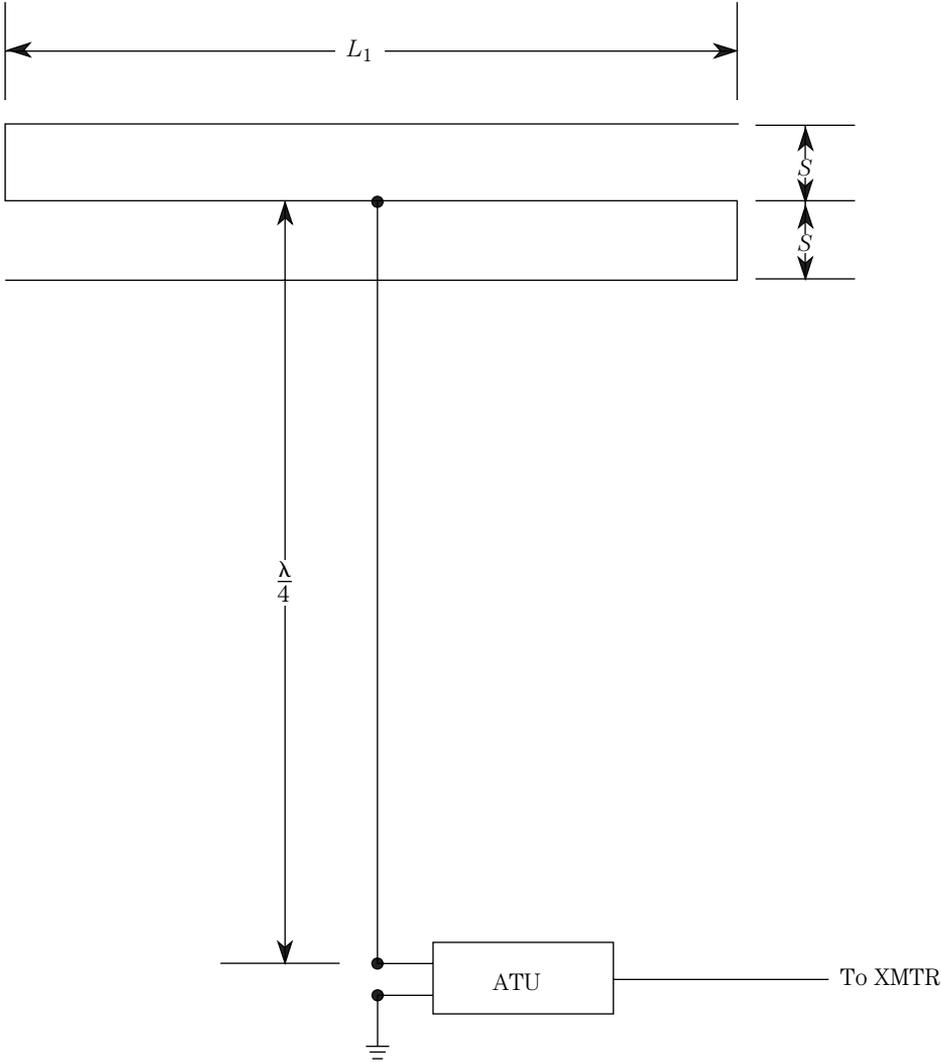
### The linearly loaded “tee” antenna

A “tee” antenna consists of a horizontal radiator element fed at the center with a single conductor wire (not coax) that is a quarter-wavelength long, and goes away from the radiator at a right angle. Such antennas were popular prior to World War II in the United States, although popularity fell off in the late 1930s as other types (and reasonably priced transmission lines) became available. The tee is possibly one of the oldest forms of radio transmitting antennas. Like other antennas at low frequencies, however, it is far too long for easy use in most locations.

We can, however, linearly load the tee by folding the ends of the radiator back on themselves to form the zigzag pattern of Fig. 26-8. Popular in Europe for some time, this antenna can provide reasonable performance on the lower frequencies without using too much horizontal space.

The radiator element consists of three sections that are each about  $\lambda/6$  long ( $L_1 = 164/F_{\text{MHz}}$ ), and spaced 8 to 12 in apart; they are parallel to each other. The quarter-wavelength transmission line is connected to the center point of the middle section of the radiator element.

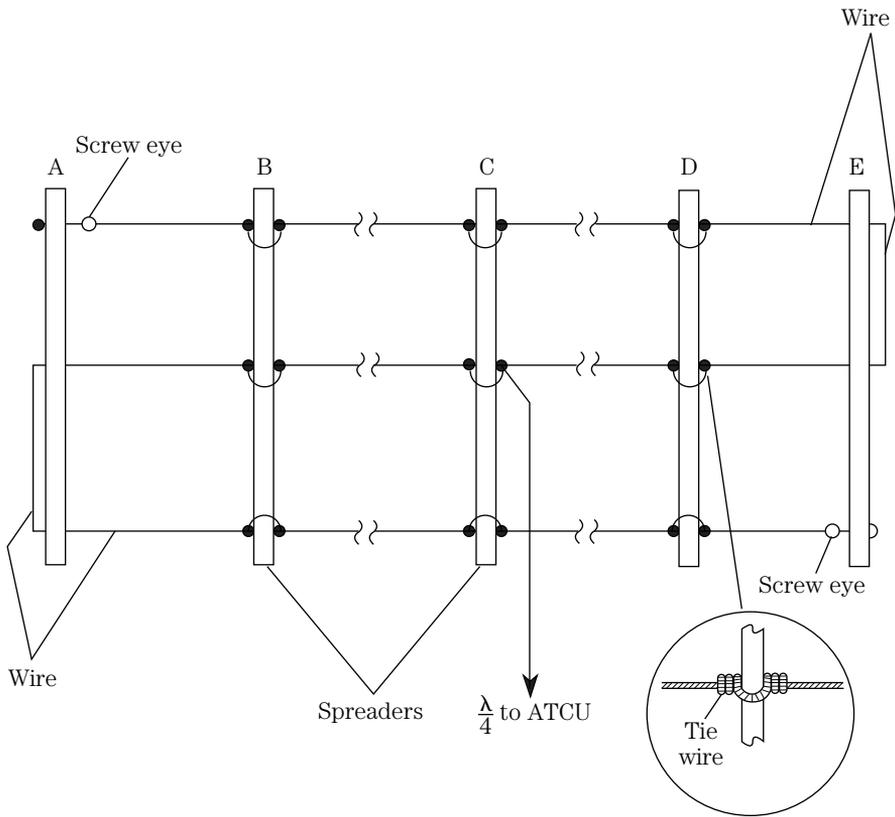
Figure 26-9 shows how the linearly loaded tee antenna could be built using wire for the radiator elements. Spreaders made of 24-in-long  $1 \times 2$ -in lumber (treated against the weather), plastic, or some other synthetic material can be used. Each spreader has three holes drilled in it spaced about 10 in apart. Spreaders A and E are end spreaders, and are identical—except for being upside down with respect to each other. In each case, one of the radiator element conductor sections is terminated, while the other passes through to the back of the spreader to join the center radiator element. The other spreaders are used either for center support, as in C (note the transmission line attachment), or for interim support between the center insulator and the end insulators.



$$L_1 = \frac{164}{F_{\text{MHz}}}$$

$$\frac{\lambda}{4} = \frac{246}{F_{\text{MHz}}}$$

26-8 Tee-loaded antenna.



26-9 Shortened tee antenna.

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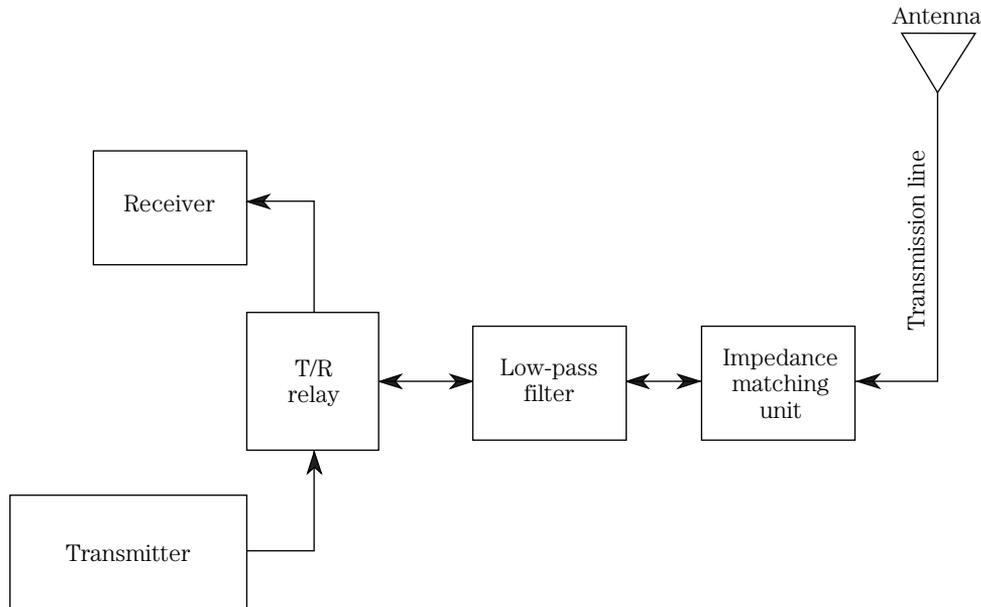
## 27 CHAPTER

# Measurements and adjustment techniques

THIS CHAPTER EXAMINES SOME OF THE INSTRUMENTS AND TECHNIQUES FOR TESTING antenna systems, whether brand new installations, or in troubleshooting situations on older antenna. The basic radio system is shown in Fig. 27-1. We have several elements in the system: transmitter, low-pass filter, impedance-matching unit, a coaxial relay if the receiver is separate from the transmitter, and the antenna. Connecting these elements are lengths of transmission line. In most modern radio systems below the microwaves, the transmission line is coaxial cable.

The low-pass filter and matching unit may be considered optional by some, but some designers regard them as essentially standard equipment, especially for HF and low-band VHF systems. The low-pass filter has the job of removing harmonics from the output signal that could interfere with other radio systems or radio and TV broadcast reception. It will pass only those frequencies below a certain cutoff point. The impedance matching unit is used to “tune out” impedance mismatches that cause the transmitter to see an excessive VSWR. These units are common in VLF, MW, and HF communications stations, as well as AM broadcasting stations. The matching unit also provides additional attenuation of the harmonics, so it makes the output even cleaner than is possible with the low-pass filter alone.

Another reason to use the matching unit is to allow the radio transmitter to put out the maximum allowable RF power. Modern solid-state final amplifiers are not tolerant of VSWR. In addition, these units use fixed-tuned low-pass filters for each band rather than the wide-range pi networks common on vacuum tube transmitters. Solid-state transmitters usually include a feedback *automatic load control* (ALC) that reduces output power when a high VSWR is sensed. The VSWR cut-in knee begins around 1.5:1 and completely shuts off the transmitter when the VSWR gets high (typically above 2.5:1 or 3:1). The external T/R relay is not used on most modern systems because the receiver and transmitter are typically housed inside of the same box. The coaxial relay was used in days past, when



**27-1** Basic communications radio station setup.

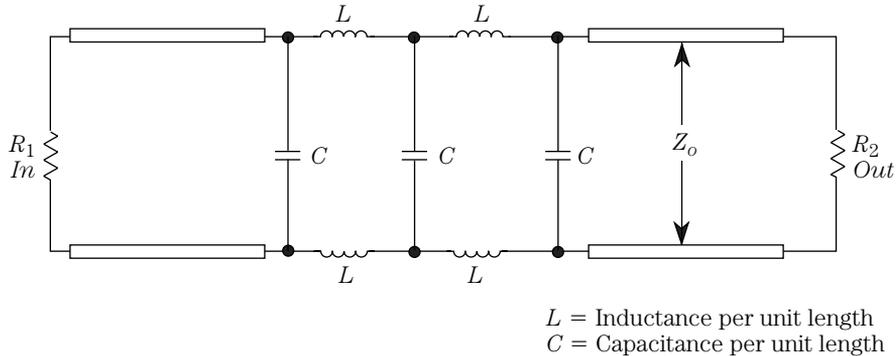
a separate receiver needed to use the same antenna. In addition, many modern solid-state equipments use PIN diode T/R switching.

## Transmission lines

The transmission line is not merely a wire that carries RF power to the antenna. It is actually a complex circuit that simulates an infinite  $LC$  network. There is a characteristic impedance  $Z_o$ , also called *surge impedance*, which describes each transmission line. This impedance is the square root of the ratio of the capacitance and inductance per unit of length. When a load having a “resistive-only” impedance equal to the surge impedance of the transmission line is connected, then we will see maximum transfer of power between the line and the antenna.

We cannot deal extensively with transmission-line theory here, and refer the reader instead to previous chapters. We must, however, have at least some idea of what the circuit looks like. Figure 27-2 shows a model of a transmission line in which  $Z_o$  is the surge impedance of the line,  $R_1$  is the load impedance of the antenna, and  $R_2$  is the output impedance of the transmitter. In a properly designed system all three impedances either will be equal ( $Z_o = R_1 = R_2$ ) or a matching network will make them equal.

We must consider the electrical situation along the transmission line in order to understand the readings that we see on our instruments. Figure 27-3 shows several possible situations. These graphs are of the RF voltage along the line, with voltage on the vertical axis and transmission line length (expressed in wavelengths of the RF signal) along the horizontal axis. When the system is matched ( $Z_o = R_1$ ), the voltage



27-2 Equivalent circuit of transmission line.

is the same everywhere along the line (Fig. 27-3A). This line is said to be “flat.” But when  $Z_o$  and  $R_2$  are not equal, then the voltage varies along the line with wavelength. In mismatched systems, not all of the power is radiated by the antenna, but rather is reflected back to the transmitter. The forward and reflected waves combine algebraically at each point along the line to form standing waves (Fig. 27-3B). We can plot the voltage maxima ( $V_{\max}$ ) and minima ( $V_{\min}$ ). Keep this graph in mind for a few minutes because you will refer again to it when we deal with VSWR.

Two special situations occur in transmission line and antenna systems that yield similar results. The entire forward power is reflected back to the transmitter (none radiated) if the load (i.e., antenna) end of the transmission line is either open or shorted. The voltage plot for an open transmission line ( $R_2$  is infinite) is shown in Fig. 27-3C, and that for the shorted line is shown in Fig. 27-3D. Note that they are very similar to each other except for where the minima ( $V_{\min} = 0$ ) occur. The minima are offset from each other by  $90^\circ$  (i.e., quarter-wavelength).

## Calculating standing wave ratio

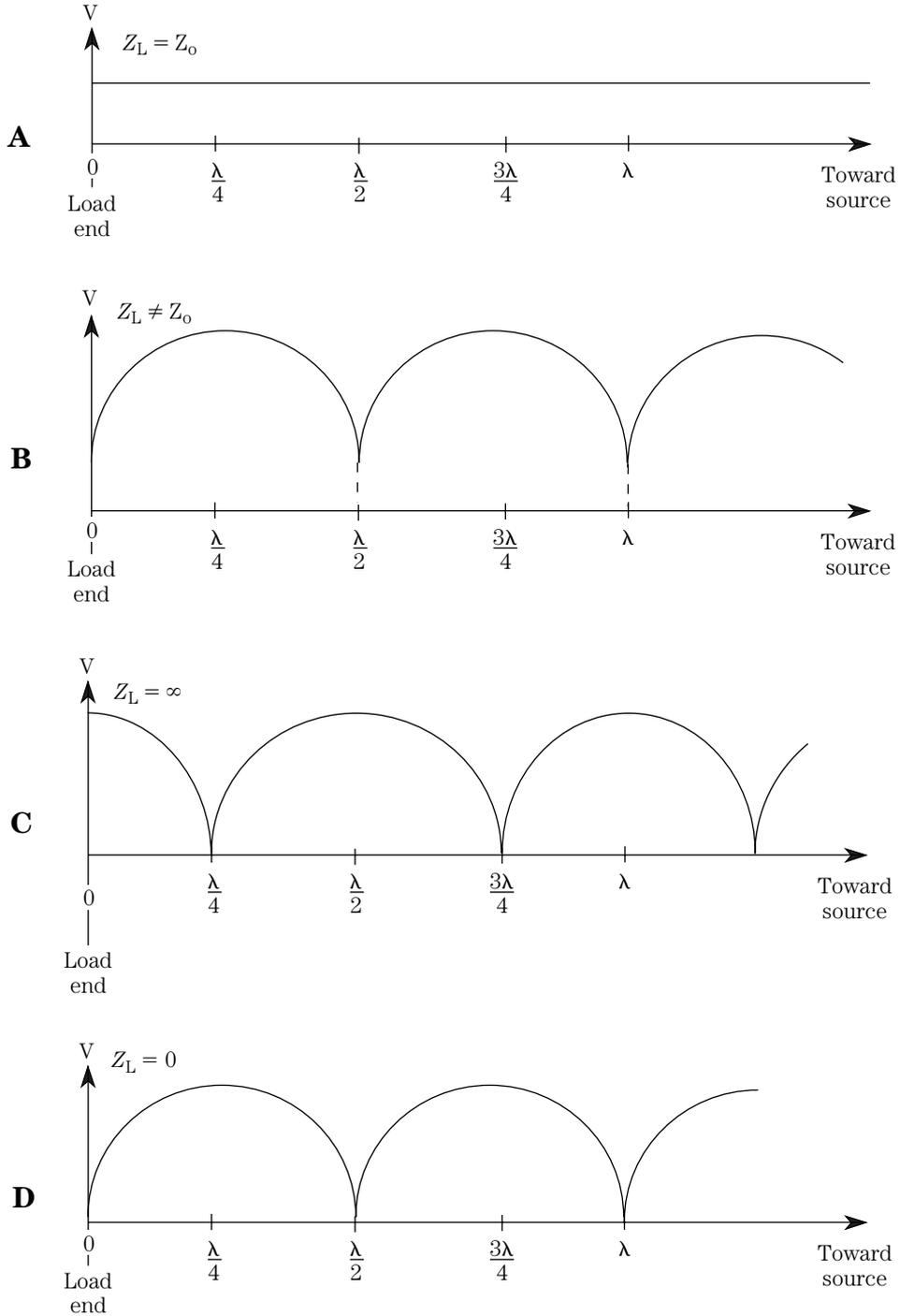
The VSWR can be calculated from any of several bits of knowledge. Even if you don't have a VSWR meter, therefore, it is possible to determine VSWR. If the antenna load impedance ( $R_2$ ) is not equal to  $Z_o$ , then we can calculate VSWR from one of the following:

If  $Z_o$  is greater than  $R_2$ :

$$\text{VSWR} = \frac{Z_o}{R_2} \quad [27.1]$$

If  $Z_o$  is less than  $R_2$ :

$$\text{VSWR} = \frac{R_2}{Z_o} \quad [27.2]$$



27-3 Voltage along line versus line electrical length: (A) flat line with matched impedance; (B) unequal impedances; (C) open load; (D) shorted load.

We can also measure the forward and reflected power, and calculate the VSWR from those readings:

$$\text{VSWR} = \frac{1 + \sqrt{(P_r/P_f)}}{1 - \sqrt{(P_r/P_f)}} \quad [27.3]$$

where

VSWR is the voltage standing wave ratio

$P_r$  is the reflected power

$P_f$  is the forward power

If we can measure either the voltage maximum and minimum, or the current maximum and minimum, then we can calculate SWR:

$$\text{VSWR} = \frac{V_{\max}}{V_{\min}} \quad [27.4]$$

Finally, if the forward and reflected voltage components at any given point on the transmission line can be measured, then we can calculate the VSWR from:

$$\text{VSWR} = \frac{V_f + V_r}{V_f - V_r} \quad [27.5]$$

where

$V_f$  is the forward voltage component

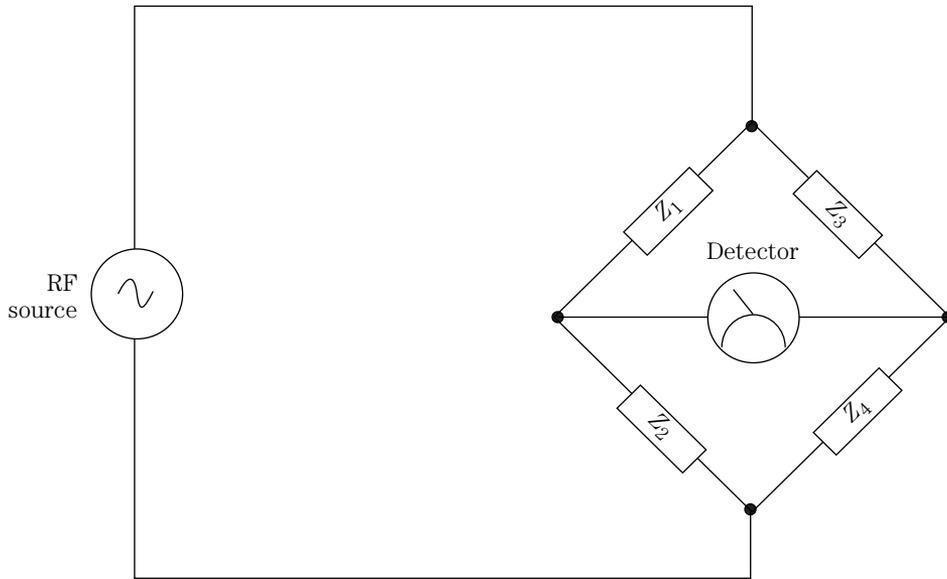
$V_r$  is the reflected voltage

The last equation, based on the forward and reflected voltages, is the basis for many modern VSWR and RF power meters.

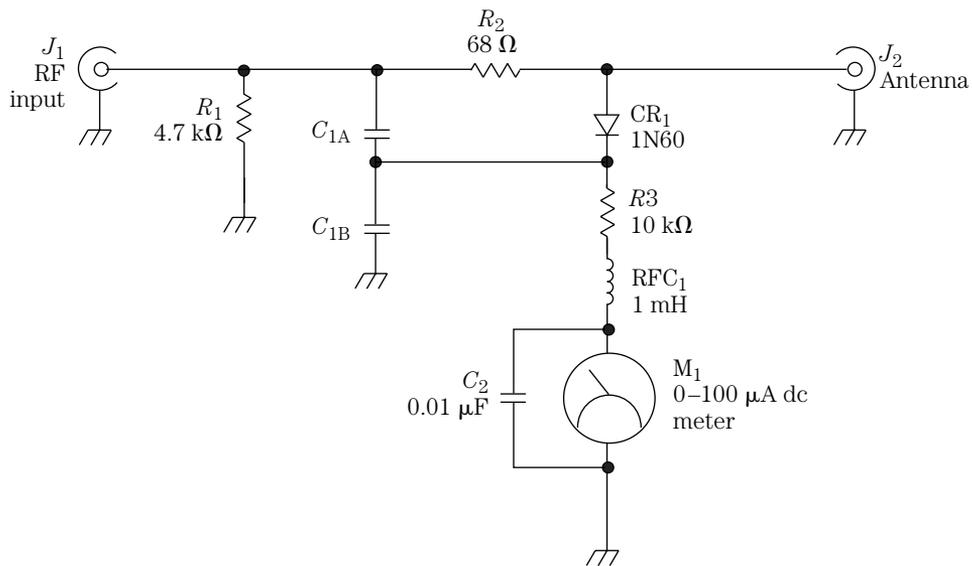
## Impedance bridges

We can make antenna impedance measurements using a variant of the old-fashioned Wheatstone bridge. Figure 27-4A shows the basic form of the bridge in its most generalized form. The current flowing in the meter will be zero when  $(Z_1/Z_2) = (Z_3/Z_4)$ . If one arm of the bridge is the antenna impedance, then we can adjust the others to make the bridge null to make the measurement. A typical example is shown in Fig. 27-4B. The antenna connected to  $J_2$  is one arm of the bridge, while  $R_2$  is a second. The value of  $R_2$  should be 50  $\Omega$  or 75  $\Omega$ , depending upon the value of the expected antenna impedance. The choice of 68  $\Omega$  is a good compromise for meters to operate on both types of antennas. The other two arms of the bridge are the reactances of  $C_{1A}$  and  $C_{1B}$ , which is a single differential capacitor. Tune  $C_1$  until the meter is nulled, and then read the antenna from the dial. At least one instrument allows the technician to plug in a resistor element equal to system impedance.

Calibrating the instrument is simple. A series of noninductive carbon composition resistors having standard values from 10 to 1000  $\Omega$  are connected across  $J_2$ . The meter is then nulled, and the value of the load resistor is inscribed on the dial at the point.



**27-4A** Basic Wheatstone bridge.



**27-4B** SWR bridge.

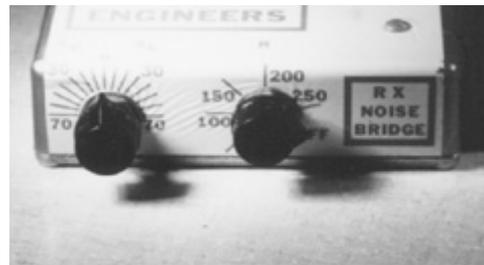
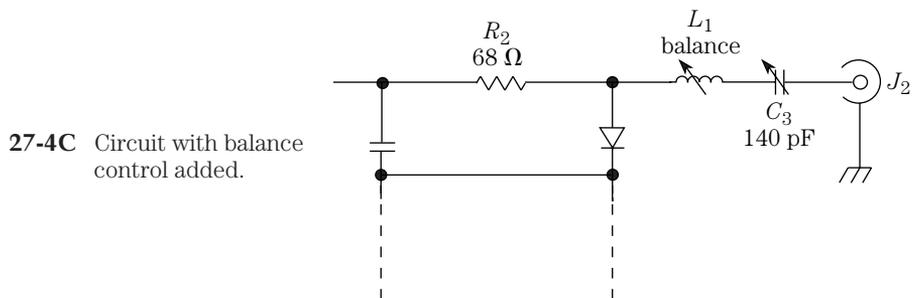
The basic circuit of Fig. 27-4B is useful only to measure the resistive component of impedance. We can modify the circuit as shown in Fig. 27-4C to account for the reactive component. An example of a commercial hand-held radio antenna impedance bridge is shown in Fig. 27-5.

## The RF noise bridge

This section explores a device that was once associated only with engineering laboratories, but turns out to have applications in general communications servicing as well: the *RF noise bridge*. It is one of the most useful, low-cost, and often overlooked test instruments in our armamentarium.

Several companies have produced low-cost noise bridges: Omega-T, Palomar Engineers, The Heath Company, and, most recently, MFJ. The Omega-T and the Palomar Engineers models are shown in Fig. 27-5. The Omega-T device is a small cube with minimal dials, and a pair of BNC coax connectors (*antenna* and *receiver*). The dial is calibrated in ohms, and measures only the resistive component of impedance. The Palomar Engineers device is a little less eye appealing, but does everything the Omega-T does; in addition, it allows you to make a rough measurement of the reactive component of impedance. The Heath Company added their Model HD-1422 to the lineup. And the latest entry is the MFJ-204B antenna bridge.

Over the years, some people have found the noise bridge very useful for a variety of test and measurement applications, especially in the HF and low-VHF regions.

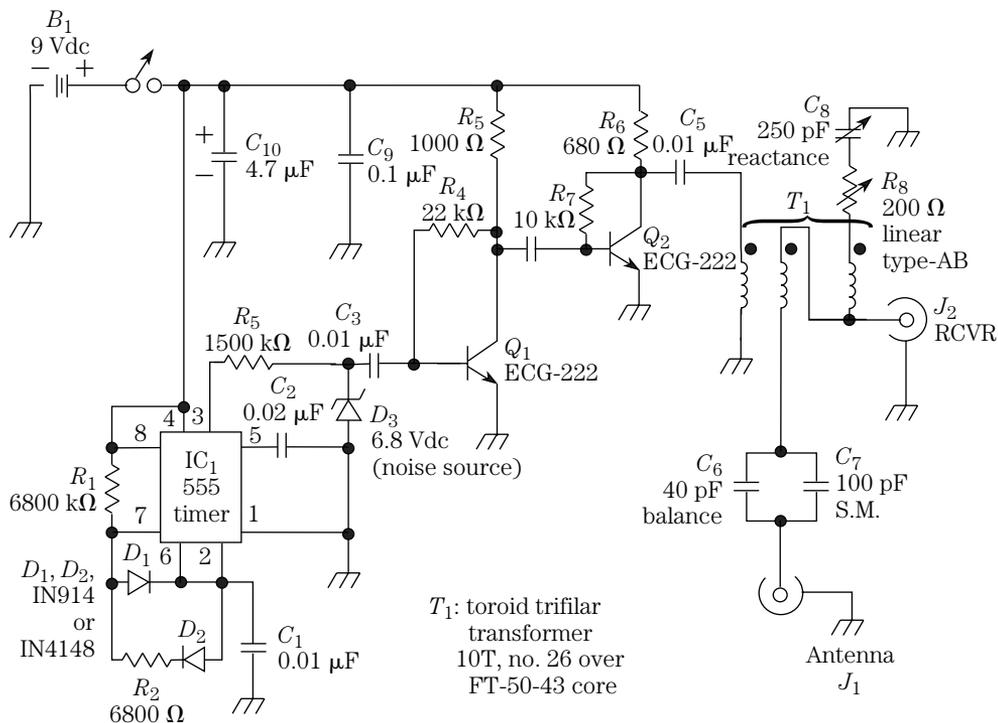


27-5 Commercially available amateur noise bridges.

Those applications are not limited to the testing of antennas (which is the main job of the noise bridge). In fact, although the two-way technician (including CB) or amateur radio operator, will measure antennas, tuned circuits, and resonant cavities with the device, consumer electronics technicians will find other applications.

Figure 27-6 shows the circuit of a noise bridge instrument. The bridge consists of four arms. The inductive arms ( $L_{1b}$  and  $L_{1c}$ ) form a trifilar wound transformer over a ferrite core with  $L_{1a}$ , so a signal applied to  $L_{1a}$  is injected into the bridge circuit. The measurement consists of a series circuit of a 200- $\Omega$  potentiometer and a 120-pF variable capacitor. The potentiometer sets the range (0 to 200  $\Omega$ ) of the resistive component of measured impedance, while the capacitor sets the reactance component. Capacitor  $C_2$  in the unknown arm of the bridge is used to balance the measurement capacitor. With  $C_2$  in the circuit, the bridge is balanced when  $C$  is approximately in the center of its range. This arrangement accommodates both inductive and capacitive reactances, which appear on either side of the "zero" point (i.e., the midrange capacitance of  $C$ ). When the bridge is in balance, the settings of  $R$  and  $C$  reveal the impedance across the unknown terminal.

A reverse-biased zener diode (zeners normally operate in the reverse-bias mode) produces a large amount of noise because of the avalanche process inherent in zener operation. Although this noise is a problem in many other applications, in a noise bridge it is highly desirable: the richer the noise spectrum the better the performance. The spec-



27-6 Noise bridge circuit.

trum is enhanced because of the 1-kHz square-wave modulator that chops the noise signal. An amplifier boosts the noise signal to the level needed in the bridge circuit.

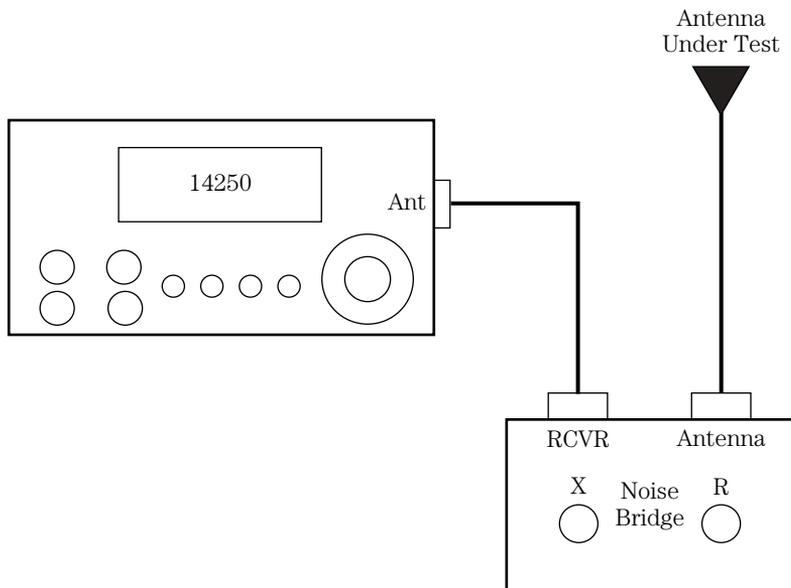
The detector used in the noise bridge is a tunable receiver covering the frequencies of interest. The preferable receiver uses an AM demodulator, but both CW (morse code) and SSB receivers will do in a pinch. The quality of the receiver depends entirely on the precision with which you need to know the operating frequency of the device under test.

### Adjusting antennas

Perhaps the most common use for the antenna noise bridge is finding the impedance and resonant points of an HF antenna. Connect the *receiver* terminal of the bridge to the *antenna* input of the HF receiver through a short length of coaxial cable as shown in Fig. 27-7. The length should be as short as possible, and the characteristic impedance should match that of the antenna feedline. Next, connect the coaxial feedline from the antenna to the *antenna* terminals on the bridge. You are now ready to test the antenna.

Set the noise bridge resistance control to the antenna feedline impedance (usually 50 to 75  $\Omega$  for most common antennas). Set the reactance control to midrange (zero). Next, tune the receiver to the *expected* resonant frequency ( $F_{exp}$ ) of the antenna. Turn the noise bridge *on*, look for a noise signal of about S9 (will vary on different receivers), and if—in the unlikely event that the antenna is resonant on the expected frequency—you will find yourself right in the middle of the null.

Adjust the *resistance* control R on the bridge for a null, i.e., minimum noise as indicated by the S meter. Next, adjust the *reactance* control C for a null. Repeat the adjustments of the R and C controls for the deepest possible null, as indicated by the lowest noise output on the S meter (there is some interaction between the two controls).



27-7 Connection of noise bridge.

A perfectly resonant antenna will have a reactance reading of zero ohms, and a resistance of 50 to 75  $\Omega$ . Real antennas might have some reactance (the less the better), and a resistance that is somewhat different from 50 to 75  $\Omega$ . Impedance-matching methods can be used to transform the actual resistive component to the 50- or 75- $\Omega$  characteristic impedance of the transmission line. The results to expect:

1. If the resistance is close to zero, then suspect that there is a short circuit on the transmission line. Suspect an open circuit if the resistance is close to 200  $\Omega$ .
2. A reactance reading on the  $X_L$  side of zero indicates that the antenna is too long, while a reading on the  $X_C$  side of zero indicates an antenna that is too short.

An antenna that is too long or too short should be adjusted to the correct length. To determine the correct length, we must find the actual resonant frequency  $F_r$ . To do this, reset the *reactance* control to zero, and then *slowly* tune the receiver in the proper direction—downband for too long and upband for too short—until the null is found. On a high- $Q$  antenna, the null is easy to miss if you tune too fast. Don't be surprised if that null is out of band by quite a bit. The percentage of change is given by dividing the expected resonant frequency  $F_{\text{exp}}$  by the actual resonant frequency  $F_r$ , and multiply by 100:

$$\text{Change} = \frac{F_{\text{exp}} \times 100\%}{F_r} \quad [27.6]$$

Connect the antenna, noise bridge, and receiver in the same manner as above. Set the receiver to the expected resonant frequency (i.e., approximately  $468/F$  for half-wavelength types and  $234/F$  for quarter-wavelength types). Set the resistance control to 50  $\Omega$  or 75  $\Omega$ , as appropriate for the normal antenna impedance and the transmission line impedance. Set the reactance control to zero. Turn the bridge on and listen for the noise signal.

Slowly rock the reactance control back and forth to find on which side of zero the null appears. Once the direction of the null is determined, set the reactance control to zero, and then tune the receiver toward the null direction (downband if null is on  $X_L$  side and upband if it is on the  $X_C$  side of zero).

A less-than-ideal antenna will not have exactly 50- or 75- $\Omega$  impedance, so some adjustment of R and C to find the deepest null is in order. You will be surprised how far off some dipoles and other forms of antennas can be if they are not in “free space,” (i.e., if they are close to the earth's surface).

### Nonresonant antenna adjustment

We can operate antennas on frequencies other than their resonant frequency if we know the impedance:

$$X_C = X = \frac{55}{68 - C} - 2340 \quad [27.7]$$

or

$$X_L = X = 2340 - \frac{159,155}{68 + C} \quad [27.8]$$

Now, plug “ $X$ ” calculated from Eq. 27.7 or 27.8 into  $X_f = X/F$  where  $F$  is the desired frequency in MHz.

## Other jobs

The noise bridge can be used for a variety of jobs. We can find the values of capacitors and inductors, determine the characteristics of series- and parallel-tuned resonant circuits, and adjust transmission lines.

**Transmission line measurements** Some antennas and non-noise measurements require antenna feedlines that are either a quarter-wavelength or half-wavelength at some specific frequency. In other cases, a piece of coaxial cable of specified length is required for other purposes: for instance, the dummy load used to service depth sounders is nothing but a long piece of shorted coax that returns the echo at a time interval that corresponds to a specific depth. We can use the bridge to find these lengths as follows:

1. Connect a short circuit across the *unknown* terminals and adjust R and X for the best null at the frequency of interest. (*Note*: both will be near zero.)
2. Remove the short circuit.
3. Connect the length of transmission line to the *unknown* terminal—it should be longer than the expected length.
4. For quarter-wavelength lines, shorten the line until the null is very close to the desired frequency. For half-wavelength lines, do the same thing, except that the line must be shorted at the far end for each trial length.

The velocity factor of a transmission line (usually designated by the letter  $V$  in equations) is a decimal fraction that tells us how fast the radio wave propagates along the line relative to the speed of light in free space. For example, foam dielectric coaxial cable is said to have a velocity factor of  $V = 0.80$ . This number means that the signals in the line travel at a speed 0.80 (or 80 percent) of the speed of light.

Because all radio wavelength formulas are based on the velocity of light, you need the  $V$  value to calculate the physical length needed to equal any given electrical length. For example, a half-wavelength piece of coax has a physical length of  $(492 \times V)/F_{\text{MHz}}$  feet. Unfortunately, the real value of  $V$  is often a bit different from the published value. You can use the noise bridge to find the actual value of  $V$  for any sample of coaxial cable as follows:

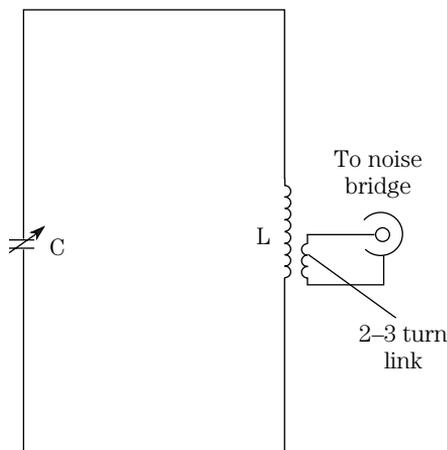
1. Select a convenient length of the coax more than 12 ft in length and install a PL-259 RF connector (or other connector compatible with your instrument) on one end, and short-circuit the other end.
2. Accurately measure the physical length of the coax in feet; convert the “remainder” inches to a decimal fraction of a foot by dividing by 12 (e.g., 32 ft 8 in = 32.67 ft because 8 in/12 in = 0.67). Alternatively, cut off the cable to the nearest foot and reconnect the short circuit.
3. Set the bridge *resistance* and *reactance* controls to zero.
4. Adjust the monitor receiver for deepest null. Use the null frequency to find the velocity factor  $V = FL/492$ , where  $V$  is the velocity factor (a decimal fraction);  $F$  is the frequency in megahertz; and  $L$  is the cable length in feet.

**Tuned circuit measurements** An inductor/capacitor ( $LC$ ) tuned “tank” circuit is the circuit equivalent of a resonant antenna, so there is some similarity between the two measurements. You can measure resonant frequency with the noise bridge to within  $\pm 20$  percent (or better if care is taken). This accuracy might seem poor, but it is better than you can usually get with low-cost signal generators, dip meters, absorption wavemeters, and the like.

A *series tuned circuit* exhibits a low impedance at the resonant frequency and a high impedance at all other frequencies. Start the measurement by connecting the series tuned circuit under test across the *unknown* terminals of the dip meter. Set the *resistance* control to a low resistance value, close to zero ohms. Set the *reactance* control at midscale (zero mark). Next, tune the receiver to the expected null frequency, and then tune for the null. Make sure that the null is at its deepest point by rocking the R and X controls for best null. At this point, the receiver frequency is the resonant frequency of the tank circuit.

A *parallel resonant circuit* exhibits a high impedance at resonance and a low impedance at all other frequencies. The measurement is made in exactly the same manner as for the series resonant circuits, except that the connection is different. Figure 27-8 shows a two-turn link coupling that is needed to inject the noise signal into the parallel resonant tank circuit. If the inductor is the toroidal type, then the link must go through the hole in the doughnut-shaped core and then connect to the *unknown* terminals of the bridge. After this, proceed exactly as you would for the series tuned tank measurement.

**Capacitance and inductance measurements** The bridge requires a 100-pF silver mica test capacitor and a 4.7- $\mu\text{H}$  test inductor, which are used to measure inductance and capacitance, respectively. The idea is to use the test components to form a series-tuned resonant circuit with an unknown component. If you find a resonant frequency, then you can calculate the unknown value. In both cases, the series tuned circuit is connected across the *unknown* terminals of the dip meter, and the series tuned procedure is followed.



**27-8** Noise bridge connection to  $LC$  tank circuits.

To measure inductance, connect the 100-pF capacitor in series with the unknown coil across the *unknown* terminals of the dip meter. When the null frequency is found, find the inductance from

$$L = \frac{253}{F^2} \quad [27.9]$$

where

$L$  is the inductance, in microhenries

$F$  is the frequency, in megahertz

Connect the test inductor across the *unknown* terminals in series with the unknown capacitance. Set the *resistance* control to zero, tune the receiver to 2 MHz, and readjust the *reactance* control for null. Without readjusting the noise bridge control, connect the test inductor in series with the unknown capacitance and retune the receiver for a null. Capacitance can now be calculated from:

$$C = \frac{5389}{F^2} \quad [27.10]$$

where:

$C$  is in picofarads

$F$  is in megahertz

## Dip oscillators

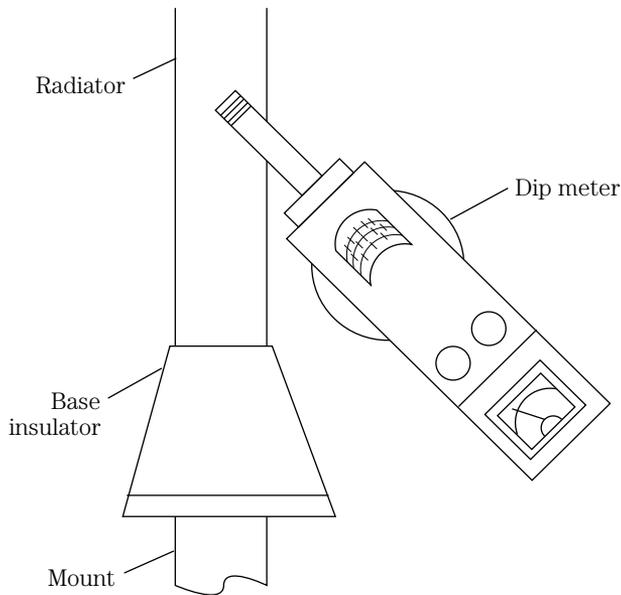
One of the most common instruments for determining the resonant frequency of an antenna is the so-called dip oscillator or dip meter. Originally called the *grid dip meter*, the basis for this instrument is the fact that its output energy can be absorbed by a nearby resonant circuit (or antenna, which is an electrically resonant LC tank circuit). When the inductor of the dip oscillator (see Fig. 27-9) is brought into



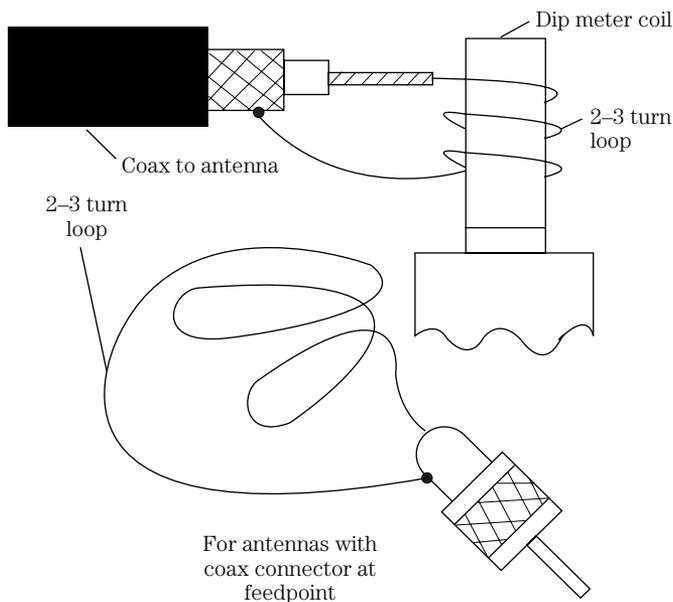
27-9 Dip meter.

close proximity to a resonant tank circuit, and the oscillator is operating on the resonant frequency, then a small amount of energy is transferred. This energy loss shows up on the meter pointer as a slight “dipping” action. The dip is extremely sharp, and is easily missed if the meter frequency dial is tuned too rapidly.

Antennas are resonant circuits, and can be treated in a manner similar to *LC* tank circuits. Figure 27-10A shows one way to couple the dip oscillator to a vertical antenna radiator. The inductor of the dipper is brought into close proximity to the base of the radiator. Figure 27-10B shows the means for coupling the dip oscillator



**27-10A** Direct coupling the dip meter to antenna.



**27-10B** Coupling the dip meter to coaxial cable.

to systems where the radiator is not easily accessed (as when the antenna is still erected). We connect a small two- or three-turn loop to the transmitter end of the transmission line, and then bring the inductor of the dipper close to it. A better way is to connect the loop directly to the antenna feedpoint.

There are two problems with dip meters that must be recognized in order to best use the instrument. First, the dip is very sharp. It is easy to tune past the dip and not even see it. To make matters worse, it is normal for the meter reading to drop off gradually from one end of the tuning range to the other. But if you tune very slowly, then you will notice a very sharp dip when the resonant point is reached.

The second problem is the dial calibration. The dial gradations of inexpensive dip meters are too close together and are often erroneous. It is better to monitor the output of the dip oscillator on a receiver and depend upon the calibration of the receiver for data.

## Selecting and using RF wattmeters and antenna VSWR meters

A key instrument required in checking the performance of, or troubleshooting, radio transmitters is the RF power meter (or “wattmeter”). These instruments measure the output power of the transmitter, and display the result in watts, or some related unit. Closely related to RF wattmeters is the antenna VSWR meter. This instrument also examines the output of the transmitter and gives a relative indication of output power. It can be calibrated to display the dimensionless units of *voltage standing wave ratio*. Many modern instruments, a couple of which will be covered as examples in this chapter, combine both RF power and VSWR measurement capabilities.

### Measuring RF power

Measuring RF power has traditionally been notoriously difficult, except perhaps in the singular case of continuous-wave (CW) sources that produce nice, well-behaved sine waves. Even in that limited case, however, some measurement methods are distinctly better than others.

The peak voltage of a waveform is 100 V (i.e., peak-to-peak 200 V). Given that the CW waveform is sinusoidal, we know that the root mean square (RMS) voltage is 0.707 V. The output power is related to the RMS voltage across the load by

$$P = \frac{(V_{\text{rms}})^2}{Z_o} \quad [27.11]$$

where:

- $P$  is the power, in watts
- $V_{\text{rms}}$  is the RMS potential, in volts
- $Z_o$  is the load impedance, in ohms

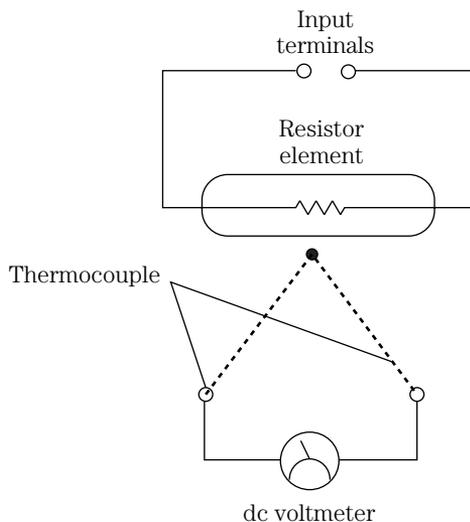
If we assume a load impedance of 50  $\Omega$ , then we can state that the power in our hypothetical illustration waveform is 100 W.

We can measure power on unmodulated sinusoidal waveforms by measuring either the RMS or peak values of either voltage or current, assuming that a constant-value resistance load is present. But the problem becomes more complex on modulated signals. The various power readings on a Bird model 4311 peak power meter, the peak (PEP) and average powers, vary markedly with modulation type.

One of the earliest forms of practical RF power measurement was the thermocouple RF ammeter (see Fig. 27-11). This instrument works by dissipating a small amount of power in a small resistance inside the meter, and then measuring the heat generated with a thermocouple. A dc current meter monitors the output of the thermocouple device, and indicates the level of current flowing in the heating element. Because it works on the basis of the power dissipated heating a resistance, *a thermocouple RF ammeter is inherently an RMS-reading device*. Because of this feature it is very useful for making average power measurements. If we know the RMS current and the resistive component of the load impedance, and if the reactive component is zero or very low, then we can determine RF power from the familiar expression:

$$P = I^2 \times R_i \quad [27.12]$$

There is, however, a significant problem that keeps thermocouple RF ammeters from being universally used in RF power measurement: those instruments are highly frequency-dependent. Even at low frequencies, it is recommended that the meters be mounted on insulating material with at least  $\frac{3}{8}$ -in spacing between the meter and its metal cabinet. Even with that precaution, however, there is a strong frequency dependence that renders the meter less useful at higher frequencies. Some meters are advertised to operate into the low-VHF region, but a note of caution is necessary. That recommendation requires a copy of the calibrated frequency response curve for that specific meter, so that a correction factor can be added (or subtracted) from



**27-11** Circuit for thermocouple RF ammeter.

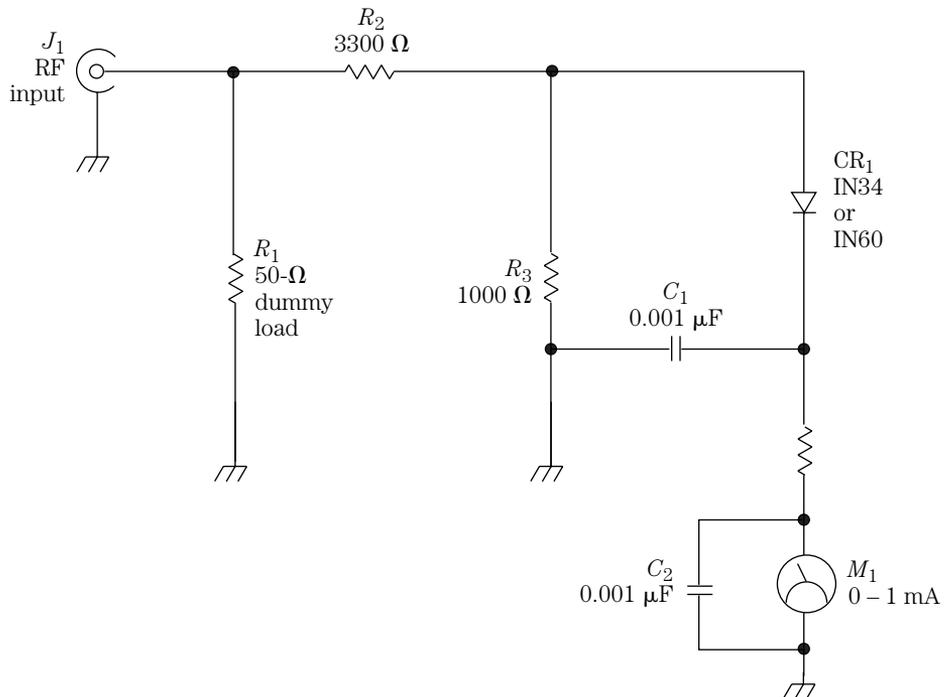
the reading. At 10 MHz and higher, the readings of the thermocouple RF ammeter must be taken with a certain amount of skepticism unless the original calibration chart is available.

We can also measure RF power by measuring the voltage across the load resistance (see Fig. 27-12). In the circuit of Fig. 27-12, the RF voltage appearing across the load is scaled downward to a level compatible with the voltmeter by the resistor voltage divider ( $R_2/R_3$ ). The output of this divider is rectified by CR<sub>1</sub> and filtered to dc by the action of capacitor C<sub>2</sub>.

The method of measuring the voltage in a simple diode voltmeter is valid only if the RF signal is unmodulated and has a sinusoidal waveshape. While these criteria are met in many transmitters, they are not universal. If the voltmeter circuit is peak reading, as in Fig. 27-12, then the peak power is

$$P = \frac{V_o^2}{R_1} \quad [27.13]$$

The average power is then found by multiplying the peak power by 0.707. Some meter circuits include voltage dividers that precede the meter and thereby convert the reading to RMS, and thus convert the power to average power. Again, it must be stressed that terms like RMS, average, and peak have meaning only when the input

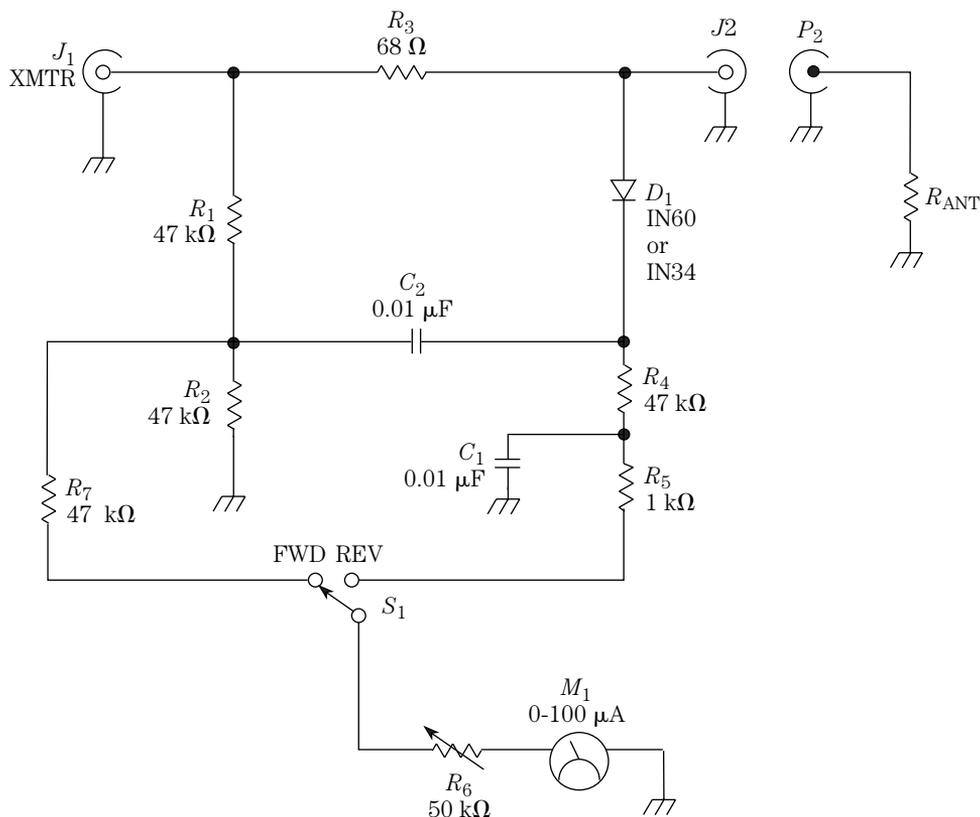


27-12 RF voltmeter "wattmeter."

RF signal is both unmodulated and sinusoidal. Otherwise, the readings are meaningless unless calibrated against some other source.

It is also possible to use various bridge methods for measurement of RF power. Figure 27-13 shows a bridge set up to measure both forward and reverse power. This circuit was once popular for VSWR meters. There are four elements in this quasi-Wheatstone bridge circuit:  $R_1$ ,  $R_2$ ,  $R_3$ , and the antenna impedance (connected to the bridge at  $J_2$ ). If  $R_{\text{ant}}$  is the antenna resistance, then we know that the bridge is in balance (i.e., the null condition) when the ratios  $R_1/R_2$  and  $R_3/R_{\text{ant}}$  are equal. In an ideal situation, resistor  $R_3$  will have a resistance equal to  $R_{\text{ant}}$ , but that might overly limit the usefulness of the bridge. In some cases, therefore, the bridge will use a compromise value such as  $67\ \Omega$  for  $R_3$ . Such a resistor will be usable on both 50- and  $75\text{-}\Omega$  antenna systems with only small errors. Typically, these meters are designed to read relative power level, rather than the actual power.

An advantage of this type of meter is that we can get an accurate measurement of VSWR by proper calibration. With the switch in the *forward* position, and RF power applied to  $J_1$  (XMTR), potentiometer  $R_6$  is adjusted to produce a full-scale deflection on meter  $M_1$ . When the switch is then set to the *reverse* position, the meter will read reverse power relative to the VSWR. An appropriate VSWR scale is provided.



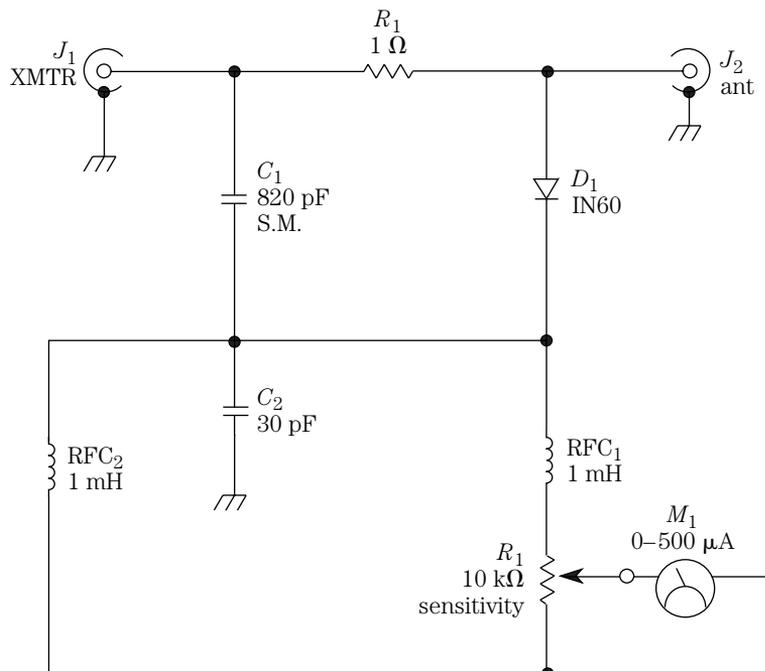
27-13 A bridge arrangement to measure both forward and reverse power.

A significant problem with the bridge of Fig. 27-13 is that it cannot be left in the circuit while transmitting because it dissipates a considerable amount of RF power in the internal resistances. These meters, during the time when they were popular, were provided with switches that bypassed the bridge when transmitting. The bridge was only in the circuit when making a measurement.

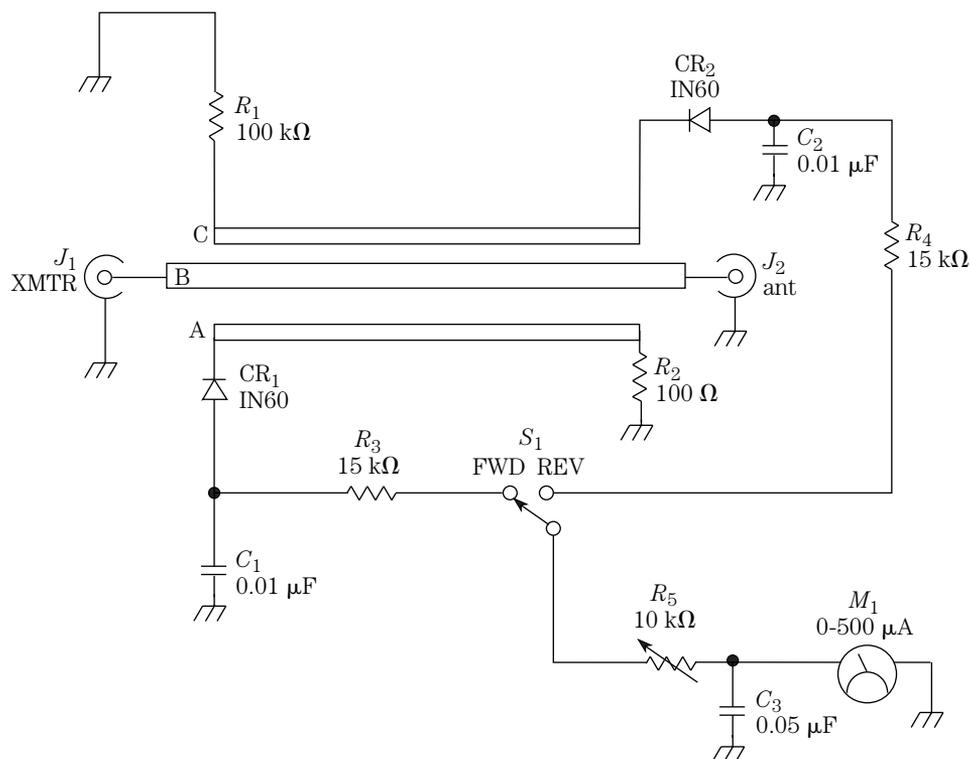
An improved bridge circuit is the capacitor/resistor bridge in Fig. 27-14; this circuit is called the micromatch bridge. Immediately, we see that the micromatch is improved over the conventional bridge because it uses only  $1\ \Omega$  in series with the line ( $R_i$ ). This resistor dissipates considerably less power than the resistance used in the previous example. Because of this low-value resistance, we can leave the micromatch in the line while transmitting. Recall that the ratios of the bridge arms must be equal for the null condition to occur. In this case, the capacitive reactance ratio of  $C_1/C_2$  must match the resistance ratio  $R_1/R_{\text{ant}}$ . For a  $50\text{-}\Omega$  antenna, the ratio is  $1/50$ , and for  $75\text{-}\Omega$  antennas it is  $1/75$  (or, for the compromise situation,  $1/68$ ). The small-value trimmer capacitor ( $C_2$ ) must be adjusted for a reactance ratio with  $C_1$  of  $1/50$ ,  $1/75$ , or  $1/68$ , depending upon how the bridge is set up.

The sensitivity control can be used to calibrate the meter. In one version of the micromatch, there are three power ranges (10, 100, and 1000 W). Each range has its own sensitivity control, and these are switched in and out of the circuit as needed.

The monomatch bridge circuit in Fig. 27-15 is the instrument of choice for HF and low-VHF applications. In the monomatch design, the transmission line is segment



27-14 Micromatch wattmeter.



**27-15** Monomatch wattmeter.

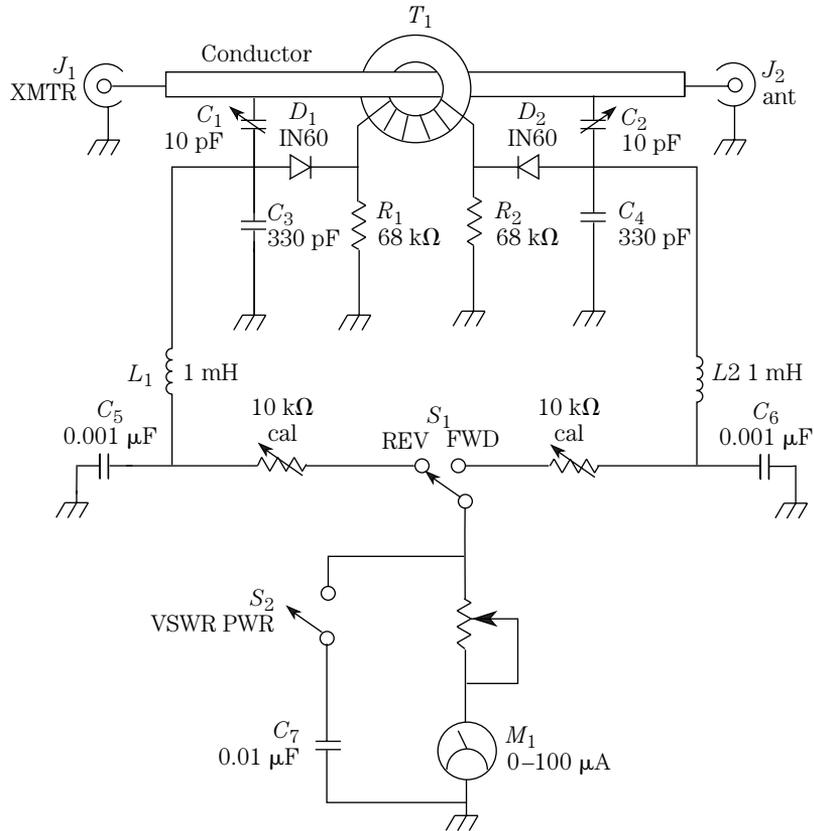
B, while RF sampling elements are formed by segments A and C. Although the original designs were based on a coaxial-cable sensor, later versions used either printed circuit foil transmission line segments or parallel brass rods for A, B, and C.

The sensor unit is basically a directional coupler with a detector element for both forward and reverse directions. For best accuracy, diodes  $CR_1$  and  $CR_2$  should be matched, as should  $R_1$  and  $R_2$ . The resistance of  $R_1$  and  $R_2$  should match the transmission line surge impedance, although in many instruments a 68- $\Omega$  compromise resistance is used.

The particular circuit shown in Fig. 27-15 uses a single dc meter movement to monitor the output power. Many modern designs use two meters (one each for forward and reverse power).

One of the latest designs in VSWR meter sensors is the current transformer assembly shown in Fig. 27-16. In this instrument, a single-turn ferrite toroid transformer is used as the directional sensor. The transmission line passing through the hole in the toroid "doughnut" forms the primary winding of a broadband RF transformer. The secondary, which consists of 10 to 40 turns of small enamel wire, is connected to a measurement bridge circuit ( $C_1 + C_2 + \text{load}$ ) with a rectified dc output.

Figures 27-17 and 27-18 show instruments based on the current transformer technique. Shown in Fig. 27-17 is the Heath model HM-102 high-frequency VSWR/power meter. The sensor is a variant on the current transformer method. This



27-16 Current transformer wattmeter.

27-17 Amateur radio RF wattmeter.



instrument measures both forward and reflected power, and it can be calibrated to measure VSWR.

The Bird model 43 Thurline RF wattmeter shown in Fig. 27-18 has for years been one of the industry standards in communications service work. Although it is slightly more expensive than lesser instruments, it is also versatile, and it is accurate and rugged. The Thurline meter can be inserted into the transmission line of an

**27-18** Bird model 43 RF wattmeter.

antenna system with so little loss that it may be left permanently in the line during normal operations. The Model 43 Thruline is popular with land mobile and marine radio technicians.

The heart of the Thruline meter is the directional coupler transmission line assembly shown in Fig. 27-19A; it is connected in series with the antenna, or dummy load, transmission line. The plug-in directional element can be rotated 180° to measure both forward and reverse power levels. A sampling loop and diode detector are contained within each plug-in element. The main RF barrel is actually a special coaxial line segment with a 50-Ω characteristic impedance. The Thruline sensor works due to the mutual inductance between the sample loop and center conductor of the coaxial element. Figure 27-19B shows an equivalent circuit. The output voltage from the sampler ( $e$ ) is the sum of two voltages,  $e_r$  and  $e_m$ . Voltage  $e_r$  is created by the voltage divider action of  $R$  and  $C$  on transmission line voltage  $E$ . If  $R$  is much less than  $X_C$ , then we may write the expression for  $e_r$  as

$$e_r = \frac{RE}{X_C} = RE(j\omega C) \quad [27.14]$$

Voltage  $e_m$ , on the other hand, is due to mutual inductance, and is expressed by

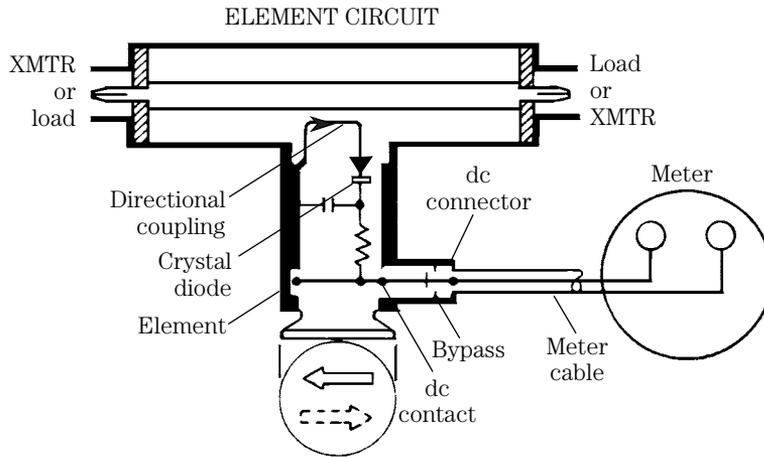
$$e_m = I(j\omega) \pm M \quad [27.15]$$

We now have the expression for both factors that contribute to the total voltage  $e$ . We know that:

$$e = e_r + e_m \quad [27.16]$$

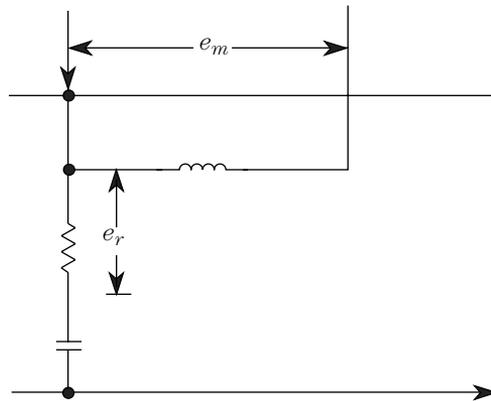
so, by substitution,

$$e = j\omega M \left[ \left( \frac{E}{Z_o} \right) \pm I \right] \quad [27.17]$$



27-19A Thru-line sensor circuit.

27-19B Circuit of pick-up element in a Model 43 wattmeter.



At any given point in a transmission line,  $E$  is the sum of the forward ( $E_f$ ) and reflected ( $E_r$ ) voltages, and the line current is equal to

$$I = \frac{E_f}{Z_o} - \frac{E_r}{Z_o} \quad [27.18]$$

where  $Z_o$  is the transmission line impedance.

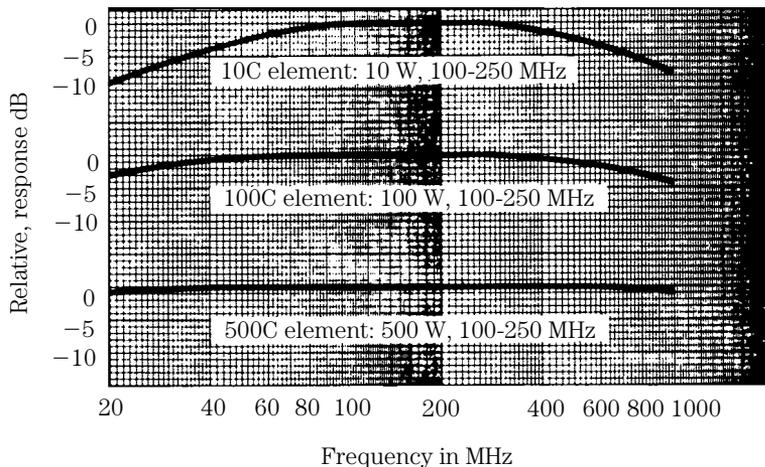
We may specify  $e$  in the forms

$$e = \frac{j \omega M(2E_f)}{Z_o} \quad [27.19]$$

and

$$e = \frac{j \omega M(2E_r)}{Z_o} \quad [27.20]$$

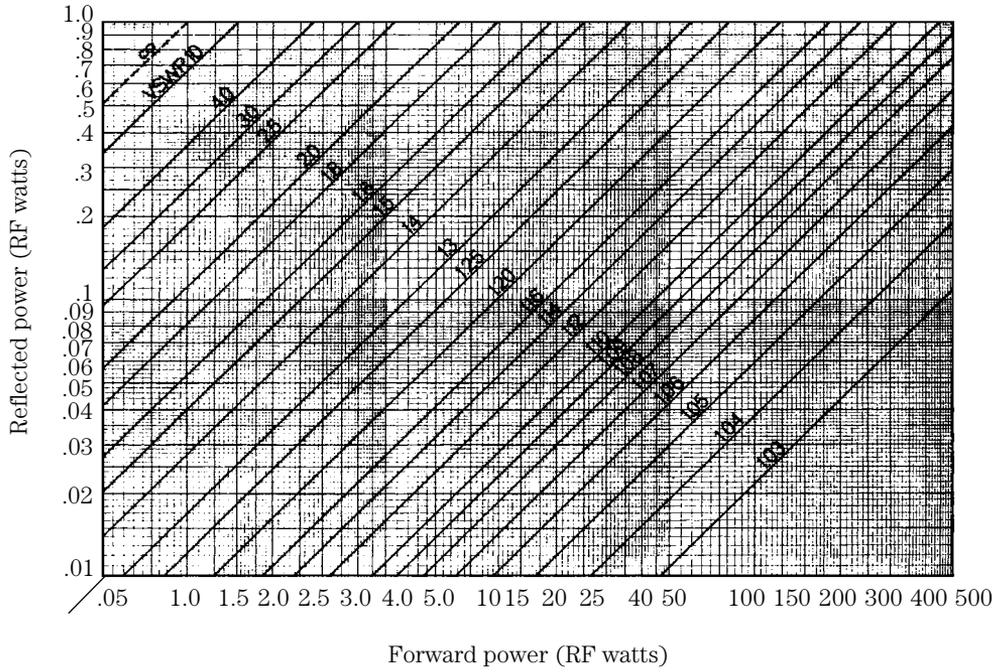
The output voltage  $e$  of the coupler, then, is proportional to the mutual inductance and frequency (by virtue of  $j\omega M$ ). But the manufacturer terminates  $R$  in a capacitive reactance, so the frequency dependence is lessened (see Fig. 27-19C). Each element is custom-calibrated, therefore, for a specific frequency and power range. Beyond the specified range for any given element, however, performance is not guaranteed. There are a large number of elements available that cover most commercial applications. The Thru-line meter is not a VSWR meter, but rather a power meter. VSWR can be determined from the formula, or by using the nomographs in Fig. 27-20. Some of the Thru-line series intended for very high power (Fig. 27-19D) applications use an in-line coaxial cable coupler (for broadcast-style hardline) and a remote indicator.



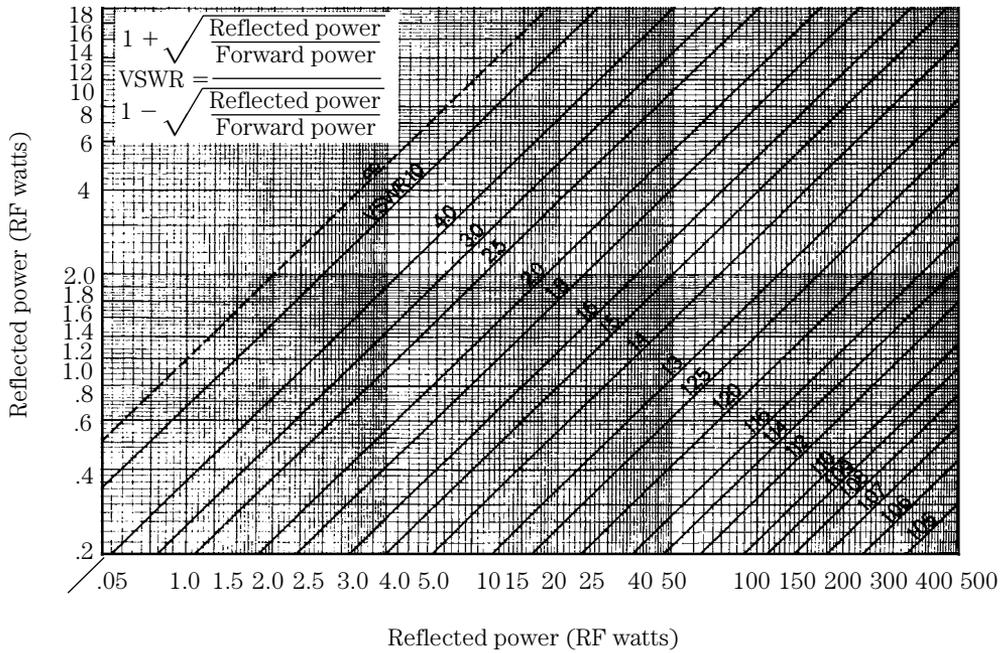
**27-19C** Power-frequency calibration.



**27-19D** High-power RF wattmeters.



**A**



**B**

27-20 Forward versus reflected power graphs.

## Dummy loads

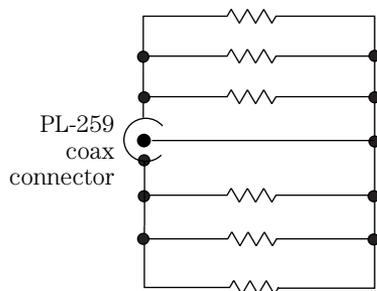
A *dummy load* is a substitute antenna for making measurements and tests. In fact, British radio engineers often refer to dummy loads as “artificial aerials.” There are several uses for these devices. Radio operators should use dummy loads to tune up on crowded channels and (only then) transfer to the live antenna.

The other use is in troubleshooting antenna systems. Suppose we have a system in which the VSWR is high enough to affect the operation of the transmitter. You can disconnect each successive element and connect the dummy load to its output. If the VSWR goes down to the normal range, then the difficulty is downstream (i.e., toward the antenna). You will eventually find the bad element (which is usually the antenna itself).

Figure 27-21 shows the most elementary form of dummy load, which consists of one or more resistors connected in parallel, series, or series-parallel so that the total resistance is equal to the desired load impedance. The power dissipation is the sum of the individual power dissipations. *It is essential that noninductive resistors be used for this application.* For this reason, carbon-composition or metal-film resistors are used. For very low frequency work, it is permissible to use special counter wound wire low-inductance resistors. These resistors, however, cannot be used over a few hundred kilohertz.

Several commercial dummy loads are shown in Figs. 27-22 through 27-24. The load in Fig. 27-22A is a 5-W model, and is typically used in Citizens Band servicing. The resistor is mounted directly on a PL-259 coaxial connector. These loads typically work to about 300 MHz, although many are not really useful over about 150 MHz. A higher-power version of the same type is shown in Fig. 27-22B. This device works to the low VHF region, and dissipates up to 50 W. I have used this dummy load for servicing high-VHF land mobile rigs, VHF-FM marine rigs, and low-VHF land mobile rigs. Very high power loads are shown in Fig. 27-23. These devices are *Bird Electronics* “coaxial resistors,” and operate to power levels up to 10, 30, and 40 kW. These high-power loads are cooled by flowing water through the body of the resistor and then exhausting the heat in an air-cooled radiator.

Our final dummy load resistor is shown in Fig. 27-24. The actual resistor is shown in Fig. 27-24A, and a schematic view is shown in Fig. 27-24B. The long, high-power noninductive resistor element is rated at 50  $\Omega$ , and can dissipate 1000 W for



**27-21** A very basic dummy load.



27-22A CB-style 5-W dummy load.

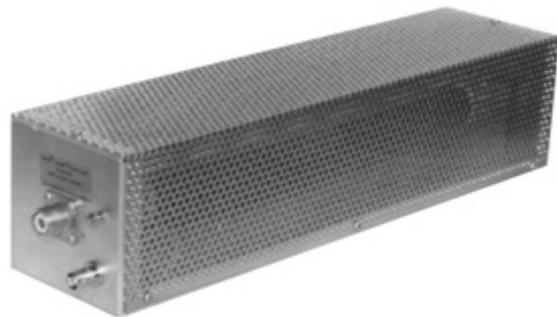


27-22B 15-W dummy load.

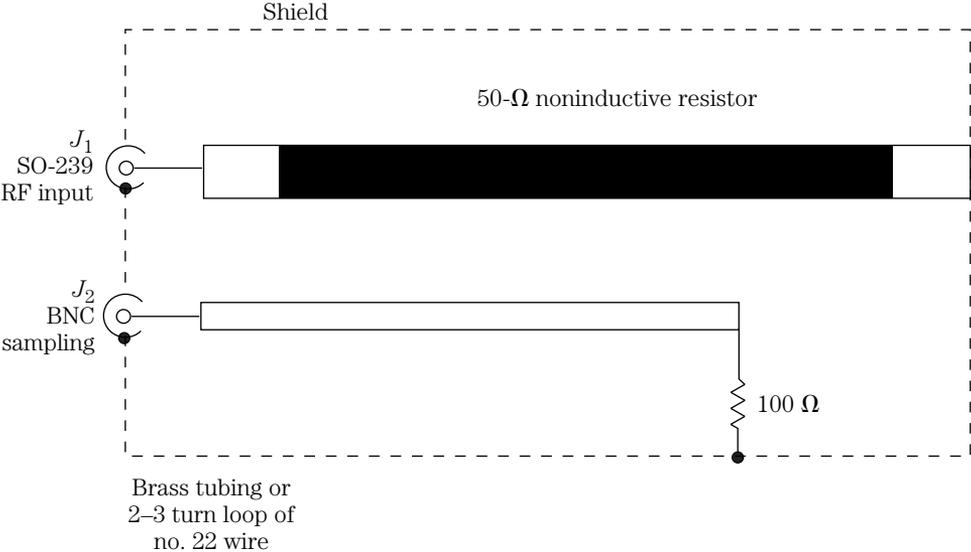
27-23 Termaline RF wattmeter.



27-24A Drake DL-1000 dummy load.



several minutes. If longer times, or higher powers, are anticipated, then forced air cooling is applied by adding a blower to one end of the cage. The device is modified by adding the BNC sampling jack. This jack is connected internally to either a two-turn loop, made of no.22 insulated hookup wire, or brass rods that are positioned alongside the resistor element. It will therefore pick up a sample of the signal so that it can be viewed on an oscilloscope, or used for other instrumentation purposes.



**27-24B** HF dummy load with pickup loop.

## 28 CHAPTER

# General antenna mechanical construction techniques

ANTENNA INSTALLATION CHORES VARY IN COMPLEXITY FROM THE SIMPLE, WHICH A teenager, with no assistance, can execute properly and safely, all the way up to large-scale projects that are best left to professional antenna riggers. If your only antenna installation chore is a dipole hanging between the peak of your house roof and a nearby tree, then little need be said besides “Be careful.”

The purpose of this chapter is to cover both wire and intermediate installation jobs that can be done by amateurs if instructed properly. Keep in mind when planning an installation, however, that the information given herein is merely informal guidance. Always have your plans reviewed by a professional mechanical or structural engineer licensed to practice in your state. In many locales, there are good insurance reasons for this step. Also, learn and abide by local mechanical, electrical, and building codes. It is simply not worth the pain, and risk, to skirt around local laws.

## Antenna erection safety

Before dealing with the radio and performance issues, let's deal first with a few antenna safety matters. You do not want to be hurt either during installation or during the next wind storm. Two problems present themselves: *reliable mechanical installation* and *electrical safety*.

Radio antennas are inherently dangerous to erect if certain precautions are not followed. It is not possible to foresee all the situations you might face in erecting an antenna. I would like to give you all possible warnings, but this is not even possible. You are on your own and must take responsibility for installing your own antenna. I can, however, give you some general safety guidelines. Knowledge of what you face and some hard-nosed, sound judgment, modulated by a little bit of common sense are the best tools on any antenna job.

One rule that is an absolute is that no antenna should *ever* be erected where the antenna, the feedline, or any part thereof crosses over or under a power line or pole transformer—*ever*! This is a “no kiddie”—don’t do it. Every year we read sad news in magazines or newspapers about a colleague being electrocuted while installing or working on an antenna. In all these tragic cases, the antenna somehow came into contact with an ac power line. Keep in mind one dictum and make it an absolute: There is never a time or situation when it is safe to let an antenna contact the electrical power lines!

This advice includes dipoles and other long wire antennas “thrown over” supposedly insulated lines, as well as antennas built from aluminum tubing. The excuse that the power lines are insulated is hogwash. Old insulation crumbles on contact with even a thin wire antenna. *Don’t do it!* The operant word is *never*!

Power lines look insulated, but there are often small breaks or weakened spots (especially more than a couple days after installation) that can bring the antenna into contact—lethal contact—with a “hot” power line. Every year or so we hear about an SWL, scanner/monitor buff, or ham operator being *killed* by tossing an antenna wire over a power line. Avoid making yourself into a high-powered resistor—it might blow your “life fuse.”

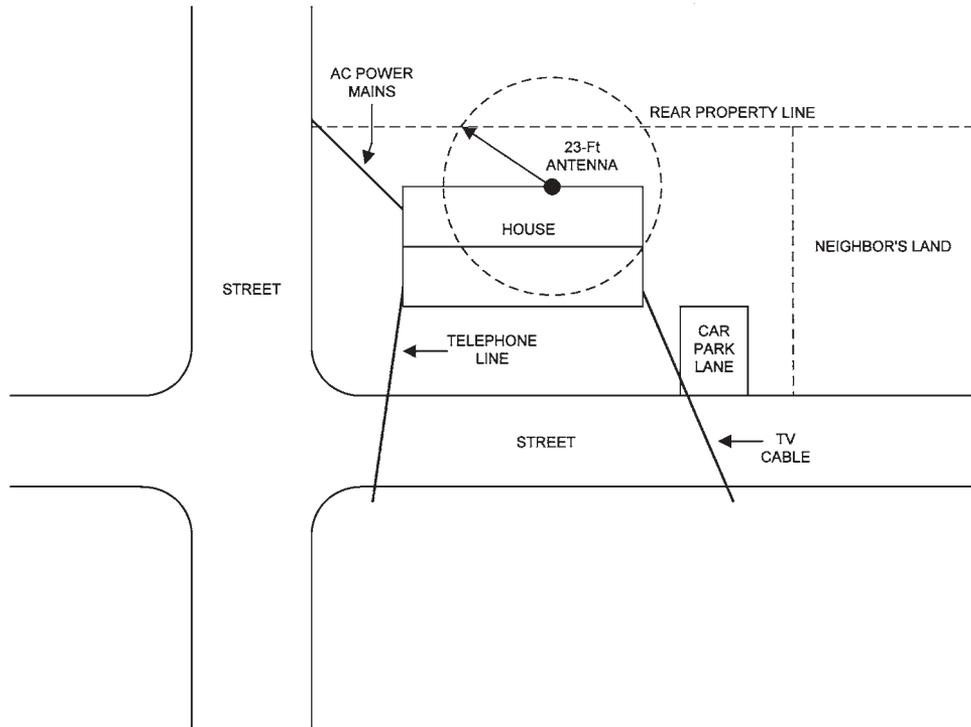
And the same rule applies to situations where the antenna can fall onto a power line if it fails mechanically. You have to examine the situation to see if there is any possible way for that antenna or its support structure to fall onto a power line if it breaks in any way whatsoever. Figure 28-1 is a diagram of my lot in Virginia. A 23-ft mast is erected at the back of the house, and its 23-ft “fall radius” is plotted. It should not intersect either the power lines or the cable TV line, even when it fails and is wind-whipped.

Note that the antenna could fall over the property line onto the next lot to the rear. After consultation with the other landowner and the county mechanical inspector, it was determined that my rear fence is sufficient protection for the other homeowner. Because it is technically on my land, the fence damage is my responsibility. However, should the other homeowner change his mind or a new homeowner buys the place, then my antenna may have to be moved.

Another caution is that you must be physically fit to do the work. While the on-the-ground portion of the work is usually not too strenuous, any climbing at all, even on ladders, can be taxing. Antenna materials are deceptively lightweight on the ground, but when you get up on even a small ladder, they are remarkably difficult to handle. Attempting to manhandle a 22-ft vertical once wiped my back out to the extent that I needed 800-mg of ibuprofen three times a day for about a week—and I consider myself lucky that the pain hit me after I had dismounted the ladder. Besides, if you could see me, you would wonder why a man my size was on any ladder in the first place!

Before using a ladder, learn how to use a ladder. A lot of homeowners, whether putting up antennas or painting the upstairs windows, fall off ladders that were being used incorrectly. There is a right way and a wrong way to use each of the different types of ladder; learn to use them correctly.

If the wind blows even a few kilometers per hour, the danger is magnified considerably. I recall a friend, who is a large, strong “bear of a man,” attempting to install



**28-1** Antenna installation on suburban lot.

a 26-element television “all channel” rotatable antenna on the roof of his two-story beach house. He wanted to be able to pick up Richmond, VA; Norfolk, VA; Washington, DC; and Baltimore, MD, on television, plus scanner bands, from his beach house on the Chesapeake Bay. The antenna was handled easily with one hand on the ground and with no wind blowing. But up on the roof, it was a different story! “Big Al” was on the peak of the roof when an off-shore gust of wind came up suddenly and caught the antenna. It acted like a hang glider and pulled Al off the roof, plunging him down two stories to the patio below. He fractured his pelvis and busted a leg. Two months of orthopedic casts and a year of physical therapy followed—not to mention lost wages. Expensive TV antenna, I reckon. *Be careful!*

One good rule is to always work under the “buddy system.” Ask as many friends as are needed to do the job safely, and always have at least one assistant, even when you think you can do it alone.

Always use quality materials and good work practices. Antennas, being potentially dangerous, always should have the best of both materials and workmanship in order to keep quality high. It is not just the electrical or radio reception workings that are important but also the ability to stay up in the air and be safe.

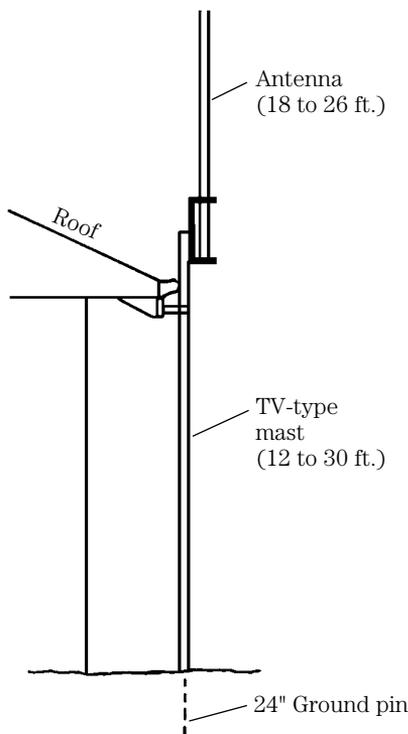
When planning an antenna job, keep in mind that pedestrian traffic in your yard possibly could affect the antenna system. Wires are difficult to see, and if an antenna

wire is low enough to intersect someone's body, then it is possible to cause very serious injuries to passersby. In the United States, even when the person is a trespasser (or even a burglar!), the courts may hold a homeowner liable for injuries caused by an inappropriately designed and installed antenna. Take care for safety not only of yourself but also of others.

Consider a typical scenario involving a four- or five-band trap vertical antenna (Fig. 28-2). It will be 18 to 26 ft tall, judging from the advertisements in magazines, and will be mounted on a roof or mast 12 to 30 ft off the ground. Consider the case of a 25-ft-tall trap vertical installed atop a 15-ft-high telescoping TV antenna mast. The total height above ground will be the sum of the two heights: 15 ft + 25 ft = 40 ft. The tip of the vertical is 40-ft above ground level. You must select a location at which a 40-ft aluminum pole can fit—and fall. This requirement limits the selection of locations for the antenna. If this is done, then a wind storm cannot result in a shorted or downed power line if the antenna falls over.

When installing a vertical antenna, especially one that is not ground-mounted, make sure that you have help. It takes at least two people to safely install a standard HF vertical antenna, and more may be needed for especially large models. If you are alone, then go find some friends to lend an arm or two. Wrenched backs, smashed antennas, crunched house parts, and other calamities simply do not happen as often on a well-organized work party that has sufficient hands to do the job safely.

The second issue pertaining to installing antennas is old-fashioned mechanical integrity. Two problems are often seen. First, you must comply with local building



28-2 Mast-mounted vertical antenna.

codes, regulations, and inspections. The local government has a reasonable interest and absolute right to impose reasonable engineering standards on the electrical and mechanical installation of radio antennas. The second issue is that it is in your own best interest to make the installation as good as possible. Consider the local regulations as the minimum acceptable standard, not the maximum.

Both these mechanical integrity issues become extremely important if a problem develops. For example, suppose that a wind or snow storm wrecks the antenna, plus part of your house. Your insurance company may not pay off if your local government requires an inspection and you failed to get one. Make sure that the mechanical and/or electrical inspector (as required by the law) leaves a certificate or receipt proving that the final inspection was done. It could come in handy when disputing the insurance company over damage.

A quality installation starts with the selection of good hardware for the installation. Any radio-TV parts distributor who sells TV antenna hardware will have what you need in most cases. I used Radio Shack standoff brackets, ground pin, and a 19-ft telescoping mast in one installation. Wherever you buy, select the best-quality, strongest materials that you can find. Opt for steel masts and brackets over aluminum, no matter what the salesperson behind the counter tells you. Keep in mind that although salespeople can be knowledgeable and even helpful, you, not they, are responsible for the integrity of the installation.

**Local government** One necessary reminder is that your local government might have some interesting ideas—legal requirements, actually—concerning your antenna installation. The electrical, mechanical, and zoning codes must be observed. There is a great deal of similarity between local codes because most of them are adaptations (even adoptions) from certain national standards. But there are enough differences that one needs to consult local authorities. Indeed, you may need a license or building permit to install an antenna in the first place.

One problem that SWLs and scanner/monitor buffs face in the United States is that their antennas are not protected by the Federal Communications Commission, as are ham antennas. Local governments have limited rights to regulate ham antennas—only “reasonable” mechanical and electrical standards can be imposed. It may be illegal for you to install any type of antenna.

Save all paperwork regarding your building permit, including the inspection certificates, decals, letters or papers, and the original drawings (with the local building inspector’s stamps). If a casualty occurs, then your insurance company may elect not to pay off if you have violated an electrical, mechanical, building, or zoning code. This advice may be overlooked by an enthusiastic antenna builder, but it could prove to be a costly oversight if something bad happens.

In some jurisdictions, there might be another limitation on antenna location. Some local governments have a requirement that the antenna be able to fall over and land entirely on your own property. About 30 years ago, a Maryland county required that antennas be installed double their own length plus 50 ft from the nearest property line—until a local ham radio club sued on grounds that about 99.99 percent of the rooftop TV antennas in the county were noncompliant. In one humorous scene, a group of amateur radio operators went down to the county offices and asked for about 50,000 complaint forms. They filled out several thousand against homeowners

with TV antennas before they were stopped by the county officer. Before installing any antenna, check with the local building codes!

## Wire antenna construction

Volumes have been written on wire antennas. Indeed, many of my antenna books spend a lot of space on the topic. What seems to be missing in many books is details on the mechanical aspects of wire antenna construction and installation. Although some books cover the mechanical details, much is often missing. Let's take a look at some of the basics.

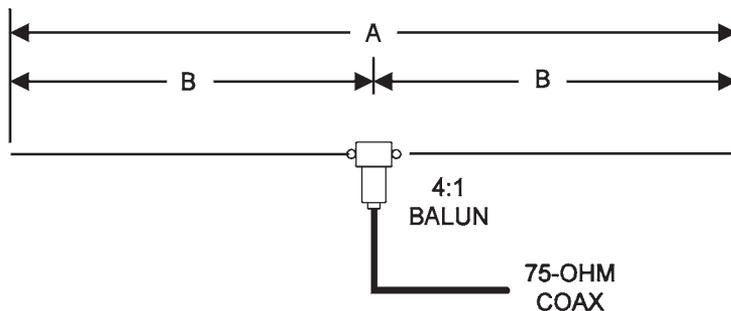
### The basic wire antenna

Wire antennas come in a bewildering variety of forms, but for our purposes, the "standard model" is our old friend the half-wavelength horizontal dipole (Fig. 28-3). The overall length ( $A$ ) in meters is found from  $144/F_{\text{MHz}}$ . Each element ( $B$ ) is one-half the overall length, or  $72/F_{\text{MHz}}$ . The antenna's nominal feedpoint impedance at resonance is  $73 \Omega$ , so it is a good match to standard  $75\text{-}\Omega$  coaxial cable. It is common practice to use a 1:1 balun transformer at the feedpoint in order to balance currents and keep the pattern according to textbook predictions.

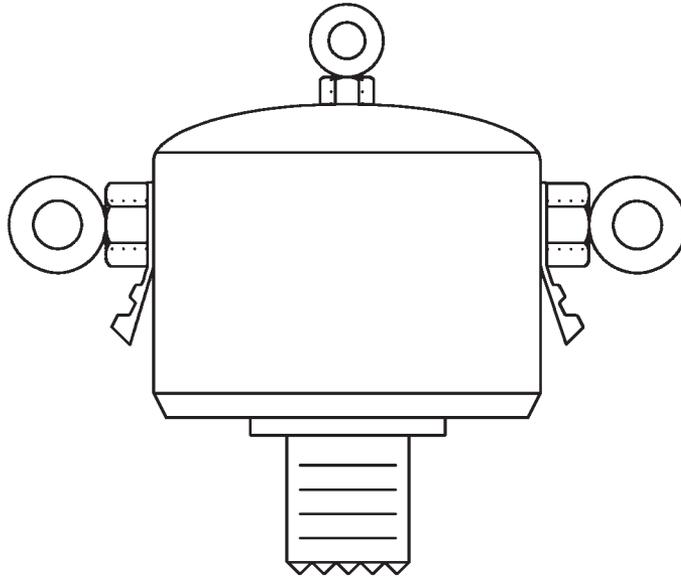
### Center insulators

It was once common (but poor) practice to strip a few centimeters of coaxial cable, part the braid and center conductor, and connect them to each side of the dipole. We could then wonder why the darn thing fell down in the next wind storm. Even if you elect to not use the balun transformer, a ready-made center insulator is a must. Figure 28-4A shows a common form of center insulator for use with dipoles and other wire antennas. There is an SO-239 UHF coaxial connector on the bottom to accept the feedline and two eyebolts mounted with solder connectors for the antenna elements on the sidewalls. At the top is a small eyebolt that is used to strain-relieve the system, or to hang it from a center support (as when inverted-V dipoles are built).

A larger form of center insulator is shown in Fig. 28-4B. The enlarged down section indicates that this device contains either a 4:1 or 1:1 impedance ratio balun



28-3 Dipole antenna.



28-4A Center insulator.

transformer. The 1:1 size is recommended for ordinary dipoles, whereas a 4:1 unit is used for folded dipoles and certain other wire antennas.

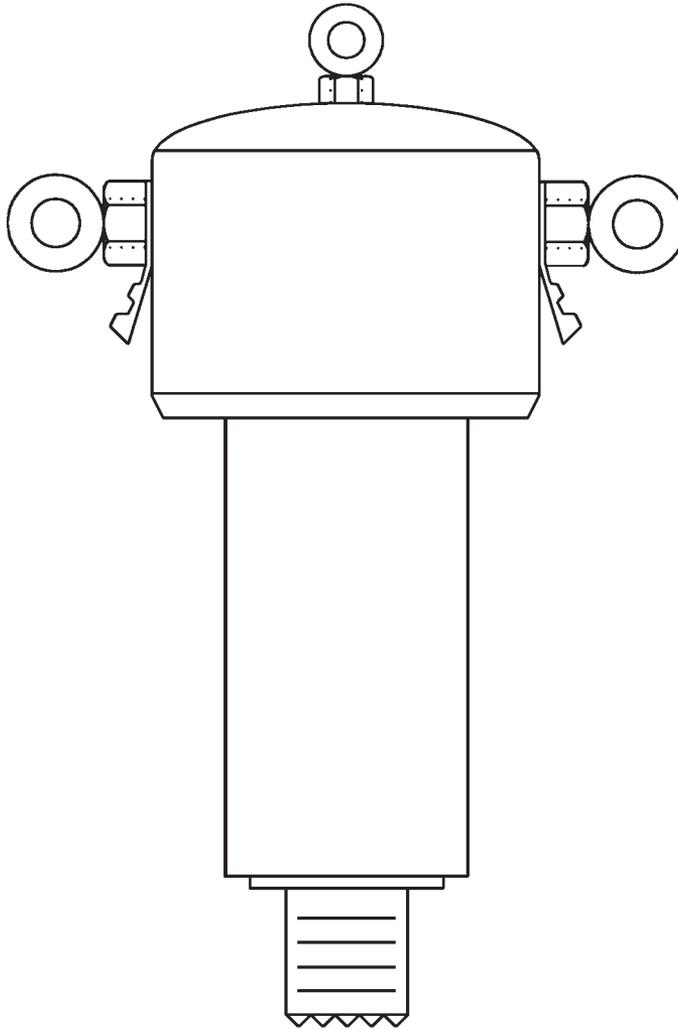
Connections to the device, whether straight center insulator or balun form, are shown in Fig. 28-5. For simplicity, only one side is shown (the other side is identical). The elements (*B* in Fig. 28-3) typically are made of no. 12 or no. 14 antenna wire. The best wire is stranded and is copper-clad steel core wire (Copperweld).

The end of the antenna wire is passed through the eyebolt and then doubled back and twisted onto itself 7 to 10 times. The remainder of the wire is then soldered to the terminal. Some people prefer to use a separate “pigtail” wire to connect the antenna element to the solder terminal. It is fastened to the antenna element wire between the 7- to 10-turn twist and the eyebolt.

When preparing the center connector/balun, it is good practice to use steel wool or sandpaper on the solder lug to remove oxidation. Oxidation makes it hard to solder and forces you to use too much heat to make a good connection. Also, be sure to have a large pair of pliers to crimp the wings of the solder lugs onto the wire.

Solder assists in making a good connection at the solder lugs, but the crimp should be good enough to do the job. The solder then provides corrosion protection. Also, tin the 7- to 10-turn twist section with solder. This does not increase the strength or provide better electrical connection—it is protection against the corrosion that will occur with copper wires.

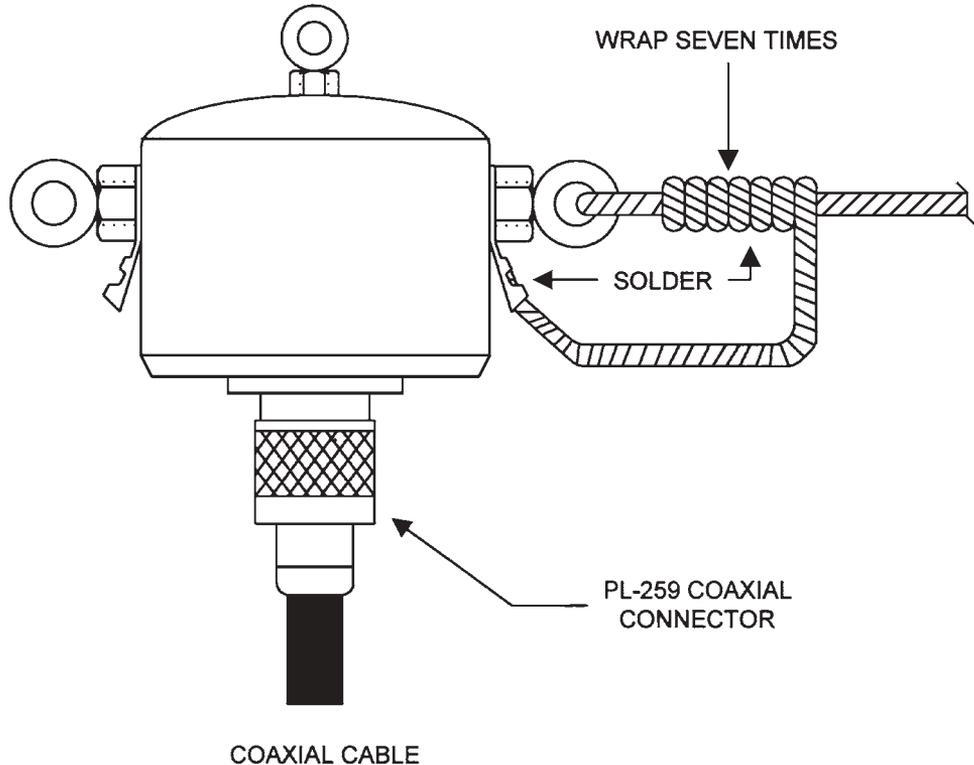
**Note** Use only resin-core solder for antennas! Typical solder is 50/50 or 60/40 lead/tin ratio and may be labeled “radio-TV” or “electronic” or something similar. Plumber’s solder is acid core and will destroy your antenna within a short time after it is erected.



**28-4B** Balun/center insulator.

Figure 28-6 shows the dipole installed. The ends of the two wire elements are connected to end insulators, which are in turn supported by rope to a vertical support (e.g., mast, tree, roof line of the house, etc.). Although shown nice and straight here, the actual installation will droop a bit in the center due to the weights of the center insulator or balun, the coaxial cable, and the wire itself. Gravity!

Note that the coaxial cable is formed into a 15- to 25-cm loop just below the coaxial connector to the center insulator or balun. This forms a species of RF choke to keep currents from flowing in the outer shield and improves the pattern. The coaxial cable loop is bound together with black electrician's tape or some similar adhesive medium and then fastened to the top eyelet with stout string or fishing line for strain relief.



28-5 Center insulator/balun connections.

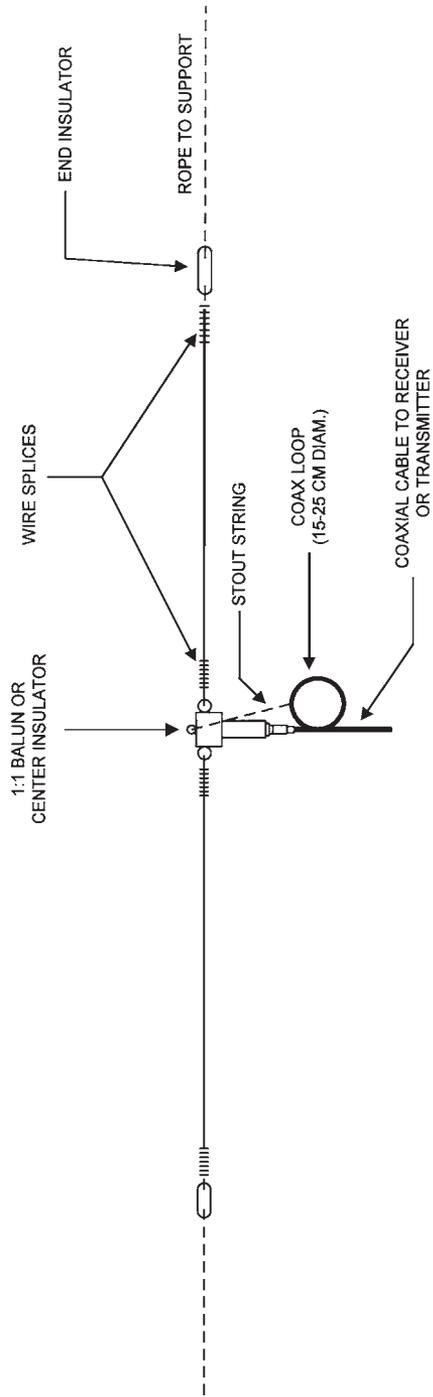
### Support rope

The support rope selected should be large enough to provide support but small enough to fit through the holes of the end insulators. Cotton rope is singularly unsuited to this task because it weathers poorly and will rot through in short order. Nylon rope, sail rope, and parachute rope are perhaps the best types, especially if a particular lot has been tested for high tensile strength. Remember, the wind loading of your antenna, even a short dipole, can be tremendous!

### End insulators

End insulators come in a large variety of shapes and sizes (I have even seen one chap use old toothbrushes, suitably drilled, of course). The main categories, however, are represented in Fig. 28-7. The version in Fig. 28-7A is the standard form, whereas that in Fig. 28-7B is an “egg” insulator. The egg insulator has two sets of wire grooves and through-holes that are orthogonal to each other (only one set is shown). Wire passes through one hold and set of grooves, and the supporting rope passes through the other hole/groove set.

The standard form shown in Fig. 28-7A has a number of grooves or ridges cut along the body. These grooves are intended to lengthen the electrical path between



28-6 Dipole and wire antenna installation.

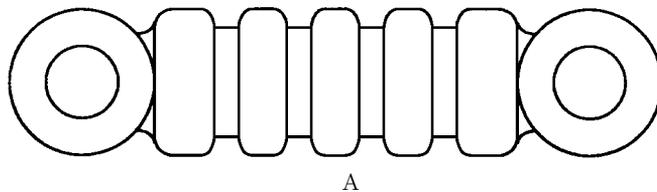
ends of the insulator and therefore serve to reduce losses in the insulator. Typical materials used to construct insulators include ceramic, glass, nylon, and plastics.

Figure 28-8 shows the connections to the end insulator. The electrical connections to the antenna wire are the same as for the center insulator. The only proviso here is that for transmitting antennas, it is wise not to leave any sharp points sticking up. The end of the dipole is at high RF voltages, so sharp points tend to cause corona sparking. Cut off any extraneous leads that form sharp points, and then smooth the whole affair down with tinning (i.e., solder).

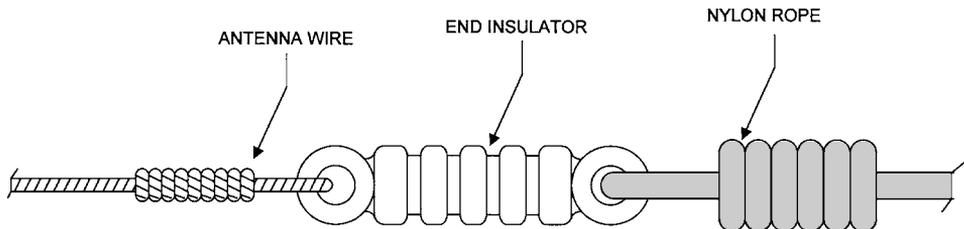
The rope is treated in a similar manner as the wire, but it must be knotted in some manner that tends to pull together as the rope is pulled into tension. Some people use the “hangman’s knot,” but this seems a bit excessive.

### Strain relief of the antenna

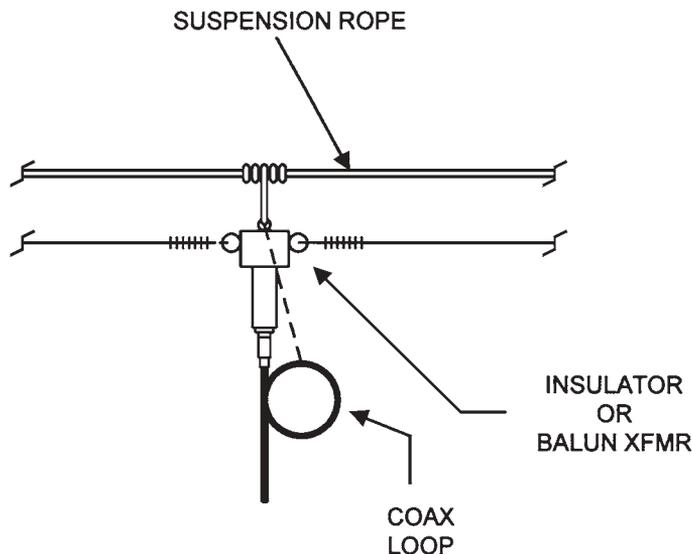
Figure 28-9 shows a method of strain relief for the center insulator or balun. This point is at the highest tension and probably is the weakest because of the number of



28-7 End insulators.



28-8 Connections to end insulators.



28-9 Strain relieving the center insulator or balun.

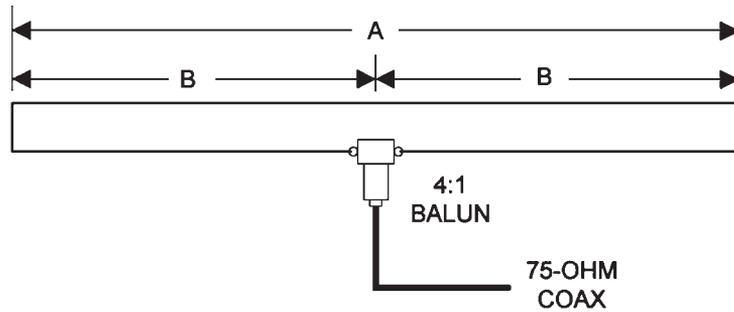
connections and things that can break. A suspension rope is run between the antenna's vertical supports, parallel to the antenna wire. It is tied to the top eyelet on the center insulator or balun by a short bit of rope about 15 to 20 cm long.

## Special case of the folded dipole

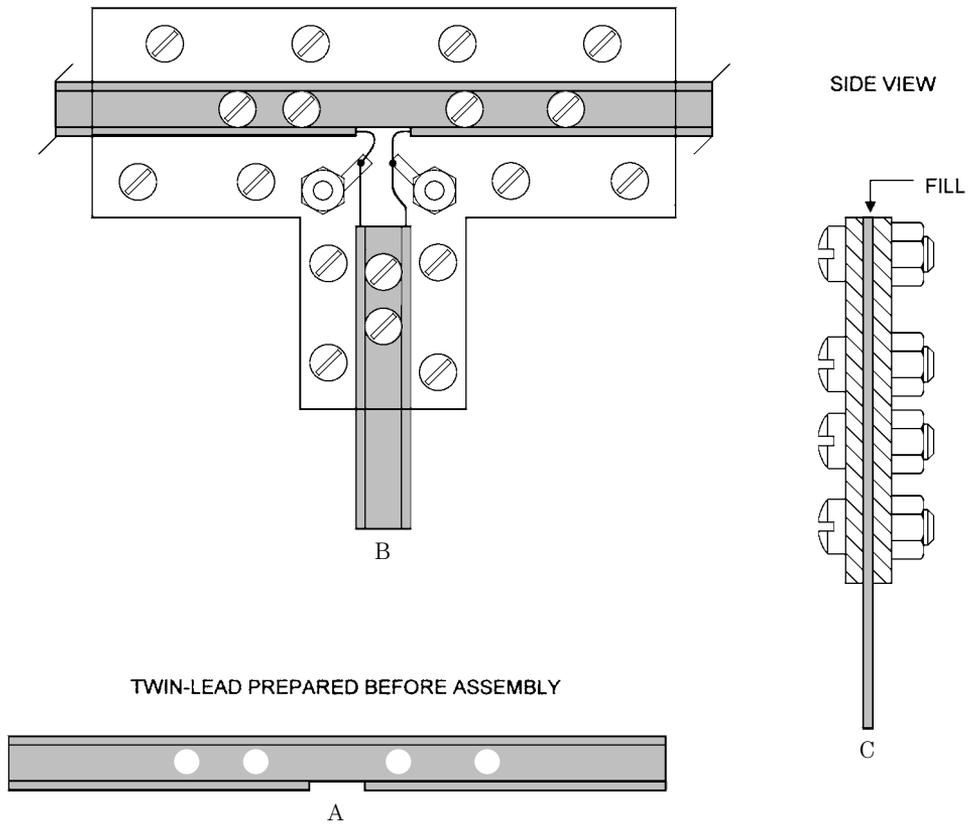
The folded dipole (Fig. 28-10) is a special case of the half-wavelength horizontal dipole. It consists of two close-spaced conductors shorted together on the extreme ends. One of the conductors is split in the center and used as the feedpoint. The feedpoint impedance at resonance is around  $280\ \Omega$ , so it makes a good match to  $300\text{-}\Omega$  television antenna twin lead. The usual practice is to build the entire antenna and feedline out of twin lead. An alternative (as shown in Fig. 28-10) is to provide a 4:1 balun transformer at the feedpoint to match  $75\text{-}\Omega$  coaxial cable. Despite the fact that  $300\text{-}\Omega$  TV twin lead is disappearing from the market, the folded dipole maintains a certain following among SWLs and hams.

The problem is how to install the antenna such that the terribly weak  $300\text{-}\Omega$  twin lead wires do not break at the slightest provocation. Figure 28-11 shows a solution used by one fellow (admittedly a fine worker in plastics and other materials). He fashioned a center insulator and the end insulators from a piece of strong Lucite material.

The  $300\text{-}\Omega$  TV twin lead is prepared as shown in Fig. 28-11A. The antenna element is notched in the center (break along the bottom) for about 1 to 1.5 cm. A hand punch is used to place two or three holes in the center insulation on either side of the break. Either a large-size leather punch or a paper punch can be used to make



28-10 Folded dipole.



28-11 Center insulator for ribbon cable folded dipole: (A) cable preparation, (B) front/back view, (C) edge view.

the holes. In a test, I was able to make the required holes using a hand-operated paper punch, but it required a large force (and turned my hand temporarily purple!).

The center insulator is shown in Fig. 28-11B. It is made from a piece of strong plastic, Lucite, or other insulating material. Two identical sections, front and back, are needed. A number of 5-mm holes are drilled into both pieces at the points shown to clamp the twin lead. A pair of solder lugs is provided to make connections between the antenna element and the transmission line (either twin lead or leads from a balun). The screws and nuts used to fasten the two halves of the insulator are made of nylon to prevent interaction with the antenna.

A side view is shown in Fig. 28-11C. Note that the twin lead causes a gap that can catch water. It also makes it possible to break one or both insulators by overtightening the nuts. In order to prevent this, a gasket of similar material is glued into the space as filler. The gasket material should be about as thick as the twin lead.

If you opt for a balun transformer, then it can be wired to the downleg of the tee insulator using the holes that normally would clamp the twin-lead transmission line.

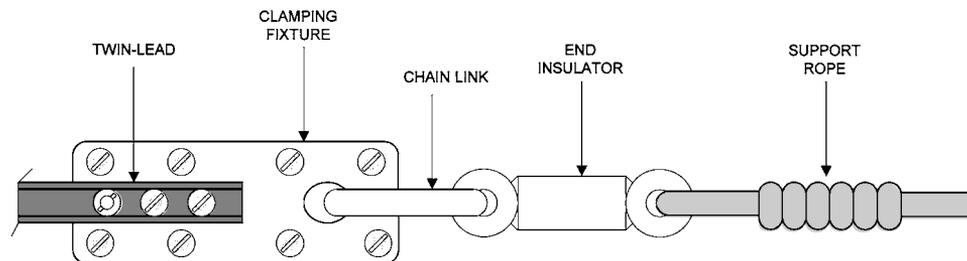
An end insulator for twin-lead antennas is shown in Fig. 28-12. It is constructed in a similar manner to the center insulator. Two scenarios are possible. The first, and that used in a test antenna, is to make the clamping fixture from metal such as 3- to 6-mm brass or copper stock. The ends of the antenna wires are shorted, so this is not a great problem.

An end insulator is then used with a rope in the normal manner. A chain-link section is used to fasten the clamping fixture and the end insulator. I attempted this with a nylon chain link, about  $1.5 \times 2.5$  in, from a do-it-yourself hardware store. It was only with a bit of difficulty that the ends were pulled apart, slipped into both the end insulator and the clamping fixture, and then reclosed.

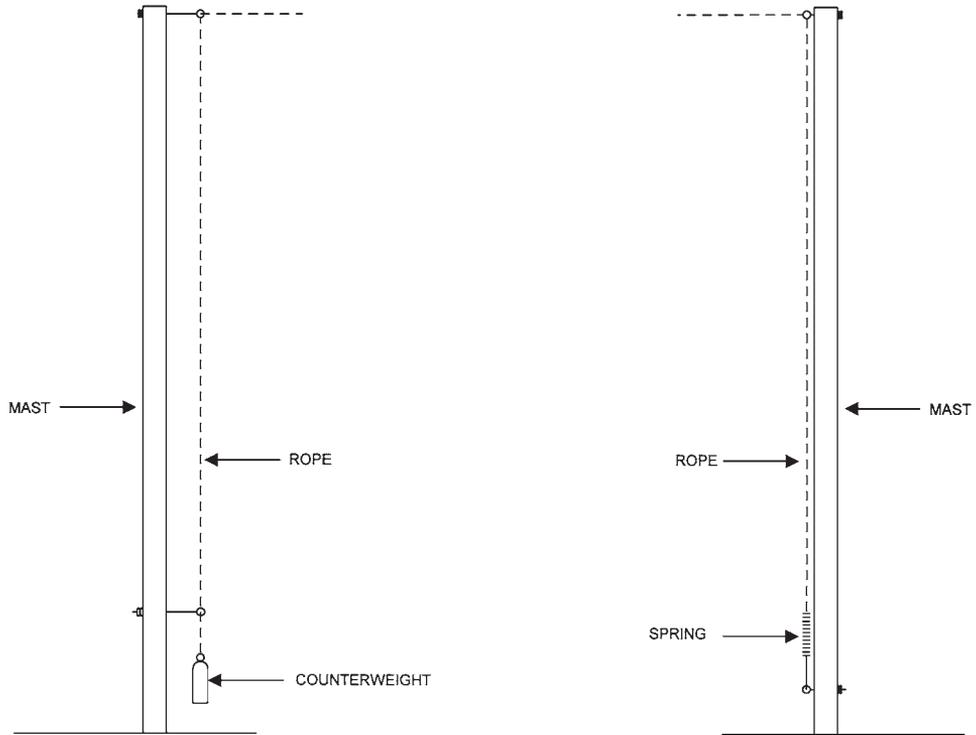
The second approach is to make the clamping fixture out of the same material as the center insulator. It then becomes the end insulator, so the other end insulator and that cursed chain link can be ignored. If you opt for this approach, it is a good idea to smooth and polish the rope hole on the clamping fixture to prevent chaffing of the rope.

## Masts and other supports for wire antennas

The end supports of the antenna can be anything that provides height: a mast, a tree, or the roof line of a building. Figure 28-13 shows the use of a mast, but either of the other forms of support could be used as well. The support rope is not tied off at the



28-12 Folded dipole end insulator.



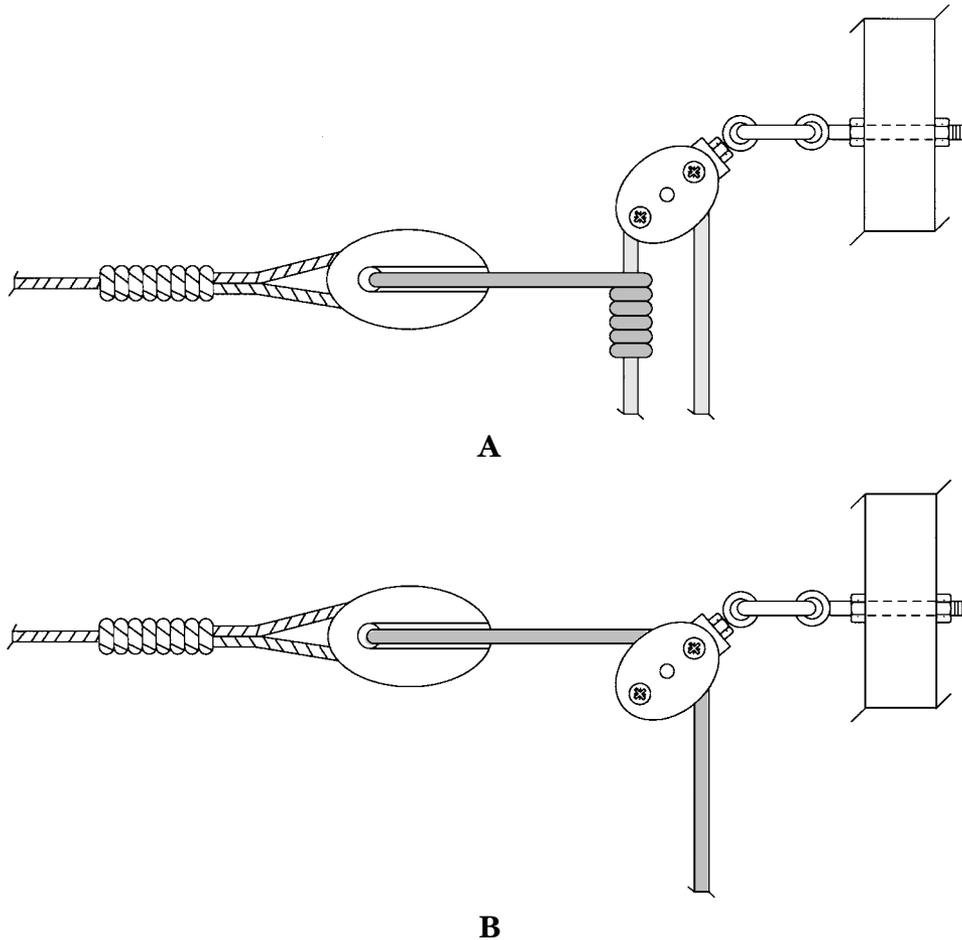
**28-13** Mounting mast schemes.

top, as is true in all too many installations, but rather is brought down to ground level. This approach facilitates raising and lowering the antenna for maintenance and tuning. Be sure to provide enough “dead slack” to make lowering the antenna feasible.

The antenna support rope is connected to a spring on one end and a counterweight on the other end. The spring should be stout enough so that it is only slightly extended when the full weight of the antenna places it in tension. I have used door-springs for this task, but only the stoutest varieties.

The counterweight should be just enough to balance the weight of the antenna and keep it not quite taut. As wind moves the antenna up and down, the counterweight raises and lowers, thereby reducing the chances of straining the antenna wires to their limit. Any number of counterweights can be used. I have seen a number of types: drapery cord weights, a small bucket of rocks, a gaggle of fishing weights, and (in one case) a burned-out automobile starter motor (the mounting hole on the front boss of the motor was ideal for accepting the rope).

Figures 28-14A and 28-14B show two methods for making the connections at the top of the mast. Although egg insulators are shown here for the sake of fairness (the alternate form was shown earlier), either form can be used. A pulley is mounted on the top of the mast with a link section and a stout eyebolt that passes through the mount. In the case of a tree, I have used a steel band of the type used to mount TV antennas to house chimneys. By making the down rope a closed loop, you gain the



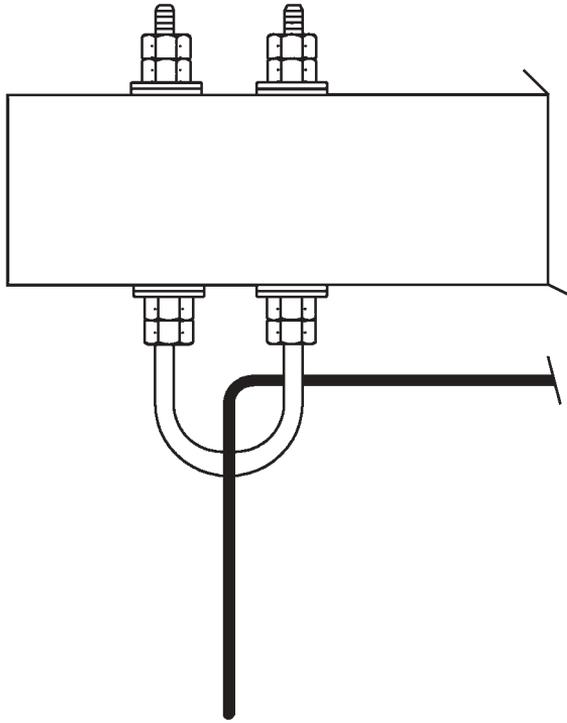
**28-14** Two alternate pulley systems for wire antenna mounting.

ability to raise and lower the antenna. But do not let a loose end slip, or you will be chasing the rope up the mast as it slips through the pulley.

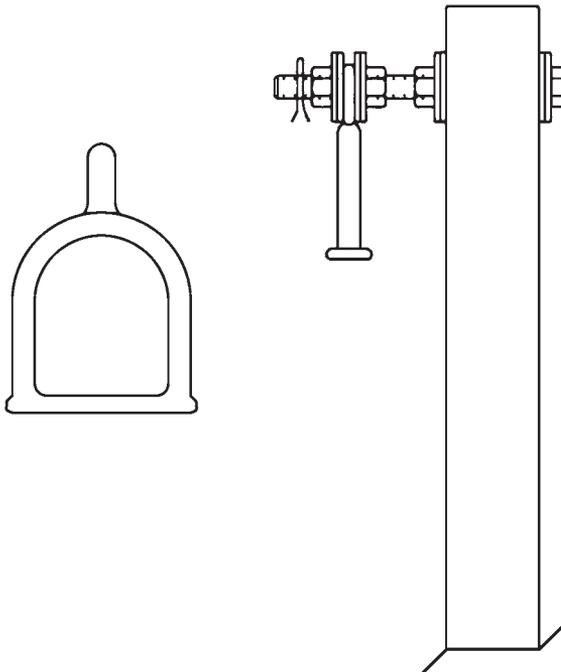
Pulleys work nicely, but they are mechanical contraptions that tend to fail occasionally. When they fail, the rope gets stuck. Pulleys also can corrode rather badly, with the same end result. All in all, a pulley is not the best solution for most people.

Figure 28-15 shows a better way. In this approach, a U-bolt is fastened to the top of the support mast. If the U-bolt is made of brass, then all the better, for it will not corrode. The rope can be passed through the U-bolt as in the pulley system but without the mechanical failure problem. Be sure to install the U-bolt all the way up to the threads so that the rope will not chafe against them.

The method shown in Fig. 28-16 can only be used by a horse lover! This method uses a brass stirrup in place of the U-bolt. If you have switched to cars or trucks and no longer need your horse stirrups, then perhaps this is a viable means of attaching the support rope to the top of the mast.



28-15 U-bolt end mounting.



28-16 Stirrup end mounting.

## Installing vertical antennas

Despite continuous (and mostly untrue) stories of very poor performance, the four- or five-band commercially manufactured trap vertical antenna remains one of the most popular amateur radio “antlers” in the HF bands. They are economical compared with directional rotating beam antennas, easy to erect, and do not have to occupy a lot of real estate (the “footprint” of the vertical can be very small, especially if you bury the radials). Unfortunately, a misinstalled vertical is both dangerous and a very poor performer. This section covers both problems in order to give you a good chance of success.

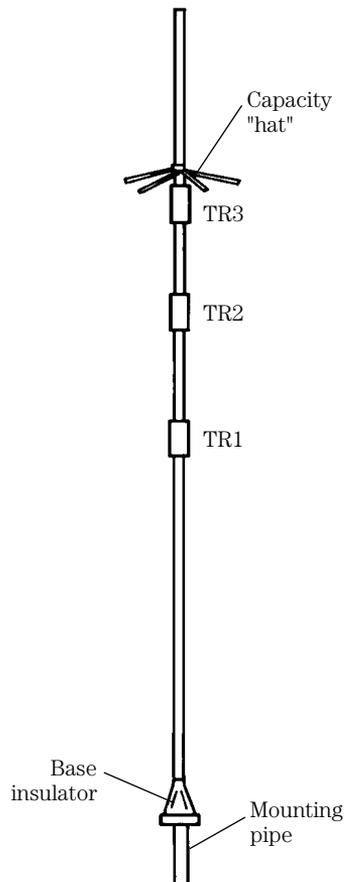
### Electrical installation

Figure 28-17 shows the usual form of multiband trap vertical antenna. Each trap ( $TR1$ – $TR3$ ) is a parallel resonant LC tank circuit that blocks a certain frequency but passes all others. In Fig. 28-17,  $TR1$  is the 10-m trap,  $TR2$  is the 15-m trap, and  $TR3$  is the 20-m trap. No 40-m trap is needed because the antenna resonates the entire length of the tubing on 40 m. Each section (except perhaps the 10-m section) is actually a little shorter than might be expected from the standard quarter-wavelength formulas. This is because the traps tend to act inductively and so lessen the length required to resonate on any given band.

The vertical antenna manufacturer might give suggested lengths for the various segments between traps. Do not make the mistake of assuming that these are absolute numbers. They are only recommended starting points, even though the literature packed with the antenna may suggest otherwise. Loosely (meaning do not tighten the clamps too much) but safely install the antenna, and then adjust each segment for resonance. Start with the 10-m band, and then work each lower-frequency band in succession: 10-15-20-40, etc. After each lower band is adjusted, recheck the higher bands to make sure that nothing has shifted, because there might be a little interaction between bands. Once the antenna is properly resonant, tighten the clamps, and make the final installation. I know this is a pain in the neck, and means putting the antenna up and taking it down a couple of times, but it pays dividends in the end. I failed to do this once and found that the 15-m band was useless; it resonated at 19.2 MHz.

Radials make or break a vertical antenna; they form the ground plane for the antenna. AM broadcast stations typically use vertical antennas and must have 120 radials for the ground plane. Antenna reference books usually contain a graph plotted to show the effectiveness of radials and demonstrate a decreasing return on investment after about 32 radials. For amateur work, I recommend not less than 2 radials per band, and preferably 4, arranged so that they are equally spaced around the antenna. If you cannot space them correctly, never fear; they will work anyway. On a four-band antenna, this means 16 radials, which really is not a lot. Of course, the general rule is that the more radials, the better is the antenna, at least up to the point where diminishing returns are realized.

The radials are made of no. 14 wire and must be a quarter wavelength ( $246/F_{\text{MHz}}$ ). Figure 28-18 shows how to mount the radials on a mast-mounted (or roof-mounted) installation, and Fig. 28-19 shows radials in a ground-mounted situa-



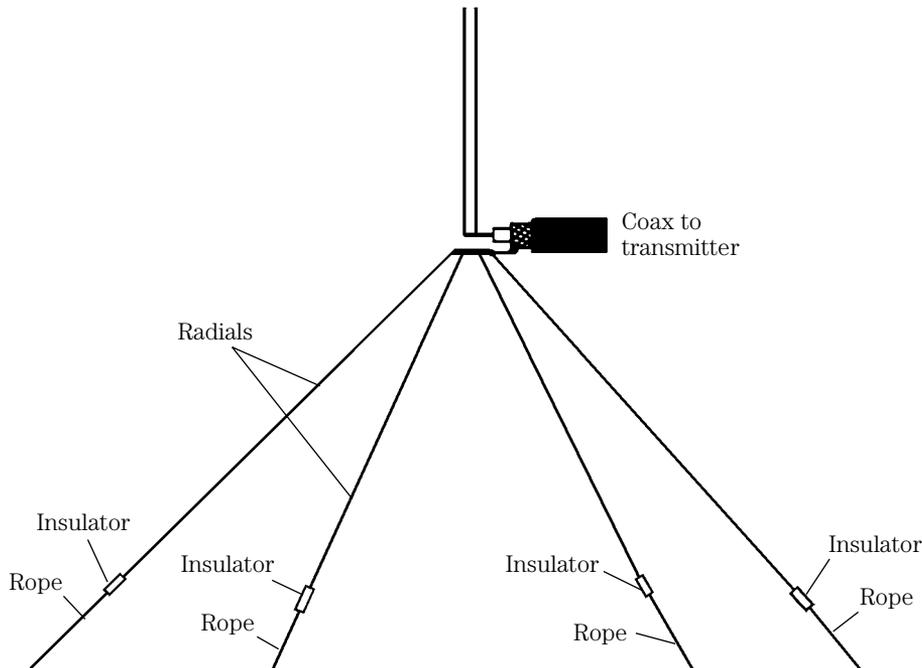
28-17 Mounting commercial trap vertical.

tion. I do not recommend radials on the surface for ground-mounted antennas. It is too easy for a guest or even a trespasser to trip over the radials, and this could land you in court. Wire radials can be buried using a spade tip to cut a 3- to 4-in slit, and they even make for a better ground.

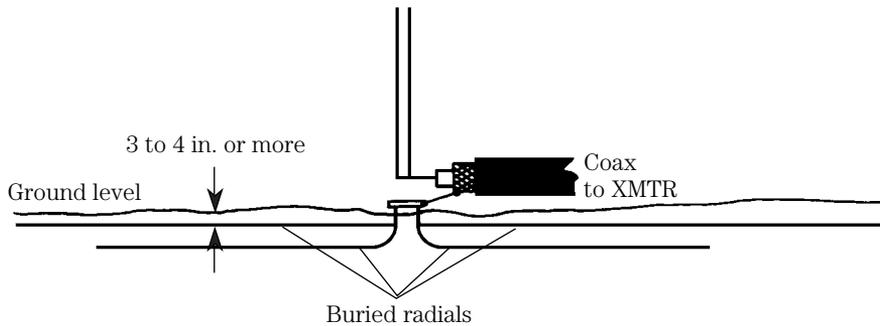
### Ground mounting verticals and masts

In the preceding section, the essential assumption was that the antenna is mounted above ground, on a mast or other support. In this section you will take a look at the methods of ground mounting the antenna or its mast. The same mast ideas are also useful for erecting supports for dipoles and other wire antennas.

Figure 28-20 shows a standard chain-link fence post used as a support for either a vertical antenna or a mast (which could be metal, PVC, or wood). The typical fence post is 1.25 to 2.00 in outside diameter and made of thick-wall galvanized steel. Thus such a pipe makes a tough installation. Some authorities believe that the pipe will last longer if painted its entire length (especially the underground portion) with a rust-inhibiting paint such as Rustoleum.

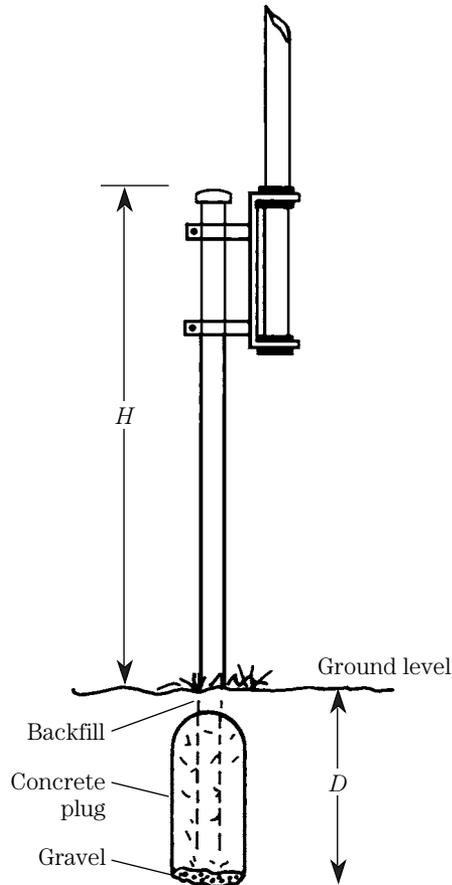


**28-18** Feeding the mast-mounted vertical antenna.



**28-19** Feeding the ground-mounted vertical antenna.

The beauty of the fence post as an antenna support is that it is relatively easy to install, and supplies are obtained easily from hardware stores. The post is mounted in a concrete plug set at depth  $D$  in the ground (see again Fig. 28-20). The depth  $D$  is a function of local climate and local building codes. In the absence of local regulations to the contrary, make the hole at least 24 in deep (or more). The depth is determined by the 100-year depth of the frost line in your area, a figure well familiar to local building experts. Keep in mind, however, that shallower depths may be legal for



28-20 Fence post mounting.

fence posts, but installation of a vertical antenna (or mast) on top of the fence post changes the mechanical situation considerably.

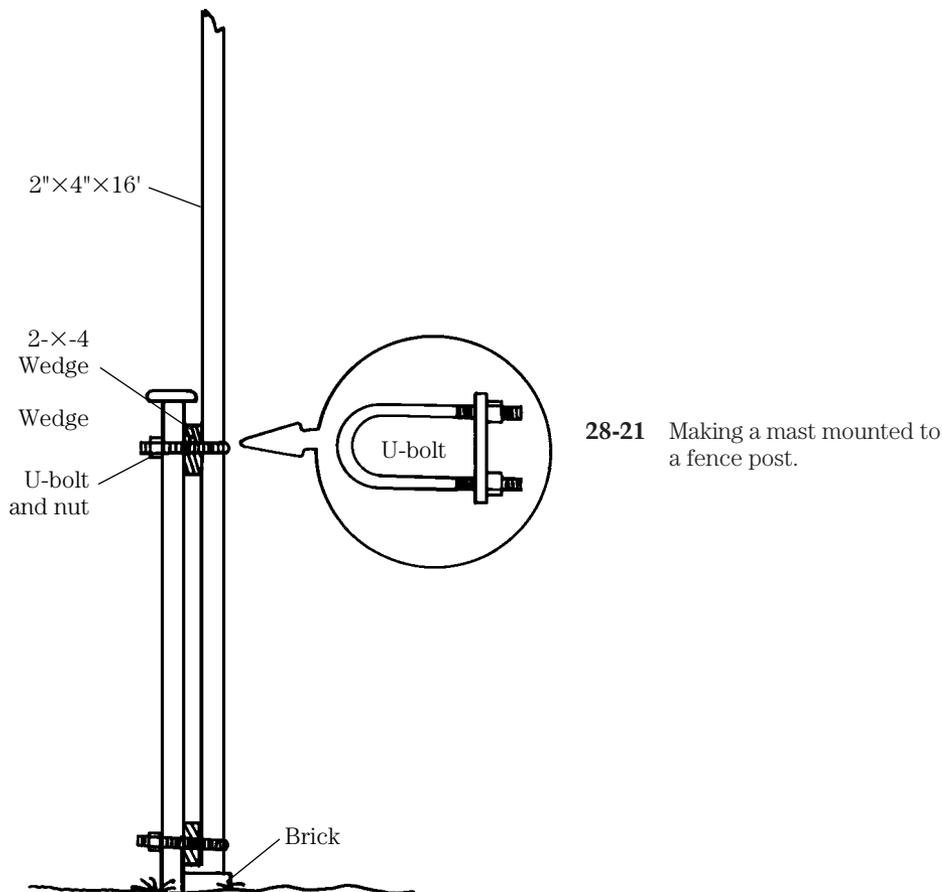
The hole can be dug with a post-hole digger tool or an earth-auger-bit tool. The latter often can be rented in either gasoline engine, electrical motor, or manual versions from tool rental stores.

Once the hole is dug, place about 4 in (or local requirement) of gravel at the bottom of the hole. This gravel can be bought in bags from hardware stores for small jobs. The post is installed at this point and made plumb (i.e., it is made to sit vertical in the hole). Gravel can be used to force the post to remain upright as the job is finished. On top of the gravel is a concrete plug that fills the hole to at least 4 in below the surface. The rest of the hole is backfilled with dirt, unless you want the plug to extend all the way to the surface. Do not put anything on top of the post for at least 4 days or whatever period of time the concrete manufacturer recommends. Concrete needs time to cure, and 4 to 7 days seems to be the recommended period of time. It is probably a good idea to install the post one weekend and then complete the antenna installation the following weekend.

Chain-link fence posts tend to be about 4 ft high, and this is sufficient for installation of masts for inverted-V and dipole support. However, for vertical installation, make the height  $H$  of the post about 18 in (otherwise, the antenna tries to act differently from the instructions).

If the antenna is a standard vertical with the usual form of offset base (as shown in Fig. 28-20), then use either the mounting brackets that come with the antenna or U-bolts fastened around the fence post. Never use a single mounting point; always use at least two-point mounting to prevent the antenna from pivoting and shearing off the mounting hardware.

Figure 28-21 shows how to mount a  $2 \times 4$  mast to a fence post. Good-quality  $2 \times 4$  lumber can be purchased in lengths up to 20 ft, although, 8-, 10-, 12-, and 16-ft lengths are the most commonly available (in my area, 16- and 20-ft lengths have to be bought at professional “contractor” lumberyards rather than those which serve homeowners). The  $2 \times 4$  selected should be kiln-dried and pressure-treated. It also should be painted in order to minimize the effects of weathering on the mechanical integrity of the installation. Good-quality  $2 \times 4$  lumber is relatively inexpensive, but

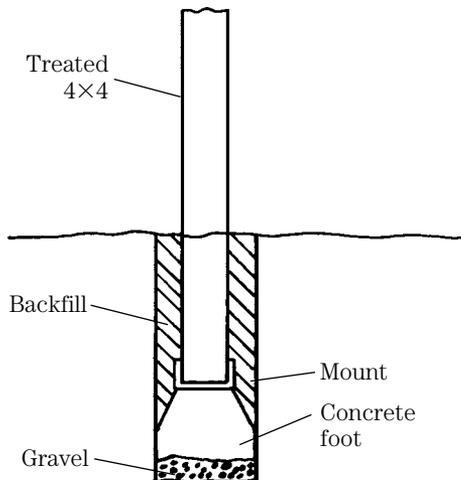


the less costly material (which is only marginally less costly than the best) will warp and ruin the installation within 1 year.

The  $2 \times 4$  is attached to the fence post using U-bolts. A  $2 \times 4$  scrap is used as a wedge to take up the difference between the post and the mast. In some installations the U-bolt will go around the perimeter of the  $2 \times 4$ , whereas in others a pair of holes will be drilled in the  $2 \times 4$  to admit the U-bolt arms. The U-bolt must be at least  $\frac{1}{4}$ -in no. 20 thread, and either a  $\frac{5}{16}$ -in bolt or a  $\frac{3}{8}$ -in bolt is highly recommend if the mast supports any significant weight. Some installers like to place a brick at the base of the  $2 \times 4$ . This support bears the static and dynamic loads (i.e., gravity and wind) that eventually would cause the mast to list because of the eccentric loading on the footer. The auxiliary purpose is to keep the butt end of the  $2 \times 4$  off the ground and thereby prevent rot. Some additional mechanical stability might be afforded by replacing the brick with a cinder block that is filled with concrete. Another approach is to install a pressure-treated  $4 \times 4$  of the sort used to support decks (Fig. 28-22). Hardware stores and lumber yards sell the specially treated  $4 \times 4$  posts and the concrete for the footers. These footers are dropped onto a gravel bed at the bottom of a hole (see the preceding section for a discussion of how deep the hole must be). A typical deck footer has a plastic or heavy-duty metal mount to support the  $4 \times 4$ . After the footer and mount are installed, the hole is backfilled with soil. Some people prefer to backfill, at least to a point above the mount, with builder's gravel (of the same sort as used beneath the concrete footer), or concrete can be used for maximum stability.

### Installing towers

Antenna towers are used extensively in amateur and commercial radio communications systems for a very obvious reason: Antennas tend to work better "at altitude" than on the ground. Amateur operators tend to install 30- to 100-ft towers, although the 40- and 50-ft models probably predominate for practical (and in some areas

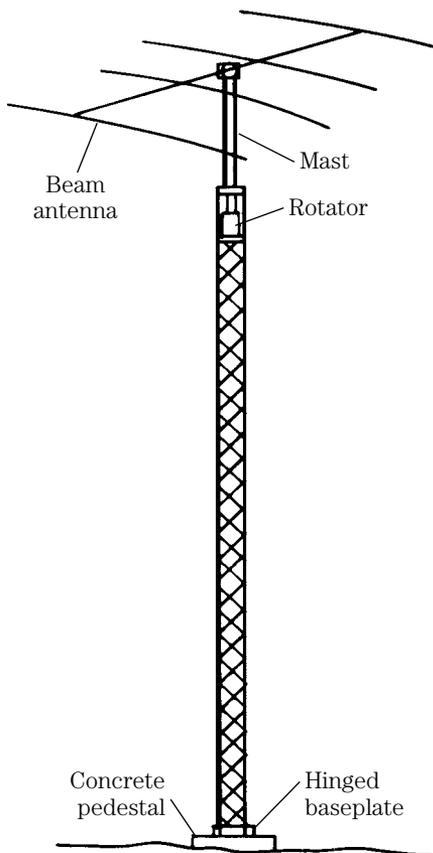


28-22 A  $4 \times 4$  antenna support.

legal) reasons. Local laws vary considerably, from one area to another. Be sure to consult the bureaucracy before erecting a tower.

In my own area, a 30-ft tower that is physically attached to a two-story house does not need a permit or inspection (but must be constructed according to certain regulations), although anything mounted away from the house (or at a height greater than 30 ft, if attached to the house) requires a permit and the usual mechanical inspection. The law requires that an inspection be performed when the hole for the footing is completed and 4 in of gravel is on the bottom. The inspector will take a yardstick and measure both the depth of the hole and the depth of the gravel. A final inspection is performed when the antenna is installed. Prior to all this mess, a permit must be obtained (which means drawings submitted).

Several different types of towers are used by amateurs. Some are square-base designs, and others are triangular-base designs. At least one commercial model is a round pipe that has a rotator at the base. In this model, the entire tower turns when the antenna is rotated. Several tower configurations are seen commonly. Figure 28-23 shows the fixed form of tower. This type of tower costs the least, and a case can be



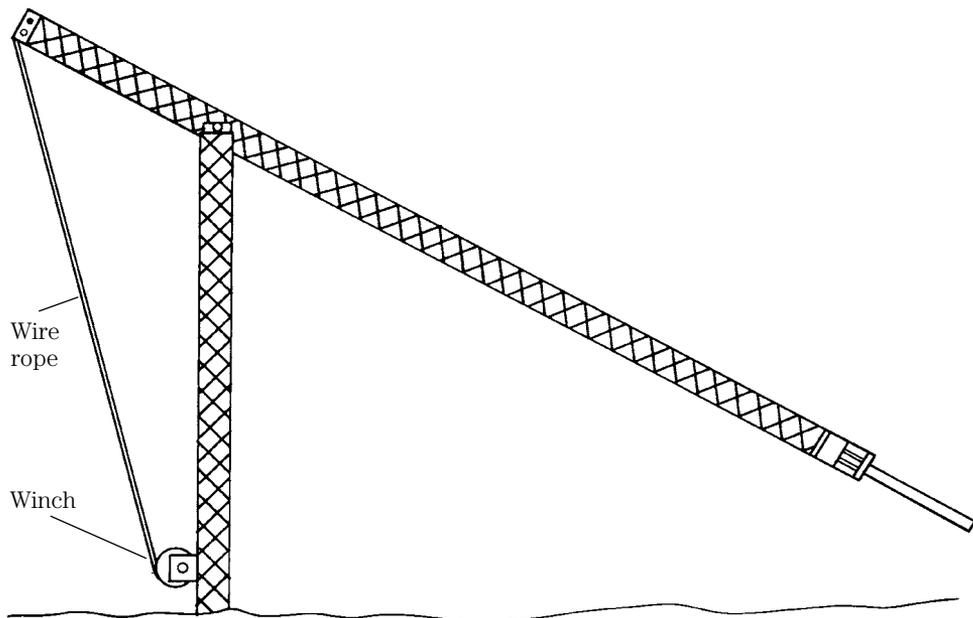
**28-23** Antenna tower on hinged baseplate.

made for it also being the most stable. It is made of 10-ft sections and an 8- to 12-ft top section that contains the thrust bearing for the antenna mast and rotator mounting platform. Some fixed towers are self-supporting up to a point, but the use of guy wires is always recommended unless a firm mounting point (against the side of a building, for example) is provided about 18 ft above the surface. The fixed tower can be made to tilt over at the base by using a hinged baseplate on the mounting pedestal.

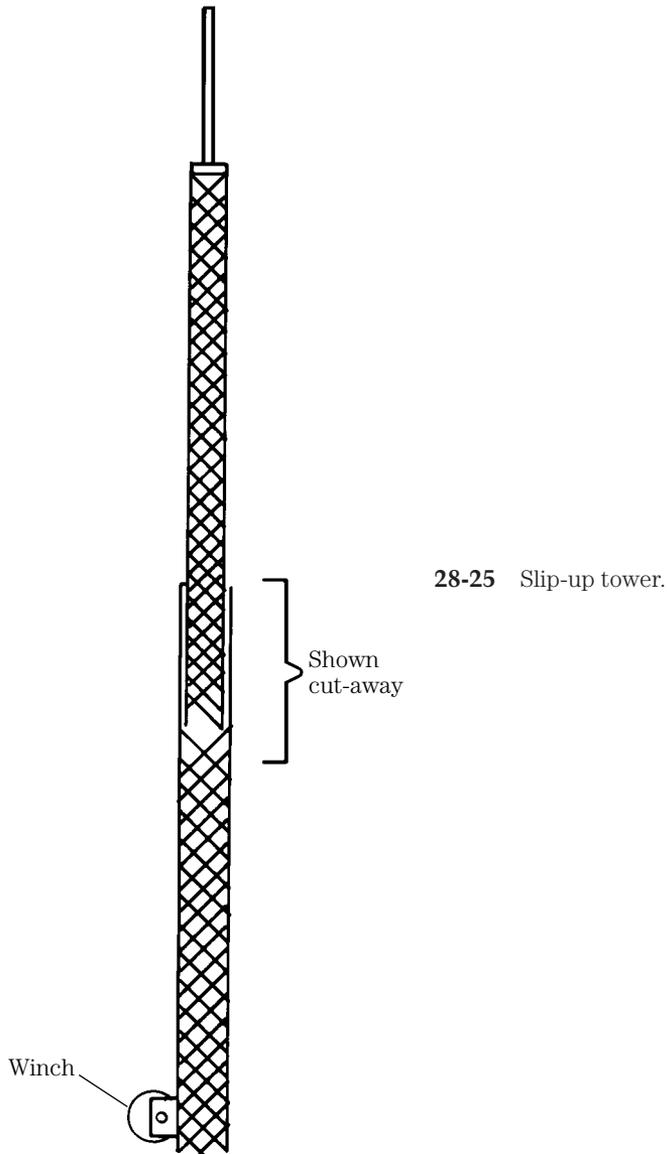
The tilt-over tower is shown in Fig. 28-24. This type of tower uses the same sort of construction as the fixed model but is hinged at a point between one-third and one-half the total height. A winch at the base drives a steel wire that is used to raise or lower the tower between fully extended and fold-over positions.

Still another configuration is shown in Fig. 28-25. This type of tower is the slip-up variety. A wide base section supports a smaller movable upper section. A winch is used to lift the upper section into its fully extended position. Typical examples will raise from a low position of 20 ft, to a height of 40 ft or so when fully extended (or 20/50 ft). When the tower is fully extended, it is locked into place at the bottom of the upper section with steel bolts. A lot of shear force is applied to these bolts, so it is wise to use several very hard stainless steel bolts for the lockdown.

There is a serious danger inherent in the design of the slip-up tower: the guillotine effect. If the upper section comes loose while you are working on it, it will come plummeting down the shaft formed by the lower section and shear off any arms, legs, or other body parts that get in the way. Whenever you work on this form of tower, use steel fence posts (or similar pieces of metal) as a safety measure (Fig. 28-26); at least two should be used, and both should be attached securely at both ends with

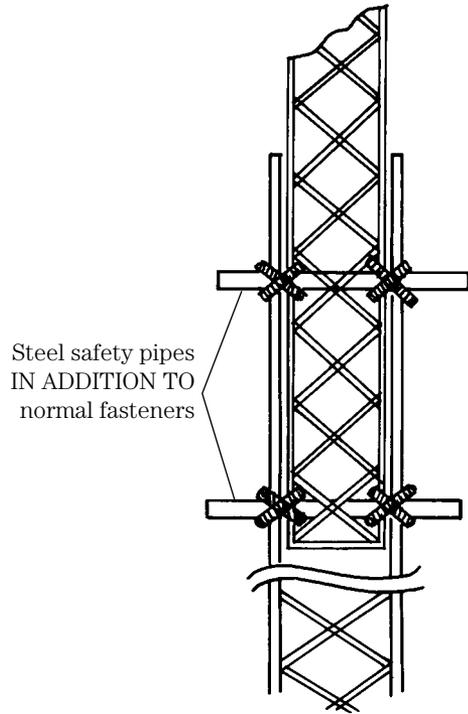


28-24 Fold-over tower.

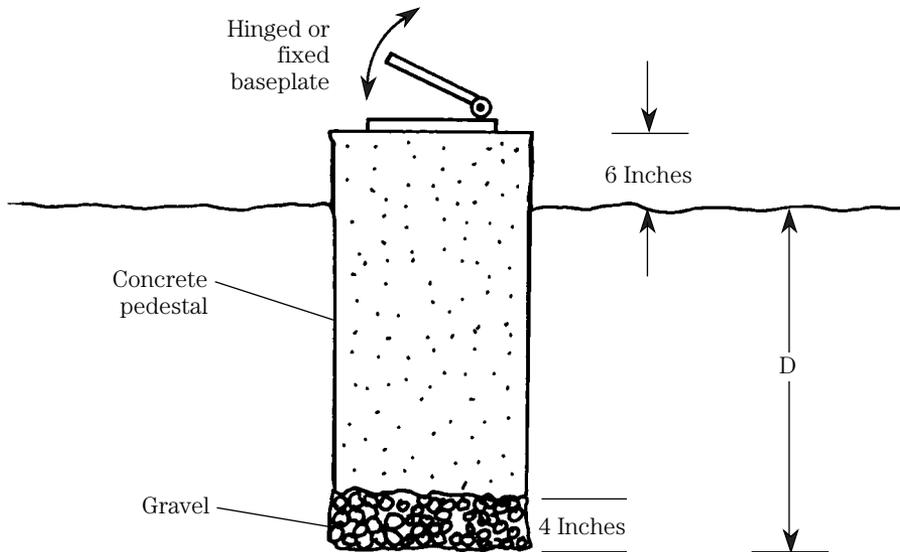


rope. The purpose of the rope is to prevent the safety pipes from working loose and falling out. These pipes are used in addition to, not instead of, the normal bolt fasteners that keep the antenna tower erect.

The antenna tower, regardless of its configuration, is typically installed on a buried pedestal made of concrete (Fig. 28-27). The construction must follow local laws, but there are some general considerations to keep in mind. The surface area of the pedestal is 242 to 362 in. The top surface is fitted with either a fixed or hinged baseplate (which must be installed before the concrete sets). The depth of the



28-26 Safety precaution.



28-27 Ground pedestal.

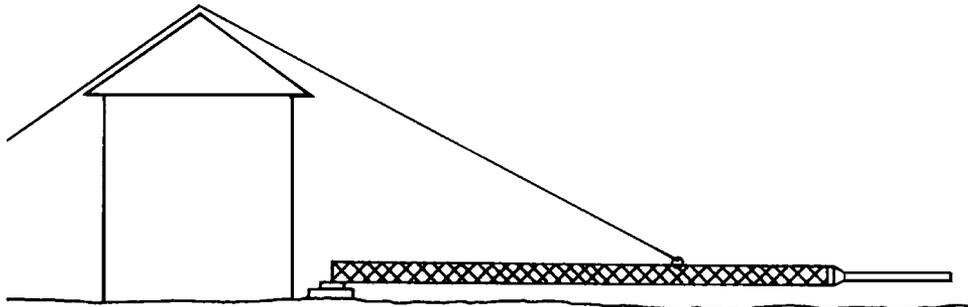
pedestal and the depth of the gravel base tend to reflect the local frost line. Regardless of how general local law is, good advice is to make the pedestal at least 28 in deep—even if local law permits less.

Raising the tower can be accomplished by either of two methods, although, in my opinion, only one of them is safe. Figure 28-28 shows a method by which the tower is laid out on the ground and then raised with a heavy rope over a high support (such as the peak of the house roof). This method is dangerous unless certain precautions are taken. For example, a support must be placed beneath the antenna as it is raised. This requirement means that a constantly increasing height of the tower demands a constantly increasing lower support.

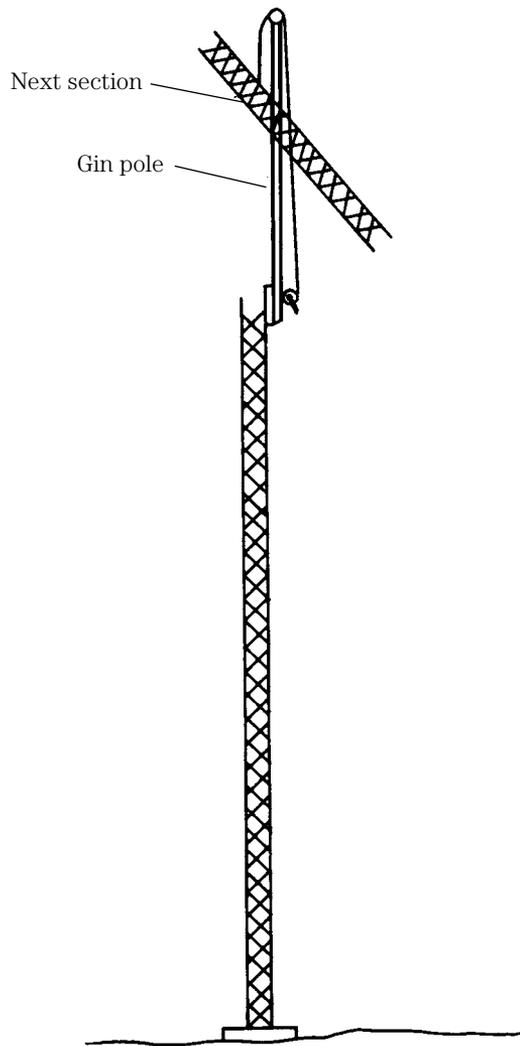
The end-over-end method of raising a tower is shown in Fig. 28-29. A “gin pole” is required for this job. The gin pole is a length of pipe fitted with a pulley at the top and a pair of clamps at the bottom. The gin pole is clamped onto the top end of the highest section so that it can be used to raise the next 10-ft section into place, where it can be bolted down. The gin pole is then raised to the top of the newly installed section, where the operation is repeated again. The end-over-end method is preferred by tower manufacturers and is the method recommended by most of them.

The antenna tower will be most stable if guy wires are installed at appropriate places determined by the tower manufacturer. Although many people are inclined to install towers “free standing” and some towers are advertised as such, it is always good practice to use guy wires attached to at least one point (or more, for tall towers). Use at least four guy wires on a square tower or three on a triangle tower (one in each direction). Figure 28-30 shows two methods for securing the guy wires to the ground. One approach is to bury a heavy concrete or metal “deadman” weight underground. Alternatively, use at least one 4- to 6-ft pipe (1 to 1.5-in in diameter) buried up to a length of 12 to 24 in (18 in is shown in Fig. 28-30).

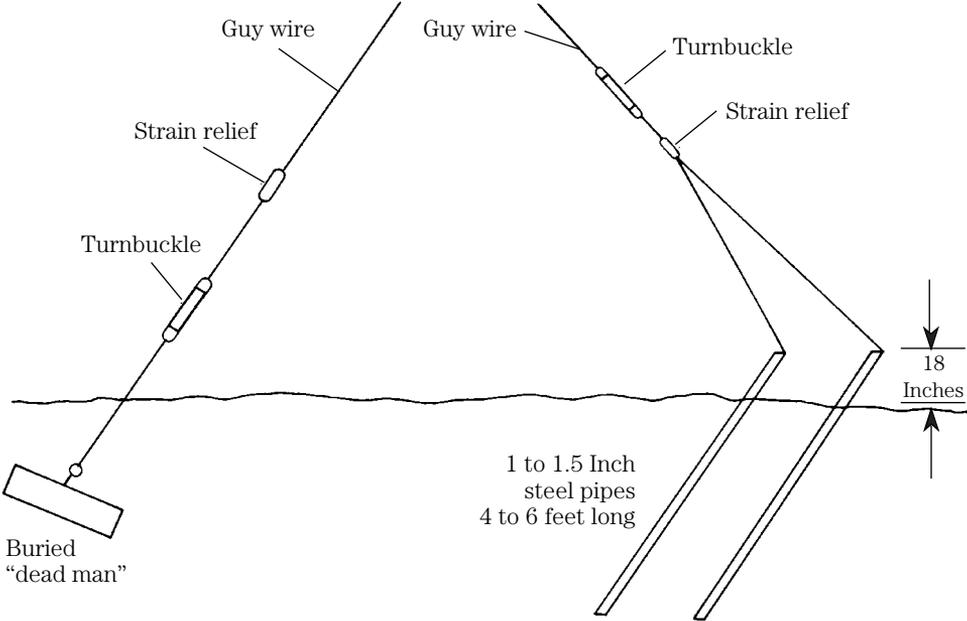
**Note** It is absolutely essential to use safety belts when climbing a tower. Use two belts that are completely independent of each other so that no “single-point failure” can eliminate the belt’s function. Always keep one belt fastened to the tower; when climbing, unhook one belt and move it up on the tower, and then disconnect the other only when the first is reattached. Be sure to inspect the belts prior to use. Leather and metal wear out and break, and you do not want to find that failure at 50 ft up. Also, do not buy cheap belts.



28-28 Dangerous method for erecting tower.



**28-29** Proper tower installation using gin pole.



**28-30** Ground stakes for guy wires.

## 29 CHAPTER

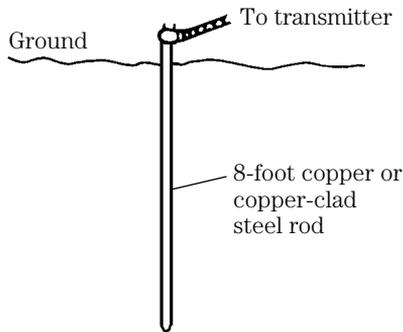
# Grounding the antenna: What is a good ground?

THE SUCCESS OR FAILURE OF A RADIO ANTENNA SYSTEM OFTEN (PERHAPS USUALLY) hangs on whether or not it has a good RF ground. Poor grounds cause most antennas to operate at less than best efficiency. In fact, it is possible to burn up between 50 and 90 percent of your RF power heating the ground losses under the antenna, instead of propagating into the air. Ground resistances can vary from very low values of  $5\ \Omega$ , up to more than  $100\ \Omega$  ( $5$  to  $30\ \Omega$  is a frequently quoted range). RF power is dissipated in the ground resistance. The factors that affect the ground resistance include the conductivity of the ground, its composition, and the water content. The ideal ground depth is rarely right on the surface, and depending on local water table level might be a couple meters or so below the surface.

It is common practice among some amateur radio operators to use the building electrical ground wiring for the RF antenna ground of their station. Neglecting to install an outdoor ground that will properly do the job, they opt instead for a single connection to the grounded “third wire” in a nearby electrical outlet. Besides being dangerous to work with, unless you know what you are doing, this is also a very poor RF ground. It is too long for even the low RF bands, and radiates RF around the house in large quantity. Stations that use the household electrical wiring as the radio ground tend to cause TV interference, broadcast interference, and other electromagnetic interference both in their own building and in nearby buildings.

Fortunately, there are some fixes that will help your situation. We can reduce the ground resistance by either altering the composition of the earth surrounding the ground point, or by using a large surface area conductor as the ground point.

Figure 29-1 shows the traditional ground rod used on small radio stations, including amateur stations. Use either a copper (or copper-clad steel) rod at least 6 ft long (8 ft preferred). Electrical supply houses, as well as amateur radio and communications equipment suppliers, also sell these ground rods. Do not use the nonclad



**29-1** Single ground rod works—almost.

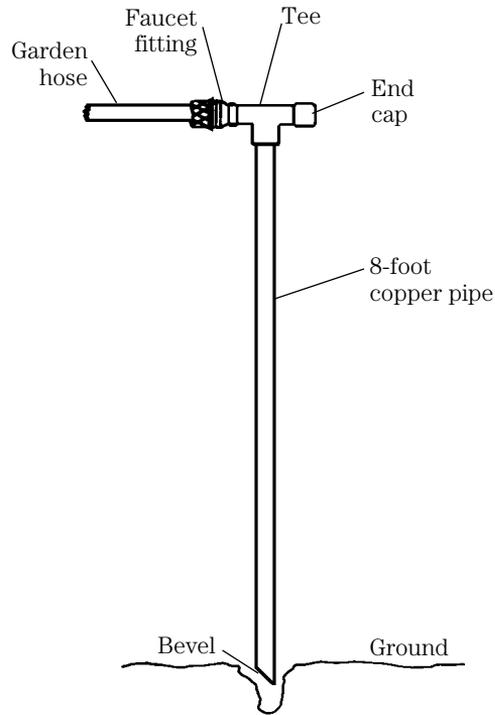
steel types sold by some electrical supply houses. They are usable by electricians when making a service entrance ground on your home or workplace, but RF applications require the low skin resistance of the copper-clad variety. The rod need not be all copper, because of skin effect forcing the RF current to flow only on the outer surface of the rod. Try to use an 8-ft rod if at all possible, because it will work better than the shorter kind. Do not bother with the small TV-antenna 4-ft ground rods; they are next to useless for HF radio stations. Drive the ground rod into the earth until only 6 in or so remains above the surface. Connect a ground wire from your station to the ground rod.

The ground wire should be as short as possible. Furthermore, it should be a low-inductance conductor. Use either heavy braid (or the outer conductor stripped from RG-8 or RG-11 coaxial cable) or sheet copper. You can buy rolls of sheet copper from metal distributors in widths from 4 in up to about 18 in. Some amateurs prefer to use 7-in-wide foil that is rated at a weight of 1 lb/linear ft. Sweat-solder the ground wire to the rod. You can get away with using mechanical connections like the electricians use, but you will eventually have to service the installation when corrosion takes its toll. I prefer to use soldered connections, and then cover the joint with either petroleum jelly or acrylic spray lacquer.

Another alternative is to use a copper plumbing pipe as the ground rod. The pipe can be purchased in 8-ft through 16-ft lengths from plumbing supply shops or hardware stores. The pipe selected should be  $\frac{3}{4}$  in or larger. Some people report using up to 2-in pipe for this application. The surface area of the hollow pipe is greater than that of a solid rod of the same diameter. Because of certain current flow geometries in the system, however, the ground resistance is not half the resistance of a rod of the same diameter, but is nonetheless lower.

Driving a long pipe into the ground is not easily done. Unlike the copper-clad steel rod, the pipe has no compression strength and will deform when you hit it with a hammer or other driving tool. To overcome this problem (see Fig. 29-2) you can use a garden hose as a water drill. Sweat solder a tee joint on the top end of the pipe, and then sweat-solder a faucet fitting that matches the garden hose end on one end of the tee joint. Cap off the other port of the tee joint. Use the tee

**29-2** Hose method of installing ground pipe.



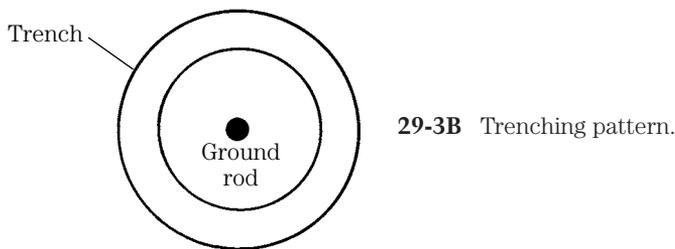
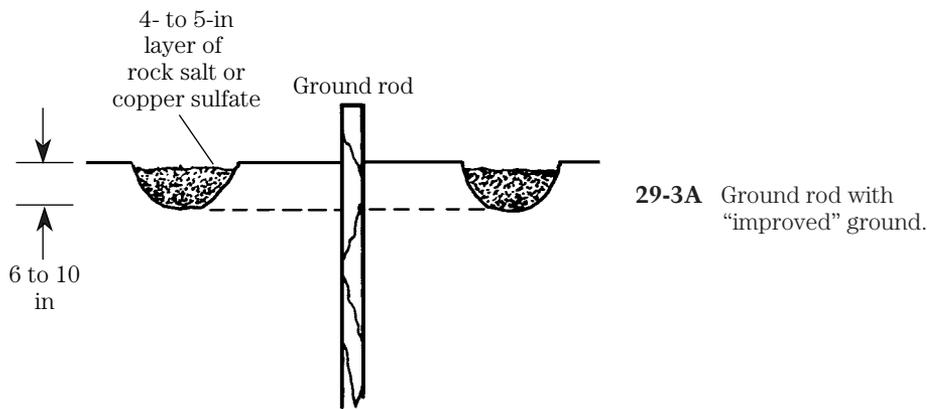
joint as a handle to drive the pipe into the ground. When water pressure is applied, the pipe will sink into the ground as you apply a downward pressure on the tee handle. In some cases, the pipe will slip into the ground easily, requiring only a few minutes. In other cases, where the soil is hard or has a heavy clay content, it will take considerably greater effort and more time. When you finish the task, turn off the water and remove the garden hose. Some people also recommend unsoldering and removing the tee joint.

## Altering soil conductivity

The conductivity of the soil determines how well, or how poorly, it conducts electrical current (Table 29-1). Moist soil over a brackish water dome conducts best (southern swamps make better radio station locations), and the sand of the western deserts makes the worst conductor. Figure 29-3 shows a method for reducing the soil electrical resistance by treating with one of two chemicals: copper sulfate or common rock salt. The rock salt is one of several salt materials used for snow and ice melting in snow-prone areas of the country. If you cannot locate rock salt in a hardware store, then look for a store that sells ice-cream-making supplies. Rock salt is a principal ingredient in the process (but not the product).

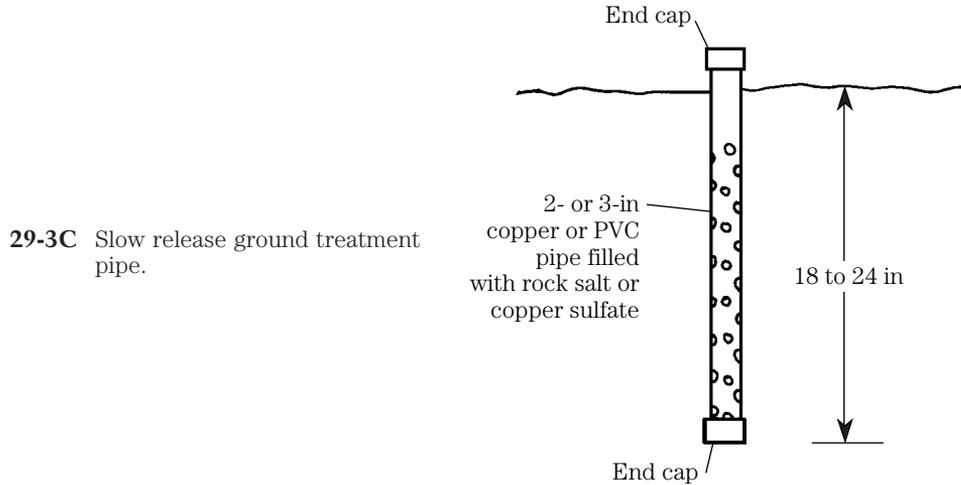
**Table 29-1. Sample soil conductivity values**

Type of soil	Dielectric constant	Conductivity (siemens /meter)	Relative quality
Salt water	81	5	Best
Fresh water	80	0.001	Very poor
Pastoral hills	14-20	0.03-0.01	Very good
Marshy, wooded	12	0.0075	Average/poor
Sandy	10	0.002	Poor
Cities	3-5	0.001	Very poor



Figures 29-3A and 29-3B show side and top views (respectively) of a slit trench method of applying the chemical treatment. Dig a 6- to 10-in-deep trench, about 12 to 24 in from the ground rod. Fill that trench with a 4- or 5-in layer of rock salt or copper sulfate. Cover the remaining depth with soil removed when you dug the trench. Water the trench well for about 15 minutes. The treatment will need to be repeated every 12 to 36 months, depending upon local rainfall and soil composition.

Figure 29-3C shows an alternative method: the salt pipe. Use either copper or PVC plumbing pipe, up to 4 in diameter (although 2 to 3 in is easier to work). The overall length of the pipe should be at least 18 to 24 in, although longer is useful.



Drill a large number of small holes in the pipe (no hole over  $\frac{3}{8}$  in), sparing only the end that will be above the surface. Fill with rock salt or copper sulfate and cap off both ends.

Install several salt pipes around the ground rod. Installation is best accomplished with a fence posthole digger. Drop the salt pipe into the hole and backfill with water. Remove the top endcap and hose down for about 15 minutes. Refill the pipe occasionally to account for the salt leaching out.

## Multiple ground rods

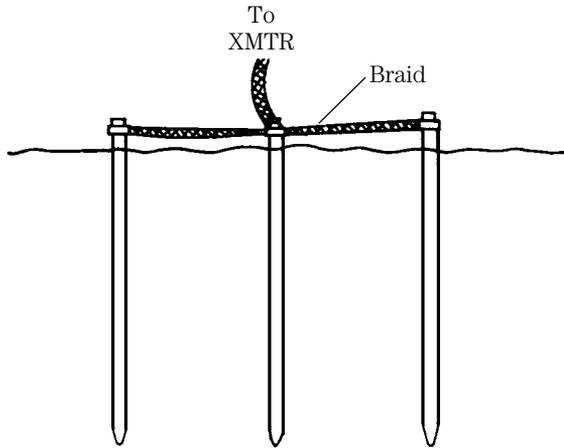
The key to a low-resistance ground is the surface area in contact with soil. One means for gaining surface area, and thereby reducing ground resistance, is to use multiple ground rods. Although multiple 4-ft rods are better than a single 4-ft ground rod, the use of 6- or 8-ft rods is recommended even in cases where multiple rods are used.

Figure 29-4A shows the use of three ground rods in the same system. The 8-ft rods are placed 12 to 18 in apart for low- and medium-power levels, and perhaps 30 in for higher amateur power levels. An electrical connection is made between the rods on the surface using either copper stripping or copper braid. The connections are sweat-soldered in the manner described above, with the feedpoint at the center rod.

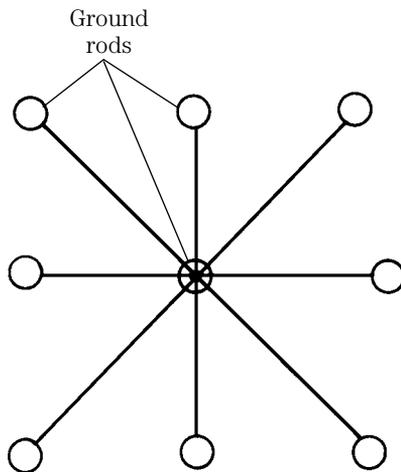
Figure 29-4B shows a more complex system with a larger number of rods laid out in an array. The rods are arrayed about 24 in apart. As in the previous case, the rods are electrically connected (by braid or foil) on the surface. Again, connection to the system is made from the rod at the geometric center of the array.

## Radials/counterpoise grounds

The effectiveness of the ground system is enhanced substantially by the use of radials either above ground or buried under the surface. Figure 29-5 shows a vertical



29-4A Multiple ground rod system.

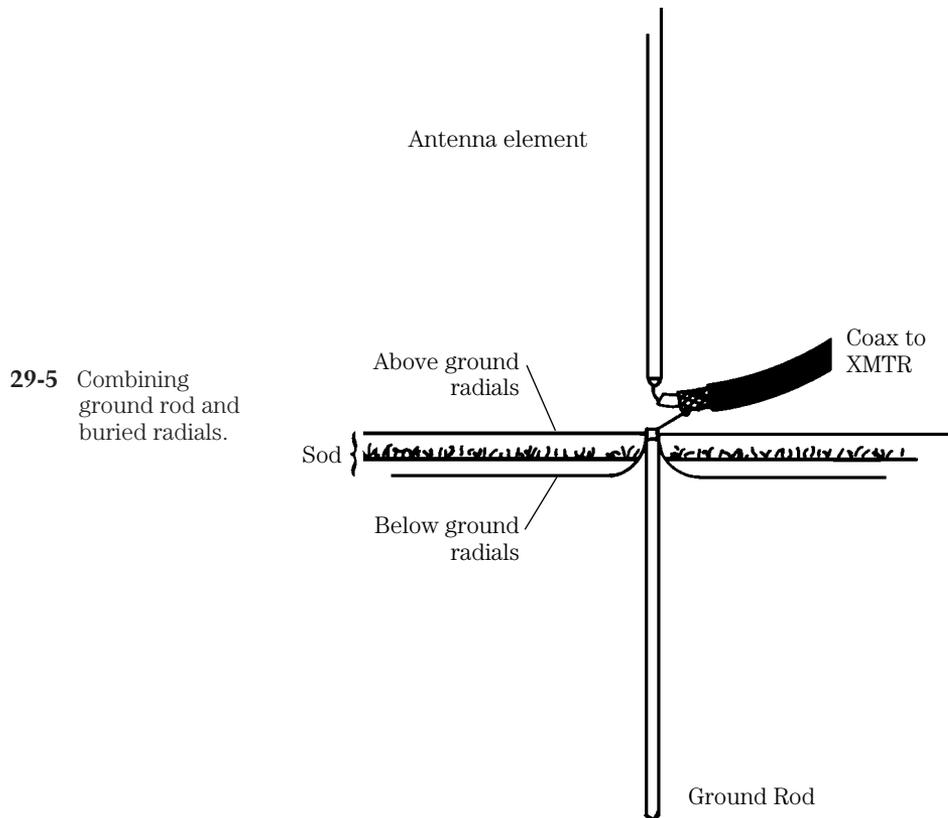


29-4B Connection scheme.

antenna with three different forms of radials: aboveground, subsurface, and ground rod. It is not unreasonable to use both radials and a ground rod. Note (from Chap. 7) that vertical antennas are relatively ineffective unless provided with a good ground system, and for most installations that requirement is best met through a system of ground radials.

An effective system of radials requires a large number of radials. Although as few as two quarter-wavelength resonant radials will provide an improvement, the best performance is to use more. Broadcasters in the AM band (550 to 1640 kHz) are advised to use 120 half-wavelength radials. Installing more than 120 radials is both expensive and time-consuming, but does not provide any substantial improvement. For amateur and small commercial stations, use a minimum of 16 quarter-wavelength radials.

Above ground, the use of insulated wire is recommended, but not required. Below ground, noninsulated wire is preferred. Although some sources claim that any

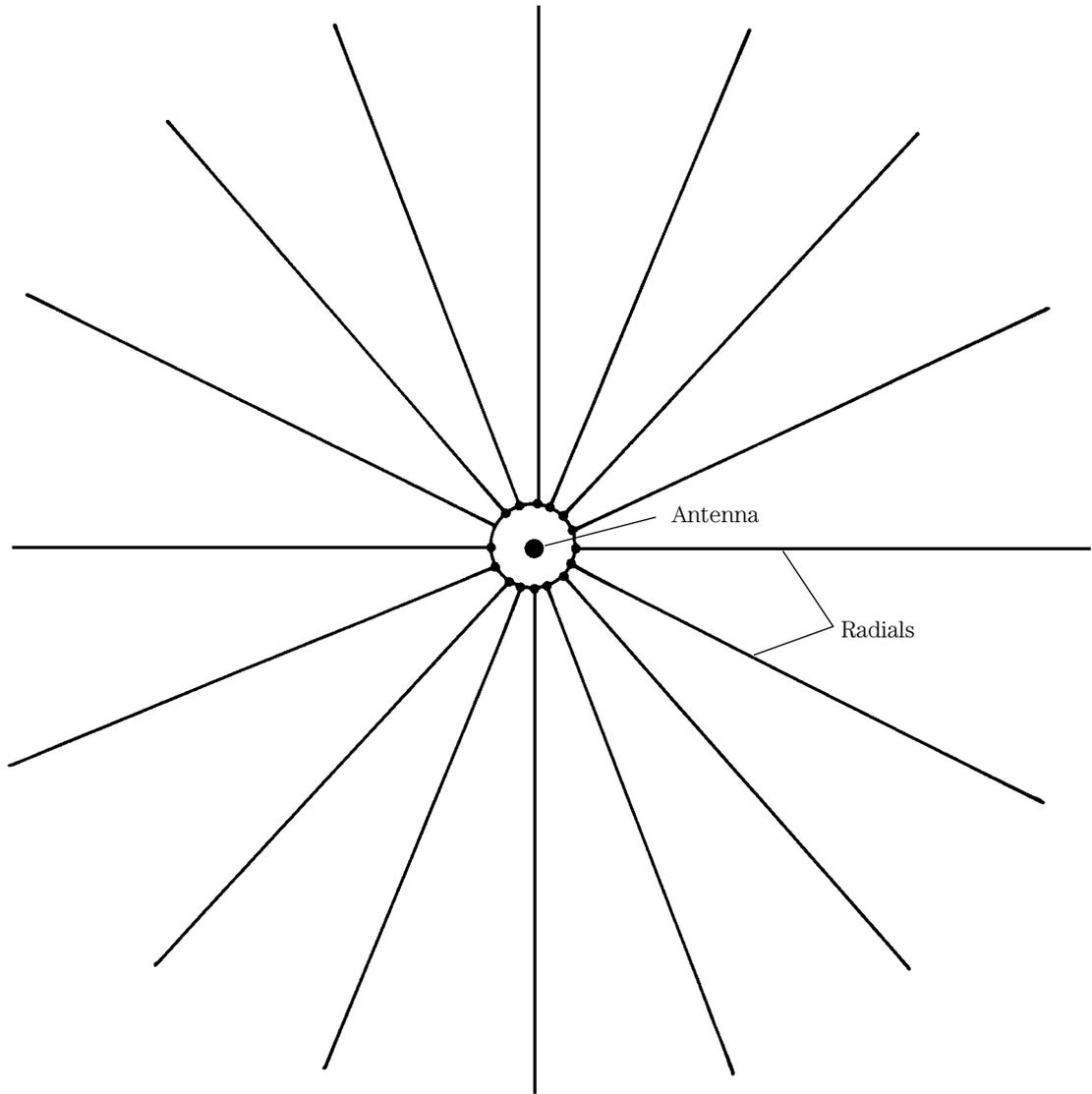


size wire from no. 26 up to no. 10 can be used, it is best to use larger sizes in that range (i.e., no. 14 through no. 10). Either solid or stranded wire can be used.

The layout for a system of radials in a vertical antenna system is depicted in a view from above in Fig. 29-6. Here, the radials are laid out in a uniform pattern around the antenna element. This coverage provides both the lowest resistance and the best radiation pattern for the antenna. Solder all radials together at a common point, which might be the ground or mounting rod used to support the vertical antenna.

## Station grounding

It does no good to provide a topflight ground system, such as those shown earlier in this chapter, if the connection between the station equipment and the ground system is substandard. Figure 29-7 shows a method used by the author to good effect. On the back of the operating position is a sheet of copper, 7 in wide, running the length of the equipment platform. This form of copper, in the 1-lb/sq ft weight, is used on older houses for roofing flashing. Each piece of equipment is connected to the ground sheet through a short length of braid harvested from RG-8 or RG-11 coaxial cable shield. RF accessories, such as the low-pass TV interference filter (if used—it should be) are mounted directly to the copper sheet. In one installation, the

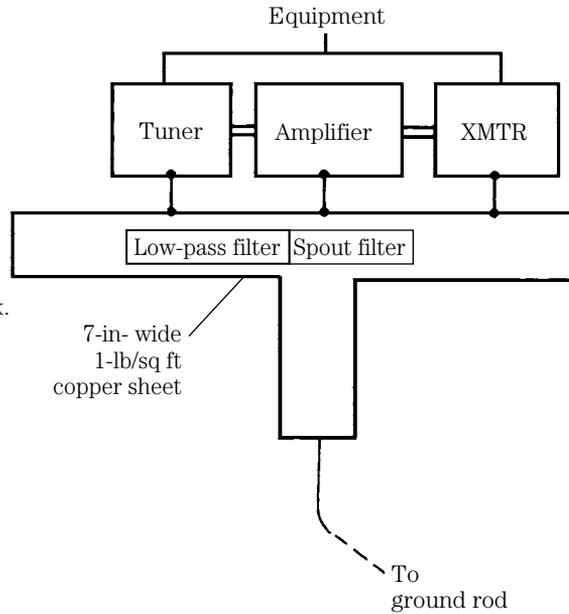


29-6 Radial pattern.

author was able to drop the copper sheet down from the table to connect directly to the ground system outside the building. The run was less than 40 in. But in other cases, a short length of braid wire will be more practical.

## Tuning the ground wire

An alternative that some operators use is the ground wire tuner. These instruments insert an inductor or capacitor (or a *LC* network) in series with the ground line. You



29-7 Ground system inside shack.

adjust the ground line tuner for maximum ground current at the operating frequency. MFJ Electronics, Inc., Mississippi State, MS, makes one of these devices.

## Conclusion

A high-quality, low-resistance ground might seem costly to install, but in reality it pays rich dividends in the functioning of your antenna. Don't overlook the quality of the ground, or you might be in the position of being penny wise and pound foolish.

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# Index

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*Note:* boldface numbers indicate illustrations; italic *t* indicates a table.

## A

- absorbed power ( $P_a$ ), transmission lines, 113–114
- absorption of microwaves, **38**
- Adcock antennas, 446–447, **448**, **449**
- Adcock, F., 446
- adjustable plunger or short-circuit termination, microwave waveguides and antennas, 388, **388**
- adjustment techniques (*See* measurement and adjustment techniques)
- admittance, 100, 105, 112–114
- air core frame loops (“box” loops), 303–305, **304**
- air-articulated line, 62–63, **63**
- airborne radio, 30
- Alexanderson alternators, 2
- aluminum vs. copper radiators, in high-frequency dipoles and doublets, 154
- AM radio, 47, 55
  - Beverage or “wave” antenna in, 224–228, **226**, **228**, **229**
  - vertically polarized HF antennas in, 186
  - radio direction finding (RDF) antennas in, 439
- AM/FM broadcast antenna tower, **4**
- amateur radio, 2, 3, 29, 49, 55
  - loop antennas, small receiver-type in, 302
  - multiband and tunable-wire antennas in, 203
  - vertically polarized HF antennas in, 176, 179, 186
- ammeter, 530, **530**
- amplifier, for loop antennas, small receiver-type, 313–316, **316**
- amplitude modulation (AM), 2, 8, 29
- angle of incidence, 16, **16**, 22
- angle of radiation, 50, 214, 274, **274**
- angle of reflection, 16, **16**, 75
- angle of reflection coefficient, **103**, 106, 112
- angle of transmission coefficient circle, Smith charts, **103**, 106
- angled or bent dipole antenna, 236–237, **240**
- antenna matching tuner, 281, **285**
- antenna noise temperature, 417–420
- antenna tuning unit (ATU) (*See also* tuning), 463–464, **464**, **465**
  - in  $(\lambda/2)$  large loop antennas, 287
  - in high-frequency dipoles and doublets, 152, 172
  - in longwires, 214
  - in multiband and tunable-wire antennas, 210
- antenna tuning unit (ATU) (*Cont.*):
  - in vertically polarized HF antennas, 173
- antinodes, 87
- aperture and aperture (slot) coupling, waveguides and antennas, 390–393, **393**, 402–403
- apex angle, rhombic beam long-wire, 223
- approximation, in high-frequency dipoles and doublets, 142–143
- arcing, 233
- Armstrong oscillator circuit, 313, **315**
- array antennas:
  - branch feed systems for, 409, **412**
  - corporate feed systems for, 409, **412**
  - flatplate array antennas, 411, 413, **414**, **415**
  - phase shifting in, 409, **412**, **413**
  - slot array antennas, 411–413, **413**
  - solid-state array antennas in, 408–411
  - Watson-Watt Adcock array antennas, 447–450, **449**, **450**
- astronomy (*See* radio astronomy antennas)
- atmospheric layers and radio propagation, 14–16, **15**, 18, 40–43, **42**
- atmospheric pressure vs. radio propagation, 25

attenuation, 9–10, 28, 32, 36, 82, 377  
 attic-installed antennas, 232–234, **233**  
 auroral propagation, 54  
 automatic load control (ALC), 515  
 azimuth aspect, high-frequency dipoles and doublets, 147

## B

backlobes in microwave waveguides and antennas, 400, **400**, **401**  
 backscatter, 52  
 balanced transmission line, in multiband and tunable-wire antennas, 208–209, **209**  
*(See also baluns)*  
 baluns:  
 coaxial cable, 465, **466**  
 in directional phased vertical antennas, 249  
 in high-frequency dipoles and doublets, 155, 171, **171**  
 in longwires, 171, **171**  
 multi-impedance, 476, **477**  
 in multiband and tunable-wire antennas, 208–209, **209**, 210  
 strain relief for, 553–554, **554**  
 tapped, 476, **477**  
 in VHF/UHF transmitting/receiving antennas, 341–346, **342**, **343**  
 bandwidth, 2  
 in loop antennas, small transmitter-type, 320–321, 320  
 in VHF/UHF transmitting/receiving antennas, 339  
 basic radio system, 515, **516**  
 battery chargers/battery packs, 126, 493–494, **494**  
 beam antennas:  
 cubical quad beam antennas, 263–267, **264**, **266**, **267**  
 inverted bobtail curtain (Thorne array) antenna, 267–269, **268**, **269**  
 quad beam VHF/UHF antenna, 356, 358, **359**  
 VHF/UHF transmitting/receiving antennas, 349–352, **351**, **352**  
 Yagi antennas, 257–262, **259**, 265, 349–352, **351**, **352**

beam antennas (*Cont.*):  
 ZL-special beam antenna, 20-m, 262–263, **262**  
 beamwidth:  
 microwave waveguides and antennas, 400, **400**, **401**  
 parabolic “dish” antennas, 407  
 bends in microwave waveguides and antennas, 388–390, **389**, **390**  
 bent dipole antenna, 236–237, **240**  
 Beverage or “wave” antenna, 224–228, **226**, **228**, **229**  
 grounding of, 227–228, **228**  
 installation of, 228  
 insulators in, 228  
 maximum effective length (MEL) in, 227  
 steerable notch Beverage antennas, 229, **229**  
 Beverage, H.H., 225  
 Binns, Jack, first maritime radio distress call, 1–2  
 bidirectionality of loopsticks in RDF antennas, 441  
 bisquare loop large antenna, 296–297, **297**  
 boundary conditions in microwave waveguides and antennas, 375–376  
 “bowtie” dipoles, 154, 155, **156**, **361**, 362  
 “box” loops, 303–305  
 branch feed systems for array antennas, 409, **412**  
 bridges:  
 impedance bridges, 519–521, **520**, **521**  
 noise bridge, 521–522, **521**, **522**, **523**  
 RF noise bridge, 522–527, **521**  
 RF power measurement using, 532, **532**  
 British Broadcasting Corporation (BBC), 54  
 broadband impedance matching, in high-frequency dipoles and doublets, 146  
 broadbanded dipoles, 154–157  
 broadside radiation pattern in directional phased vertical antennas, 246  
 Brooke, E.F., 209

burn or shock hazard through antennas, 191, 195, 215, 233, 276, 279  
 Byrne, John, 209

## C

cable TV, 67  
 cage dipoles, 154, 155, 157, **157**  
 call signs, 3  
 capacitance, 126, 138, 526–527  
 coaxial cable, 69  
 loop antennas, small receiver-type, 317  
 loop antennas, small transmitter-type, 321  
 transmission lines, 64  
 capacitance tuning of loop antennas, small receiver-type, 307–309  
 capacitive (probe) coupling in microwave waveguides and antennas, 390–393, **392**  
 capacitive antennas, 126  
 capacitive reactance, 140  
 in high-frequency dipoles and doublets, 143  
 impedance matching and, 458  
 using Smith charts, 100, **102**, 117  
 capacitors, 124–126, **124**  
 in longwires, 215  
 in loop antennas, small receiver-type, 304, 307–309  
 in loop antennas, small transmitter-type, 320, 321  
 in Yagi antennas, 261  
 cardioid radiation pattern of sense antenna in RDF antennas, 441, **442**  
 Cartesian coordinate system used in modeling software, 328, **329**, **330**  
 Cassegrain feed for parabolic “dish” antennas, 408–409, **409**  
 cavity antennas, 404, **405**  
 center insulators, 548–550, **549**, **550**, **551**  
 characteristic impedance ( $Z_0$ ), 516  
 of Beverage antenna, 226  
 of longwires, 223  
 of microwave waveguides and antennas, 383–384  
 of transmission line, 64–67, 89

- charge, 124–126, **124**, 127,  
131–137, **132**, **133**
- charged solar particles vs. propaga-  
tion, 39
- choke joints in microwave wave-  
guides and antennas, 389, **390**
- circular loop transmitter antenna,  
323, **323**
- Citizens' Band (CB) radio, 29, 48,  
166, 176, 191, 341, 496
- coaxial cable, 59, 61–62, **61**, **62**  
baluns, 465, **466**  
capacitance in, 69  
characteristic impedance of, 66  
coaxial vertical scanner antenna  
using, 365–368, **367**, **368**  
cutoff frequency ( $F_c$ ) of, 69  
doping out, 437–438  
Faraday connection in, 318  
Faraday shield in, 318  
length of, calculating, 438  
loop antennas made using, 318,  
**318**  
losses in, 369, 438  
matching section of, quarter-  
wavelength coax, 292  
multiband and tunable-wire  
antennas and, 208  
testing, 437–438, **437**, **438**  
velocity factor ( $V_p$ ) of, 248  
vertical coax VHF/UHF antenna,  
346–349, **347**, **348**
- coaxial vertical antennas, 191–194,  
**192**, **193**
- coaxial vertical scanner antenna,  
365–368, **367**, **368**
- cochannel interference in loop  
antennas, small receiver-type,  
302
- coil-loaded mobile antenna,  
480–481, **480**
- collinear “Franklin” array antenna,  
165–166, **166**
- collinear vertical VHF/UHF antenna,  
349, **350**
- commercial operators, multiband  
and tunable-wire antennas, 203
- complex admittance, Smith charts,  
105
- complex conjugates in impedance  
matching, 458
- CONALRAD station frequencies,  
439
- conductance of transmission lines,  
64
- conductivity of soil, 27, 28*t*
- cone of silence, 17
- constant reactance circles, Smith  
charts, 100–105
- constant resistance circles, Smith  
charts, 97–100, **99**
- construction techniques (*See* in-  
stallation and construction  
techniques)
- constructive interference, 19
- continuous loading in low-  
frequency operation antennas,  
**503**, 504–505, **506**
- continuous wave (CW), 55
- continuous wave trains, 7, **7**
- coordinate system to denote  
dimensions in microwave wave-  
guides and antennas, 376–377,  
**377**
- coordinate system used in model-  
ing software, 328, **329**, **330**
- copper plumbing pipes as ground  
rod, 574
- copper sulfate to alter soil conduc-  
tivity, 575–576
- copperweld wire, 214
- core material of transformers, 474
- corner reflector antennas, 404,  
**405**, **406**
- corona effects, 233
- corporate feed systems for array  
antennas, 409, **412**
- cosmic radiation noise, 37
- counterpoise grounds, 577–579,  
**579**  
in emergency operations antennas,  
487–488, **488**  
in longwire, 170–172, **171**
- coupling loop antennas, 306–307,  
**306**
- coupling methods for microwave  
waveguides and antennas,  
390–393
- CQD/CQE distress calls, 2
- critical angle, 48
- critical frequency ( $F_o$ ), 43, **44**
- cross-polarization, 13
- cubical quad beam antennas,  
263–267, **264**, **266**, **267**, 421
- current, 123, 126, 128, 129,  
131–137, **132**, **133**
- current (*Cont.*):  
in directional phased vertical  
antennas, 248  
in ( $\lambda/2$  large loop antennas, 289  
in high-frequency dipoles and  
doublets, 143–146, **143**  
in loop antennas, large designs,  
299  
in loop antennas, small receiver-  
type, 299  
in loop antennas, small transmitter-  
type, 320  
standing waves of, 134, **135**, 136,  
**137**, 139  
in vertically polarized HF antennas,  
173, **174**  
in loop antennas, small transmitter-  
type, 323
- current standing wave ratio  
(ISWR), 88
- current transformer wattmeter,  
534–538, **535**, **536**, **537**
- cutoff frequency ( $F_c$ ):  
in microwave waveguides and  
antennas, 381–383  
in parabolic “dish” antennas, 406
- cycles per second, 7
- ## D
- D layer of ionosphere, 41, **42**, 46,  
51
- D loop ( $1\lambda$ ) large antenna,  
290–293, **291**
- damping, 6
- dead zones, 23
- Dellinger fade, 46–47
- delta feed match systems:  
in low-frequency operation  
antennas, 507, **509**  
in VHF/UHF transmitting/receiving  
antennas, 343, **344**, 345
- delta loop ( $1\lambda$ ) large antenna,  
290–293, **291**, **295**
- delta-fed grounded vertical  
antenna, 200–201, **200**
- demiquad loop large antenna, 293,  
**294**
- depth wound loop antennas, small  
receiver-type, 303
- design equations for loop antennas,  
small transmitter-type,  
321–322
- destructive interference, 19

- diamond loop (1 $\lambda$ ) large antenna, 290–293, **291**
- dielectric constants, 138
- dielectric loss, 369
- dielectric properties of air, 29–30, **30**
- diffraction, 16–17, **17**, 32–35
- dip oscillator/dip meter, 527–529, **527, 528**
- dipoles and doublets, 141–172, **142**
- aluminum vs. copper radiators in, 154
  - angled or bent dipole, 236–237, **240**
  - antenna tuning unit (ATU) for, 152, 172
  - approximation of, 142–143
  - attic installation of, 232–234, **233**
  - azimuth aspect of, 147
  - baluns in, 155, 171, **171**
  - “bowtie” dipoles in, 154, 155, **156, 361, 362**
  - broadband impedance matching in, 146
  - broadbanded dipoles in, 154–157
  - cage dipoles in, 154, 155, 157, **157**
  - capacitive reactance in, 143
  - center insulators, 548–550, **549, 550, 551**
  - collinear “Franklin” array antenna in, 165–166, **166**
  - counterpoise longwire in, 170–172, **171**
  - current in, 143–146, **143**
  - directivity in, 146–247
  - double dipoles in, 160
  - double extended Zepp antenna in, 164–165, **165**
  - drooping dipole scanner antenna in, 363, **363**
  - electrical vs. physical length of, 141, 142–143, 149–151, 160, **162**
  - emergency operations antennas, 490, **490**
  - end or egg insulators, 551–553, **553**
  - feedpoint in, 143–146, 256–257, **257**
  - floor flanges mounting hardware in, 256
  - folded dipole, 154–155, **156, 362, 362, 554–556, 555, 556**
- dipoles and doublets (*Cont.*):
- gain in, 146–147
  - ground plane verticals in, 172
  - height of antenna vs. radiation patterns in, 148–149, **148**
  - height of antenna vs. resistance in, 145–146, **145**
  - hidden and limited-space antennas, 232–234, **233**
  - high-Q models of, 160
  - high-frequency dipoles and doublets, 141–172, **142**
  - impedance and impedance matching in, 143–146, 149, 151–152, 155, 237
  - inductive reactance in, 143, 157, 158*t*, 159–160, **161**
  - installation plan for, **552**
  - insulators for, 141, 256, 548–550, **549, 550, 551, 551–553, 553**
  - inverted-vee dipole in, 152–153, **153, 490–491, 491**
  - isotropic radiation in, 146
  - length of, 141, 142–143, 149–151, 160, 232, 255–257
  - limited-space designs for, 236–244
  - loading coils in, 157–160, **157**
  - losses in, 143–146
  - microwave waveguides and antennas and, elements of, 396–398, **397, 399**
  - mounting of, 141, 233, 256
  - multiband fan dipole in, 169–170, **170**
  - off-center-fed full-wave doublet (OCFD) in, 163, **163**
  - off-center-fed nonresonant sloper (OCFS), 164, **164**
  - ohmic loss in, 143–146
  - omnidirectionality of, 146
  - phase reversal stubs in, 165
  - Q-factor in, 160
  - Q-sections in, 146
  - radiation patterns of, 146–152, **147**
  - radiation resistance in, 143–146
  - radiators of, 141
  - reactance in, 143
  - resistance in, 143–146, **145**
  - resonance in, 143, 149–151, **150, 237**
  - rhombic inverted-vee antenna in, 169, **169**
- dipoles and doublets (*Cont.*):
- right-angle flange mounting hardware in, 256
  - RLC networks and, 143
  - rotatable dipole antennas, 239, **242, 255–257, 257**
  - shortened coil-loaded dipoles in, 157–160, **158**
  - shortwave reception antennas, 272, **273, 274**
  - sloping (“sloper” or “slipole”) dipole in, 153–154, **154, 155, 236, 238**
  - space diversity in, 160
  - stacked dipoles in, 160–163, **162**
  - stagger tuning of, 157
  - standing waves and standing wave ratio (SWR) in, 144
  - strain relief in, 553–554, **554**
  - support ropes for, 551
  - three-dimensional (ideal) radiation pattern of, 147–148, **147**
  - tilted, center-fed, terminated, folded dipole (TCFTFD or T + 2 + FD, TTFD) in, 166–168
  - transformers for, 146
  - tunable dipoles in, 160
  - tuning of, 149–152
  - vee dipoles, 236, **239**
  - vee-sloper antenna in, 168–169, **168**
  - vertical dipole scanner antenna in, 364, **364, 365**
  - vertical plane radiation pattern of, 148–149, **148**
  - voltage in, 143–146, **143**
  - voltage standing wave ratio (VSWR) in, 144, 146, 149–151, **150, 160**
  - wavelength measurement of, 141–142
  - wire antenna, basic design, 548, **548**
  - Zepp antenna in, 164–165, **165, 171**
- direct vs. reflected space waves, 17–19, **18**
- direction of radio waves, 54
- directional array, 360-degree
- phased, 250–254, **253, 254*t***
- directional beam antennas, 255–269

- directional beam antennas (*Cont.*):  
 cubical quad beam antennas,  
 263–267, **264**, **266**, **267**  
 directivity of, 255  
 effective radiated power (ERP) in,  
 255  
 electromagnetic interference  
 (EMI), 263  
 feedpoints, 256–257, **257**  
 260–261, **261**, 263, **265**  
 floor flanges mounting hardware  
 in, 256  
 gamma match feed system for,  
 260–261, **261**  
 insulators, 256  
 inverted bobtail curtain (Thorne  
 array) antenna, 267–269, **268**,  
**269**  
 length of, 255–257, 259–260  
 mounting of, 256  
 quad loop antennas in, 263–267,  
**264**, **266**, **267**  
 right-angle flange mounting hard-  
 ware in, 256  
 rotatable dipole as, 255–257, **257**  
 Yagi antennas as, 257–262, **259**,  
 265  
 Yagi-Uda antenna as, 258  
 ZL-special beam antenna, 20-m,  
 262–263, **262**
- directional discontinuity ring radia-  
 tor (DDRR) antenna, 424–427,  
**426**, 427*t*
- directional phased vertical antennas,  
 245–254  
 broadside radiation pattern  
 in, 246  
 current in, 248  
 directivity of, 246–250  
 effective radiated power (ERP) in,  
 245–246  
 feed phasing table for, 254*t*  
 feedpoint in, 247–248, **247**, **248**  
 length of, variable lengths in,  
 248–249, **249**  
 noise in, 245  
 omnidirectionality of, 245  
 phase-shifting circuit for, **252**  
 phasing in, 246–250  
 phasing switch in, 249–250, **252**  
 radiation patterns in, 246–250,  
**246**, **247**  
 signal-to-noise ratio (SNR) in, 246
- directional phased vertical  
 antennas (*Cont.*):  
 360-degree directional array, 250,  
**253**, 254*t*  
 three-element, 250, **253**  
 transformers in, 249–250, **250**,  
**251**  
 two-port phasing transformer for,  
 249–250, **250**, **251**  
 velocity factor ( $V_p$ ) of coax in, 248  
 directional wire antenna, for short-  
 wave reception, 281–286, **284**
- directivity:  
 in directional beam antennas, 255  
 in directional phased vertical  
 antennas, 246–250  
 in high-frequency dipoles and  
 doublets, 146–247  
 in microwave waveguides and  
 antennas, 395, 397, 398–403,  
**400**  
 in shortwave reception antennas,  
 272  
 in VHF/UHF transmitting/receiving  
 antennas, 349
- directivity gain ( $G_d$ ) in microwave  
 waveguides and antennas,  
 401–402
- directors, Yagi antennas, 258–259
- directrix, parabolic “dish” antennas,  
 404
- discrete loading in low-frequency  
 operation antennas, **503**,  
 504–505
- dish antennas (*See* parabolic “dish”  
 antennas)
- distance between antennas, 28
- distance calculation, for radio,  
 26–27, **26**
- distress calls, 2
- distributed capacitance, in loop  
 antennas, small transmitter-  
 type, 321
- disturbance, ionospheric, 45–47
- diurnal (*See* time-of-day  
 variations)
- diversity propagation/reception,  
 55–58, **57**, **58**, 160
- dominant mode in microwave  
 waveguides and antennas,  
 376–377
- Doppler antennas, 450–451, **451**
- double dipoles, 160
- double ducky direction finders  
 (DDDF), 455–456, **455**
- double extended Zepp antenna,  
 164–165, **165**
- double vee longwire antennas, 223
- double-shielded coaxial, 63, **63**
- doublet antennas (*See* high-  
 frequency dipoles and doublets)
- drooping dipole scanner antenna,  
 363, **363**
- drooping dipole multiband  
 antennas, 205
- “drooping radials,” in vertically po-  
 larized HF antennas, 176, **177**
- ducting, 29–31, **31**, 36
- dummy loads, 385–388, **386**,  
 540–542, **540**, **541**, **542**
- DXing using loop antennas, small  
 receiver-type, 311–313
- ## E
- E layer of ionosphere, 41, **42**, 46
- effective aperture ( $A_e$ ), 11, 394,  
 407
- effective Earth’s radius factor (K  
 factor), 23–27
- effective height of antennas, 304
- effective radiated power (ERP), 45  
 in directional beam antennas,  
 255  
 in directional phased vertical  
 antennas, 245–246  
 in microwave waveguides and  
 antennas, 401  
 in VHF/UHF transmitting/receiving  
 antennas, 349
- efficiency, in loop antennas, small  
 transmitter-type, 321
- E-field components, 27
- electric field, 123, 124–126, **125**,  
 127, **128**  
 in longwires, 219  
 in microwave waveguides and  
 antennas, 383, 395–396  
 in vertically polarized HF antennas,  
 173
- electrical installation, vertical  
 antennas, 560–561
- electrical vs. physical length,  
 137–140, **139**  
 of high-frequency dipoles and  
 doublets, 141–143, 149–151,  
 160, **162**

- electrical vs. physical length  
(*Cont.*):  
impedance matching and,  
470–471  
of longwires, 213–214, 217–219,  
221  
of transmission line, 67–68  
electromagnetic fields, 8–16, **9**,  
123, 125–127  
electromagnetic interference  
(EMI), 263  
electromagnetic radiation vs. prop-  
agation, 39  
electromagnetic waves, 5  
electrostatic fields/discharge, in  
longwires, 219  
eleven-year cycle of radio propaga-  
tion, 46  
EM Scientific, 331  
emergency operations antennas,  
485–493  
battery packs for, 493–494, **494**  
counterpoise ground plane for,  
487–488, **488**  
dipoles in, 490, **490**  
grounding and ground systems  
for, 487–488  
increasing signal strength from  
portable/mobile stations for,  
486  
inverted-vee dipole in, 490–491,  
**491**  
mobile antennas and, 486–487,  
**487**  
mounting of, on vehicles, 491–492  
multiple radials from vehicle-  
mounted antenna for, 488, **489**  
radial-to-mobile-antenna connec-  
tion detail in, 488, **489**  
repeaters and, 487  
telescoping antennas for, 491  
temporary antenna construction  
for, 487–488, **488**  
temporary mast mount for, 492,  
**493**  
VHF band for, 487  
end effect, 138–139  
end or egg insulators, 551–553, **553**  
end plates/end terminators, in  
microwave waveguides and  
antennas, 385, 388, **388**  
end-fed longwire, 217–219, **217**  
end-fed Zepp antenna, 208, **208**  
environmental/natural noise  
(QRN), 245  
erecting antennas safely, 543–548  
escape angle, 48  
ether, 8  
Everett, W.L., 209  
Expert MiniNEC modeling soft-  
ware, 331  
EZNEC for Windows modeling soft-  
ware, 331, 334–337, **335, 336,**  
**337**  
EZNEC Professional modeling soft-  
ware, 329
- F**  
F layer of ionosphere, 41–43, **42, 46**  
fading, 35–36, 35, 55–58, 55  
Dellinger fade, 46–47  
selective fading, 55  
fan dipole, multiband, 169–170,  
**170**  
far field calculations in microwave  
waveguides and antennas,  
395–396  
Faraday connection, 318  
Faraday shield, 318  
Federal Communications Commis-  
sion (FCC), 2  
feed geometries for parabolic “dish”  
antennas, 408–409, **408, 409**  
feed illumination in parabolic “dish”  
antennas, 406  
feeding transformer, 476, **477**  
feedpoints:  
 $\lambda/2$  large loop antennas, 287–288  
dipoles, 256–257, **257**  
directional beam antennas,  
256–257, **257**, 260–261, **261**  
directional phased vertical  
antennas, 247–248, **247, 248**  
high-frequency dipoles and  
doublets, 143–146  
longwires, 217–218  
multiband and tunable-wire  
antennas, 204, 210  
one/ $\lambda$  large loop antennas, 292  
quad beam antennas, 263, **265**  
vertically polarized HF antennas,  
197, **198**  
Yagi antennas, 260–261, **261**  
fence posts as supports for vertical  
antennas, 561–565, **563, 564**  
ferrite core inductors, 471–477  
ferrite rods, 475, **475**  
fiber optics, 370  
field strength meters (FSM), for  
mobile HF antennas, 483–485,  
**484, 485**, 485*t*  
field strength, 21, **22**  
figure of merit, 418  
filters, 55, 515  
five-eighths-wavelength antennas,  
197–200  
2-meter VHF/UHF antenna, 353,  
**354**  
vertical antenna, 181, **182**  
flagpole-installed antennas,  
234–236, **235, 237**  
flatplate array antennas, 411, 413,  
**414, 415**  
flattop antennas, shortwave recep-  
tion, 279, **280**  
floor flanges mounting hardware,  
256  
flux, magnetic, 126, 127  
FM, marine radio antennas, 495  
fold- or tilt-over tower, 567, **567**  
folded dipole, 154–155, **156**  
installation of, 554–556, **555, 556**  
tilted, center-fed, terminated,  
folded dipole (TCFTFD or  
 $T^2$ FD, TTFD) in, 166–168,  
**167**  
VHF/UHF antennas, 362, **362**  
forward power, 538, **539**  
forward scatter, 52  
“fox hunting” antenna, 444, **445**  
“Franklin” array antenna, 165–166,  
**166**  
free space, 14, 19, **21**, 67–68,  
137–138  
free-space velocity (C) in micro-  
wave waveguides and antennas,  
377, 380  
frequency, 2, 7, 9  
in microwave waveguides and  
antennas, 373, **374**, 375, **376**,  
380, 383–384, **384**  
using Smith charts, 117–121, **119,**  
**120**  
tilt angle vs., 28, 28*t*  
VHF/UHF transmitting/receiving  
antennas, 339, 341  
VSWR vs., 434–437, **435**  
frequency diversity, 36, 55, **57**  
frequency modulation (FM), 32

- frequency of optimum traffic (FOT), 45
- Fresnel zones, 33–34, **34**, 35
- fundamentals of radio antennas, 123–140
- ## G
- G5RV multiband dipole antenna, 206–208, **207**
- gain:
- in  $\lambda/2$  large loop antennas, 288–289, **289**
  - in high-frequency dipoles and doublets, 146–147
  - in longwires, 214, 218, **218**, 223
  - in loop antennas, small transmitter-type, 319–320
  - in microwave waveguides and antennas, 394–395, 398–403
  - in microwave waveguides and antennas, vs. aperture in, 402–403
  - in parabolic “dish” antennas, 405–408
  - in shortwave reception antennas, 272
  - in vertically polarized HF antennas, 179–181
  - in VHF/UHF transmitting/receiving antennas, 349
- galactic noise, 39, 421
- gamma match feed system:
- in Yagi antennas, 260–261, **261**
  - in VHF/UHF transmitting/receiving antennas, 346, **346**
- gamma rays, 39
- gamma-fed grounded vertical antenna, 200–201, **201**
- gas-filled line, 62, **63**
- ghost radiator, vertically polarized HF antennas, 173
- gin pole to raise towers, 570, **571**
- Global Positioning System (GPS), 443
- GNEC modeling software, 329–330
- graphite-sand dummy load in microwave waveguides and antennas, 385
- gray line transmission, 51, **53**
- great circle paths, 50–51, **52**
- grid dip meter, 527–529, **528**
- ground losses, vertically polarized HF antennas, 183
- ground plane vertical antennas, 172, 176, **177**
- ground rods, 573–574, **574**, 575, **575**, 577, **578**
- ground stakes and guy wires, 570, **572**
- ground wave, 17–29, **18**, 47–48, **48**, 302
- ground-mounted vertical antennas, 561–565, **562**
- grounded vertical antennas, 200–201
- grounding and ground systems, 573–581
- in Beverage antenna, 227–228, **228**
  - copper plumbing pipes as ground rod, 574
  - copper sulfate to alter soil conductivity in, 575–576
  - counterpoise grounds, 577–579, **579**
  - in emergency operations antennas, 487–488
  - ground rods for, 573–574, **574**, 574–575, **575**, 577, **578**
  - in hidden and limited-space antennas, 238–239
  - impedance matching and, 459–460
  - interference and, 573
  - length of ground wires in, 574
  - in longwires, 215–216, **216**
  - in low-frequency operation antennas, 501–502, **502**
  - in marine radio antennas, 495–496, **495**, **496**
  - radials for, 577–579, **579**, **580**
  - “salt pipe” to apply chemical soil-conductivity improvers, 576–577, **577**
  - salt to alter soil conductivity in, 575–576
  - in shortwave reception antennas, 276, **277**, 278–279, **279**
  - soil conductivity and, altering, 575–577, 576*t*, **576**, **577**
  - station grounding, 579–580, **581**
  - trenching to apply chemical soil-conductivity improvers, 576, **576**
- grounding and ground systems (Cont.):
- tuning ground wire, 580–581
  - in vertically polarized HF antennas, 176, 185–187, 195–197, **196**
- groundplane VHF/UHF antenna, 356, **357**
- groundplane scanner antenna, 364, **366**
- group velocity ( $V_g$ ) in microwave waveguides and antennas, 377–379
- Grover’s equation, for loop antennas, small receiver-type, 302–303, 303*t*
- guy wires and ground stakes, 570, **572**
- ## H
- “hairpin” loading scheme, 505, **506**
- half-delta sloper (HDS), 295–296, **296**
- half-slope vertical antenna, 238, **240**
- half-wave antennas, 126, 181, **182**
- current distribution in, 131–137, **132**, **133**
  - electric vs. physical length of antenna, 137–139, **139**
  - impedance matching in, 139–140, **139**
  - standing waves and, 129
  - voltage distribution in, 131–137, **132**, **133**
- half-wavelength line, 83
- halo VHF/UHF antenna, 356, **358**
- hardline, 62, **62**
- harmonics, 140, 209, 210, 515
- height of antenna, 19, **20**, 25, 28
- distance calculation for, 26–27, **26**
  - in high-frequency dipoles and doublets, vs. radiation patterns in, 148–149, **148**
  - in high-frequency dipoles and doublets, vs. resistance in, 145–146, **145**
  - in loop antennas, small receiver-type, 304
  - in vertically polarized HF antennas, 181, 187
- helical antennas, 427–428, **429**
- helical line, 62

helical whip mobile antenna, 481, **482**

helically wound antennas, **239**, 242, **243**

hertz as unit of frequency, 7

Hertz or hertzian antenna (*See* high-frequency dipoles and doublets)

Hertz, Heinrich, 5, 7

heterodyning, 55

hidden and limited-space antennas, 231–244

  angled or bent dipole, 236–237, **240**

  attic-installed, 232–234, **233**

  dipoles as, 232–234, **233**, 236

  flagpole installation for, 234–236, **235**, **237**

  ground system for, 238–239

  half-slope vertical antenna as, 238, **240**

  helically wound antennas as, **239**, 242, **243**

  L-section couplers in, 238–239

  limited-space designs for, 236–244

  longwires as, 232

  mobile antenna used as windowsill-mounted, 239, **241**

  mounting of, 233

  nonresonant loop antennas as, 234–235, **234**

  rotatable dipole antennas as, 239, **242**

  safety issues for, 233

  sloping (“sloper” or “slipole”) dipole as, 236, **238**

  vee dipoles as, 236, **239**

  voltage standing wave ratio (VSWR) in, 238

  window-sill antennas as, 238, **241**

high frequency (HF), 3, 13, 17, 29, 35, 37, 41, 48, 61, 170–171, 203, 479–485

high-frequency dipoles and doublets (*See* dipoles and doublets)

high-power RF wattmeters, 538, **538**

high-Q dipoles and doublets, 160

hinged baseplate tower, 566–567, **538**

history of radio broadcasting and communications, 1–4

hopping, 36

horizontal low-frequency operation antennas, 505

horizontal polarization, 12–13, **13**, 27, 55, 173

  Beverage antenna and, 226–227

  helically wound antennas and, **239**, 242, **243**

horn antenna radiators, 403–404, **403**

Huygens–Fresnel wave theory, 32

**I**

$I^2R$  losses in microwave waveguides and antennas, 385

image antenna, vertically polarized HF antennas, 173

imaginary operator ( $j$ ), 22

impedance and impedance matching, 140, 433, 457–477, 525

  antenna tuners for, 463–464, **464**, **465**

  in antennas, 139, **139**

  approaches to, 458–462, **458**

  in Beverage antenna, 226

  capacitive reactance in, 458

  characteristic impedance ( $Z_0$ ) of transmission line and, 64–67

  in coaxial cable baluns, 465, **466**

  complex conjugates in, 458

  in dipoles, 237

  electrical vs. physical length calculations in, 470–471

  feeding transformer for, 476, **477**

  ferrite core inductors in, 471–477

  ferrite rods in, 475, **475**

  grounding and, 459–460

  in high-frequency dipoles and doublets, 143–146, 149, 151–152, 155

  impedance bridges in, 519–521, **520**, **521**

  impedance matching network (Q sections) in, 85

  inductive reactance in, 458

  L-section network for, 458–460, **459**

  in  $\lambda/2$  large loop antennas, 287–288

  in longwires, 214, 223

  in low-frequency operation antennas, 505

impedance and impedance matching (*Cont.*):

  matching stub systems for, 465–466, **466**, **467**

  matching units for, 515

  in microwave waveguides and antennas, 383–384, **384**, 394, 396

  multi-impedance balun for, 476, **477**

  in multiband and tunable-wire antennas, 210

  noise and, 417–418

  normalization of impedance in, 469–470

  in  $1\lambda$  large loop antennas, 292–293, **292**

  pi networks for, 461, **461**

  quarter-wave matching (Q section) sections for, 467–468, **468**, **469**

  radials in, 460

  series matching section for, 468–471

  using Smith charts, 110–112, **111**, 115, 119–121, **119**

  using Smith charts, impedance radius in, 100

  using Smith charts, normalized impedance line in, 95–97, **98**

  SPC transmatch network in, 463, **463**

  split-capacitor networks for, 461–462, **462**

  stub matching systems for, 115

  tapped balun for, 476, **477**

  tee-network transmatch for, 463, **463**

  toroid broadbanded transformers for, 475–477, **476**

  toroid cores and, 471, **472**

  toroidal transformers and, 471–475, **472**

  in towers, 507, **508**

  transmatch circuits for, 462–471, **462**

  in transmission lines, 71, 72, 82–86, 109, 110–112, **111**, 112

  in vertically polarized HF antennas, 173, 175–176, 183, **184**, **186**, 195–197, 199–200, **199**

  in VHF/UHF transmitting/receiving antennas, 341–346, **342**

- impedance and impedance matching (*Cont.*):  
 Wheatstone bridge in, 519–521, **520**  
 in Yagi antennas, 260–261, **261**
- impedance bridges, 519–521, **520**, **521**
- incident angle, 48
- incident power ( $P_i$ ) of transmission lines, 89
- incident waves, 128, 134
- index of refraction, 23–24
- inductance, 526–527  
 ferrite core inductors in, 471–477  
 Grover's equation in, 302–303, 303*t*
- in longwires, 214
- in loop antennas, small receiver-type, 307
- in loop antennas, small transmitter-type, 321
- in transmission lines, 64
- induction fields, 5, 123, 127
- inductive (loop) coupling in  
 microwave waveguides and antennas, 390–393, **393**
- inductive loading in low-frequency operation antennas, **503**, 504–505
- inductive reactance, 140  
 in high-frequency dipoles and doublets, 143, 157, 158*t*, 159–160, **161**
- impedance matching and, 458
- in  $\lambda/2$  large loop antennas, 288–289, **289**
- in loop antennas, small transmitter-type, 319, 321  
 using Smith charts, 100, **101**, 117
- infrared (IR) radiation, 9
- installation and construction techniques, 433–438, 543–572  
 center insulators, 548–550, **549**, **550**, **551**  
 electrical installation, vertical antennas, 560–561  
 end or egg insulators, 551–553, **553**  
 fence posts as supports for vertical antennas, 561–565, **563**, **564**  
 fold- or tilt-over tower, 567, **567**
- installation and construction techniques (*Cont.*):  
 folded dipole installation, 554–556, **555**, **556**  
 four/4x4 support posts for, 565, **565**  
 gin pole to raise towers, 570, **571**  
 ground-mounted vertical antennas, 561–565, **562**  
 guy wires and ground stakes, 570, **572**  
 hinged baseplate tower, 566–567, **538**  
 insulators, 548–550, **549**, **550**, 551–553, **551**, **553**  
 mast-mounted vertical antenna, 546, 560–561, **546**, **561**, **562**  
 masts and supports, 556–559, **557**, **558**  
 pedestals for towers, 568–570, **569**  
 permits, zoning, and regulations, 547–548  
 pulley system for mounting, 557–558, **558**  
 raising towers, 570, **570**, **571**  
 safety issues, 543–548  
 slip-up tower, 567–568, **568**, **569**  
 stirrup-end mounting hardware, 558, **559**  
 strain relief, 553–554, **554**  
 support ropes, 551  
 suburban lot installation plan, **545**  
 tower installation, 565–572, **538**  
 U-bolt mounting hardware, 558, **559**  
 vertical antenna installation, 560–572  
 wire antenna, basic design, 548, **548**
- insulators (or spreaders), 548–550, **549**, **550**, **551**, 551–553, **553**  
 in Beverage antenna, 228  
 in dipoles, 256  
 in directional beam antennas, 256  
 in high-frequency dipoles and doublets, 141  
 in microwave waveguides and antennas, 371–372  
 in multiband and tunable-wire antennas, 206  
 in vertically polarized HF antennas, 175, 195
- intensity of radio waves, 9–10
- interference, 19, 36, **36**, **37**, 573  
 electromagnetic interference (EMI), 263  
 in longwires, 214  
 in loop antennas, small receiver-type, 302  
 in microwave waveguides and antennas, 400–401  
 in shortwave reception antennas, 272, **273**, 274
- interferometer antennas, 431, **431**, **432**
- intermediate frequency (IF), 55
- International Telecommunications Union (ITU), 2
- inverse square law ( $1/D^2$ ), 9–10, **11**, 369–370
- inverted bobtail curtain (Thorne array) antenna, 267–269, **268**, **269**
- inverted-L antenna, 510, **510**
- inverted-vee antenna, rhombic, 169, **169**
- inverted-vee dipole, 152–153, **153**  
 emergency operations antennas, 490–491, **491**  
 multiband and tunable-wire antennas, 205, **206**
- ionization, 5, 14–16, **15**, 18, 37, 41, 46, 47, 51, 52, 54
- ionograms, 43
- ionosphere, 14–16, **15**, 18, 37, 40–43, **42**, 176
- ionospheric disturbances, 45–47
- ionospheric propagation, 17–18, **18**, 37–43, 176
- ionospheric scatter propagation, 51–54, **53**
- ionospheric storms, 47
- ionospheric variation, 45–47
- isothermal region of atmosphere, 14, **15**
- isophase plane, 32
- isopotential line, loop antennas, small receiver-type, 303
- isoresistance circles, using Smith charts, 97–100, **99**
- isotropic antenna waveguide, 394–403  
 backlobes in, 400, **400**, **401**  
 beamwidth in, 400, **400**, **401**

isotropic antenna waveguide  
(*Cont.*):  
dipole antenna elements and,  
396–398, **397, 399**  
directivity in, 397, 398–403, **400**  
gain in, 398–403  
gain vs. aperture in, 402–403  
impedance in, 396  
interference and, 400–401  
near field and far field calculations  
in, 395–396  
ohmic loss in, 396  
radiation patterns and, 397, **398**  
radiation resistance in, 396  
sidelobes in, 400, **400, 401**  
isotropic radiation:  
in high–frequency dipoles and  
doublets, 146  
in microwave waveguides and  
antennas, 394  
in shortwave reception antennas,  
272  
in vertically polarized HF antennas,  
179–181, **181**  
isotropic sources, 10–14  
ISWR (*See* current standing  
wave ratio)

**J**

J-pole antennas, 201, 353–355,  
**355**  
Jansky, Carl, 421  
JFET amplifier for loop antennas,  
small receiver-type, 313–316,  
**316**  
Johnson noise, 417  
joints and bends in microwave  
waveguides and antennas,  
388–390, **389, 390**  
Jupiter reception antenna,  
421–422, **423**

**K**

K factor, 23–27  
kilohertz as unit of frequency, 7  
Ku bands, 69

**L**

L-section coupler, 214, **215**,  
238–239  
L-section network for impedance  
matching, 458–460, **459**  
lag, voltage, 126

$\lambda/2$  large loop antennas, 287–289,  
**288, 289**  
 $\lambda/4$  inverted-L antenna, 510, **510**  
Law of Reciprocity, 226, 271, 394  
LC tank circuit, 526, **526**  
leakage resistance, 140  
left-hand rule, 126, **127**  
length of antenna elements:  
in coaxial cable, 438  
in dipoles, 141, 142–143, 149–151,  
160, 232, 255–256  
in directional beam antennas,  
255–257, 259–260  
in directional phased vertical an-  
tennas, variable lengths in,  
248–249, **249**  
in high–frequency dipoles and  
doublets, 141, 142–143,  
149–151, 160, 232, 255–256  
in longwires, 213–215, 218–219,  
**218**, 221, 223  
in low-frequency operation  
antennas, 504  
in microwave waveguides and  
antennas, 392  
in multiband and tunable-wire  
antennas, 204  
in vertical antennas, 504  
in vertically polarized HF antennas,  
181, 183, **184**  
in VHF/UHF transmitting/receiving  
antennas, 339  
in Yagi antennas, 259–260  
Lewallen, Roy, 329, 331, 334  
licensing, 3  
light, speed, 8, 9, 13–14, 39, 525  
limited-space antennas, 231–244  
line attenuation constant, 82  
line flattener tuner for marine radio  
antennas, 500, **500**  
line-of-sight transmission, 19, **20**,  
23–24, **23**, 35, 339  
linear loading in low-frequency  
operation antennas, **504–505**  
linearly loaded tee antenna, 511,  
**512**  
link-tuned dipole, multiband and  
tunable-wire antennas, 205  
loaded tower designs, 507  
loading, **503**, 504–505, **506**  
automatic load control (ALC), 515  
in low-frequency operation anten-  
nas, **503**, 504–505, **506**

loading (*Cont.*):

in low-frequency operation anten-  
nas, loaded tower designs for,  
507  
loading coils, dipoles and doublets  
in, 157–160, **157**  
long path transmission, 51  
long wave band, 8  
longwires, 213–229  
angle of radiation in, 214  
antenna tuning unit (ATU) in,  
172, 214  
apex angle in, 223  
baluns in, 171, **171**  
Beverage or “wave” antenna in,  
224–228, **226, 228, 229**  
capacitors for, 215  
characteristic impedance in, 223,  
226  
counterpoise, 170–172, **171**  
double vee antennas, 223  
electrical vs. physical length of,  
213–214, 217, 218–219, 221  
electrostatic fields and electric  
charge in, 219  
end-fed longwire, 217–219, **217**  
feedpoint in, 217–218  
flagpole installation of, 234–236,  
**235, 237**  
gain in, 214, 218, **218**, 223  
ground plane verticals in, 172  
grounding of, 215–216, **216**  
as hidden and limited-space  
antennas, 232  
impedance and impedance  
matching in, 214, 223, 226  
inductors for, 214  
interference and, 214  
L-section coupler for, 214, **215**  
length of, 213–215, 218–219, **218**,  
221, 223  
as marine radio antennas, 497,  
**498**  
maximum effective length (MEL)  
in, 227  
as multiband and tunable-wire  
antennas, 209, **210**  
non-resonant single-wire longwire  
antennas in, 220  
nonresonant vee beams,  
223, **224**  
quarter-wavelength, 214, 217  
radials for, 215

- longwires (*Cont.*):  
 radiation patterns in, 219, **220**,  
 220–221, **222**  
 random-length type, 213–214, **214**  
 resonant vs. non-resonant types  
 in, 213  
 rhombic beam, 221, 223–224,  
**225**  
 safety issues in, 215  
 as shortwave reception antennas,  
 277–278, **278**  
 as standing wave antennas, 220  
 steerable notch Beverage antennas  
 and, 229, **229**  
 terminated antennas as, 220, **221**  
 tilt angle in, 223  
 transmission line for, 217–218  
 traveling wave antennas in, 220,  
 223  
 true longwire antenna, 217–219,  
**217**  
 tuning of, 214, 215, **215**  
 vee beam, 221–223, **222**  
 VHF/UHF transmitting/receiving  
 antennas, 340–341  
 voltage standing wave ratio  
 (VSWR) in, 215  
 wavelength-measurement in, 217,  
 218–219  
 window-sill antennas, 238, **241**  
 wire used in, 214  
 Zepp antenna in, 164–165, **165**  
 “looking in” impedance, 82–86, 372  
 loop antennas, 129, 131  
 loop antennas, large design,  
 287–297, 299  
 (*See also* loop antennas, small  
 receiver-type; loop antennas,  
 small transmitter-type)  
 antenna tuning unit (ATU) in, 287  
 bisquare loop large antenna,  
 296–297, **297**  
 current in, 289  
 delta loop ( $1\lambda$ ) large antenna,  
 290–293, **291**  
 delta loop large antenna, 293, **295**  
 demiquad loop antenna, 293, **294**  
 diamond loop ( $1\lambda$ ) large antenna,  
 290–293, **291**  
 feedpoints in, 287–288, 292  
 gain in, 288–289, **289**  
 half-delta sloper (HDS), 295–296,  
**296**  
 loop antennas, large design (*Cont.*):  
 impedance and impedance match-  
 ing in, 292–293, **292**  
 impedance in, 287–288  
 inductive reactance in, 288–289,  
**289**  
 $\lambda/2$  large loop antennas in,  
 287–289, **288, 289**  
 loss in, 288–289, **289**  
 nulls in, 299–300, **300, 301**  
 ( $1\lambda$ ) large loop antennas,  
 290–293, **290, 291**  
 orthogonal lobes in, 299–300, **300,**  
**301**  
 radiation patterns in, 287,  
 299–300, **300, 301**  
 square loop ( $1\lambda$ ) large antenna,  
 290–293, **290**  
 loop antennas, small receiver-type,  
 299–318  
 (*See also* loop antennas, large  
 design; loop antennas, small  
 transmitter-type)  
 air core frame loops (“box” loops)  
 in, 303–305, **304**  
 amateur radio, 302  
 amplifier for, 313–316, **316**  
 angle of arrival of signal in, voltage  
 vs., 303, **305**  
 Armstrong oscillator circuit for,  
 313, **315**  
 capacitance and capacitors in,  
 304, 307–309, 317  
 capacitance tuning of, 307–309  
 coaxial cable loop antennas, 318,  
**318**  
 cochannel interference, 302  
 coupling loops in, 306–307, **306**  
 current in, 299  
 depth wound loops in, 303  
 DXing using, 311–313  
 effective height of, 304  
 Faraday connection and Faraday  
 shield in, 318  
 Grover’s equation in, 302–303, 303t  
 height of, 304  
 inductance in, 307  
 isopotential line in, 303  
 low frequency (LF), 302  
 nulls in, 299–300, **300, 301**, 303,  
**305**  
 parallel tuning of, 307–309, **308**  
 planar windings in, 303  
 loop antennas, small receiver-type  
 (*Cont.*):  
 preamplifiers for, 304  
 Q–multipliers for, 313, **315**  
 radiation patterns in, 299–300,  
**300, 301**, 303, 313  
 radio direction finding (RDF)  
 using, 301  
 remote tuning in, 307–309, **310**  
 self-resonant point in, 317, **317**  
 series resonant tuning in,  
 307–309, **308**  
 shape constants in, 303, 303t  
 sharpening of, 313  
 shielded loop antennas as,  
 309–311, **311, 312**  
 split frequencies in, 302  
 spoiler loop for, 313, **314**  
 sports fan’s loop as, 309  
 transformer loops in, 306–307, **306**  
 triangulation of radio signal using,  
 302  
 tuning of, 304, 307–309, **308**,  
 312–313  
 UHF, 302  
 use of, 311–313  
 varactor tuning in, 309, **310**  
 VHF, 302  
 VLF/LF loops, special problems in,  
 317  
 voltage in, 303, 304  
 loop antennas, small transmitter-  
 type, 319–325  
 (*See also* loop antennas, large  
 designs ; loop antennas, small  
 receiver-type)  
 bandwidth of, 320–321  
 capacitance and capacitors in,  
 320–321  
 capacitor voltage in, 321  
 circular loop transmitter antenna  
 as, 323, **323**  
 current in, 320, 323  
 design equations for, 321–322  
 distributed capacitance in, 321  
 efficiency in, 321  
 gain in, 319–320  
 inductance in, 321  
 inductive reactance in, 319, 321  
 loss resistance in, 321  
 octagonal loop transmitter an-  
 tenna in, 319, **320, 322, 322**,  
 323, **324**

- loop antennas, small transmitter-type (*Cont.*):  
 Patterson octagonal loop transmitter antennas, 322, **322**  
 quality factor (Q) in, 320–321  
 radiation patterns in, 324, **325**  
 radiation resistance in, 319, 321  
 resistance in, 319, 321  
 square loop transmitter antenna in, 319, **320**  
 tuning capacitance in, 320–321  
 voltage in, 320, 323  
 voltage standing wave ratio (VSWR) in, 320
- loss, 11–12, 19, 27, 140  
 in coaxial cable, 438  
 in high-frequency dipoles and doublets, 143–146  
 in  $\lambda/2$  large loop antennas, 288–289, **289**  
 loss coefficient scale in, 109  
 in microwave waveguides and antennas, 396  
 in multiband and tunable-wire antennas, 210  
 in Smith charts, 108–109  
 in transmission lines, 83–84, 91–94, 369  
 in vertically polarized HF antennas, 183
- loss coefficient scale, 109, 112–113  
 loss resistance, in loop antennas, small transmitter-type, 321  
 lossy circuits, Smith charts, 117, **118**
- low frequency (LF), 421  
 loop antennas, small receiver-type, 302  
 low-frequency operation antennas, 501–513  
 VHF/UHF transmitting/receiving antennas vs., 339
- low-frequency operation antennas, 501–513  
 continuous loading in, **503**, 504–505, **506**  
 delta feed systems in, 507, **509**  
 discrete loading in, **503**, 504–505  
 grounding and ground systems for, 501–502, **502**  
 “hairpin” loading scheme in, 505, **506**  
 impedance matching in, 505
- low-frequency operation antennas (*Cont.*):  
 impedance matching in, of towers, 507, **508**  
 inductance loading of vertical antennas in, **503**  
 inductive loading in, **503**, 504–505  
 inverted-L antenna for, 510, **510**  
 $\lambda/4$  inverted-L antenna for, 510, **510**  
 length of vertical antenna in, 504  
 linear loading in, **503**, 504–505  
 linearly loaded tee antenna for, 511, **512**  
 loaded tower designs for, 507  
 loading in, **503**, 504–505, **506**  
 radial layout for, 502, **502**  
 random-length Marconi antenna for, 507–510, **509**  
 shortened horizontal antennas for, 505  
 shortened tee antenna, 511, **513**  
 shortened vertical antennas for, 502–505, **503**  
 shunt feed systems in, 507
- lower sideband (LSB), 55  
 lowest usable frequency (LUF), 45
- M**  
 magnetic fields, 123, 126, 127, **128**,  
 in microwave waveguides and antennas, 375–376, **375**, 383, 395–396  
 magnetic flux, 126, 127  
 man-made noise (QRM), 245  
 Marconi, Guglielmo, 1  
 marine radio antennas, 1–2, 494–497, **497**  
 Citizens Band radios and, 496  
 grounding and ground systems for, 495–496, **495**, **496**  
 HF SSB mode in, 495  
 line flattener tuner for, 500, **500**  
 longwire antennas, 497, **498**  
 tuners, 497–500, **499**, **500**  
 VHF-FM use in, 495  
 whip antennas for, 496–497, **497**
- mast-mounted vertical antenna, 546, **546**, 560–561, **561**, **562**  
 masts and supports, 556–559, **557**, **558**  
 matching section, quarter-wavelength coax, 292
- matching stub systems, 465–466, **466**, **467**  
 maximum effective length (MEL), 227  
 maximum usable frequency (MUF), 43–45, 47  
 Maxwell’s equations, 8–9  
 measurement and adjustment techniques, 433–438, 515–542  
 antenna adjustment in, 523–524  
 automatic load control (ALC) in, 515  
 basic radio system in, 515, **516**  
 bridge to measure RF power in, 532, **532**  
 capacitance measurement in, 526–527  
 current transformer wattmeter in, 534–538, **535**, **536**, **537**  
 dip oscillator/dip meter in, 527–529, **527**, **528**  
 dummy loads in, 540–542, **540**, **541**, **542**  
 filters in, 515  
 forward power in, 538, **539**  
 grid dip meter in, 527–529, **528**  
 harmonics in, 515  
 high-power RF wattmeters for, 538, **538**  
 impedance bridges in, 519–521, **520**, **521**  
 impedance in, 525  
 inductance measurement in, 526–527  
 LC tank circuit in, 526, **526**  
 matching unit in, 515  
 micromatch wattmeter in, 533, **533**  
 monomatch wattmeter in, 533–534, **534**  
 nonresonant antenna adjustment in, 524–525  
 parallel resonant circuits in, 526  
 power in, 515, 531–532  
 power-frequency calibration in, 538, **538**  
 reactance testing/adjustment in, 524  
 reflected power in, 538, **539**  
 resistance testing/adjustment in, 523–524  
 resonance testing/adjustment in, 524

- measurement and adjustment techniques (*Cont.*):  
 RF noise bridge in, 521–522, **521**  
 RF power measurements in, 529–539  
 RF wattmeters for, 529–539, **531**  
 series tuned circuits in, 526  
 surge impedance ( $Z_0$ ) in, 516  
 thermocouple for RF ammeter in, 530, **530**  
 transmission lines in, 516–517, 525  
 tuned circuit measurements in, 526  
 tuning antennas in, 524  
 voltage measurement in, 531  
 voltage standing wave ratio (VSWR) in, 515, 517–519, 529, 532  
 Wheatstone bridge in, 519–521, **520**
- medium wave (MW), 8, 17, 29, 48, 61, 224–228, **226, 228, 229**
- meteor scatter propagation, 54
- micromatch wattmeter, 533, **533**
- microwave transmissions, 3, 5, 8, 13, 17, 18, 19, 22, 23, 29, 30, 32, 37, 40, 61, 69, 369
- microwave waveguides and antennas, 369–415, 421  
 (*See also* parabolic “dish” antennas)  
 adjustable plunger or short-circuit termination in, 388, **388**  
 antenna design using, 394  
 aperture (slot) coupling in, 390–393, **393**  
 aperture in, 402–403  
 array antennas in, 409–411  
 attenuation in, 377  
 backlobes in, 400, **400, 401**  
 beamwidth in, 400, **400, 401**  
 boundary conditions in, 375–376  
 branch feed systems for, 409, **412**  
 capacitive (probe) coupling in, 390–393, **392**  
 cavity antennas, 404, **405**  
 characteristic impedance in, 383–384  
 choke joints in, 389, **390**  
 coordinate system to denote dimensions in, 376–377, **377**  
 corner reflector antennas, 404, **405, 406**
- microwave waveguides and antennas (*Cont.*):  
 corporate feed systems for, 409, **412**  
 coupling methods for, 390–393  
 cutoff frequency ( $F_c$ ) in, 381–383  
 dielectric loss, 369  
 dipole antenna elements and, 396–398, **397, 399**  
 directivity gain ( $G_d$ ) in, 401–402  
 directivity in, 395, 397, 398–403, **400**  
 dominant mode in, 376–377  
 dummy loads in, 385–388, **386**  
 effective aperture ( $A_e$ ) in, 394  
 effective radiated power (ERP) in, 401  
 electric fields in, 383, 395–396  
 end plate in, 385, 388, **388**  
 flatplate array antennas, 411, 413, **414, 415**  
 free-space velocity ( $C$ ) in, 377, 380  
 frequency in, 373, **374, 375, 376, 380, 383–384, 384**  
 gain in, 394, 395, 398–403  
 gain vs. aperture in, 402–403  
 graphite-sand dummy load in, 385  
 group velocity ( $V_g$ ) in, 377, 378–379  
 horn antenna radiators in, 403–404, **403**  
 $I^2R$  losses in, 385  
 impedance and impedance matching in, 383–384, 394, 396  
 inductive (loop) coupling in, 390–393, **393**  
 insulators in, 371–372  
 interference and, 400–401  
 inverse square law ( $1/D^2$ ), 369–370  
 isotropic antenna waveguide, 394–403  
 isotropic radiation in, 394  
 joints and bends in, 388–390, **389, 390**  
 law of reciprocity and, 394  
 length in, 392  
 losses in, 396  
 magnetic fields in, 375–376, **375, 383, 395–396**  
 materials used in, 370  
 near field and far field calculations in, 395–396
- microwave waveguides and antennas (*Cont.*):  
 ohmic loss in, 369, 371, 396  
 parabolic “dish” antennas as, 404–409, **407, 408, 409**  
 phase shifting in, 409, **412, 413**  
 phase velocity ( $V_p$ ) in, 377, 379, 380  
 pipe analogy of, 369–370, **370**  
 power gain ( $G_p$ ) in, 401–402  
 propagation constant ( $B$ ) in, 384  
 propagation modes in, 373–381  
 quarter-wavelength shorted stub vs., 372–373, **372, 373, 374**  
 radiating vs. nonradiating slots in, 393  
 radiation loss in, 369, 372  
 radiation patterns and, 397, **398**  
 radiation resistance in, 396  
 rectangular waveguide using parallel transmission lines and, 370–373, **371**  
 reflector antennas, 404, **405, 406**  
 resistive vane dummy load in, 385  
 rotating joints in, 390, **391**  
 safety issues, 413–415  
 semipermanent joints in, 389  
 shielding in, 392  
 sidelobes in, 400, **400, 401**  
 skin effect in, 369, 371  
 slot array antennas in, 411–413, **413**  
 solid-state array antennas in, 408–411  
 terminations for, 385–388  
 transducer action of, 394  
 transverse electric mode (TE mode) in, 375  
 transverse electromagnetic (TEM) fields in, 373  
 transverse magnetic mode (TM mode) in, 375  
 velocities in, 377–381, **378**  
 voltage standing wave ratio (VSWR) in, 388  
 wavelength in, 377–381, **378**  
 wedge dummy loads in, 385, **386, 387**
- military radio, multiband and tunable-wire antennas, 203
- MiniNEC modeling software, 331
- miniNEC-3/miniNEC-4 for modeling software, 328–329

“mirroring” of radio waves, 43, 54  
 mismatch (VSWR) losses, transmission lines, 91–94  
 mixing, 55  
 mobile HF antennas, 479–485  
   coil-loaded antenna as, 480–481, **480, 351**  
   field strength meters (FSM) for, 483–485, **484, 485, 485t**  
   helical whip antenna as, 481, **482**  
   multiband top-loaded antenna as, 481, **482**  
 Q factor in, 483  
 radial-to-mobile-antenna connection detail in, 488, **489**  
 resonance in, 483  
 top-loaded antenna as, 481, **482**  
 tuning of, 483–485  
 voltage standing wave ratio (VSWR) in, 483  
 used as windowsill-mounted, 239, **241**  
 modeling software for antennas, 327–338  
   coordinate system used in, 328, **329, 330**  
   Expert MiniNEC, 331  
   EZNEC for Windows, 331, 334–337, **335, 336, 337**  
   EZNEC Professional, 329  
   GNEC, 329–330  
   MiniNEC, 331  
   miniNEC-3/miniNEC-4, 328–329  
   NEC-2, 330  
   NEC4Win95, 331  
   NECWin Basic, 331–334, **332**  
   NECWin Plus, 330–331  
 Numerical Electromagnetic Computation (NEC) programs in, 328  
 radiation patterns displayed in  
   EZNEC for Windows, 336–337, **337**  
   radiation patterns displayed in  
   NECWin Basic, 332, **333, 334**  
 voltage standing wave ratio (VSWR) modeling in, 333, **335**  
 voltage standing wave ratio (VSWR) modeling in, 337, **338**  
 monomatch wattmeter, 533–534, **534**  
 monopulse feeds, parabolic “dish” antennas, 409, **410**

monthly variations in radio propagation, 46  
 Morse code, 1, 2  
 mounting (*See also* construction and installation techniques)  
   Beverage antenna, 228  
   dipole, 233, 256  
   directional beam antenna, 256  
   emergency operations antenna, on vehicles, 491–492  
   fence posts as supports for vertical antenna, 561–565, **563, 564**  
   floor flanges in, 256  
   hidden and limited-space antenna, 233  
   high-frequency dipole and doublet, 141  
   mast-mounted vertical antenna, **546**  
   pulley system for, 557–558, **558**  
   right-angle flange hardware in, 256  
   shortwave reception antenna, 277–278, **278**  
   stirrup-end hardware in, 558, **559**  
   temporary mast mount for emergency antenna, 492, **493**  
   U-bolt hardware, 558, **559**  
   vehicle-mounted emergency antenna, 491–492  
   vertically polarized HF antenna, 186, 189–190, 194–195, **194, 195–197, 196, 560–565, 561**  
   VHF/UHF transmitting/receiving antenna, 347  
 multi-impedance balun, 476, **477**  
 multiband and tunable-wire antennas, 203–212  
   antenna tuning unit (ATU) for, 210  
   balanced transmission line and, 208  
   balancing line, 208–209, **209**  
   baluns for, 208–209, **209, 210**  
   coaxial cable in, 208  
   dipoles in, 204–205, **205**  
   end-fed Zepp antenna in, 208, **208**  
   feeding parallel transmission line for, 208–212  
   feedpoint in, 204, 210  
 G5RV multiband dipole antenna in, 206–208, **207**

multiband and tunable-wire antennas (*Cont.*):  
   harmonics in, 210  
   impedance in, 210  
   insulators (spreaders) in, 206  
   inverted-vee dipole for, 205, **206**  
   length of, 204  
   link-tuned dipole for, 205  
   longwires for, 209, **210**  
   losses in, 210  
   quad beam antennas for, 203  
   random-length longwire antennas in, 209, **210**  
   transformers for, 210, **211**  
   transmatches in, 208–209, **209**  
   trap dipoles in, 204–206, **204, 205**  
   tunability of, 203  
   tuned feeder antennas in, 206, **207**  
   tuning of, 205  
   voltage standing wave ratio (VSWR) and, 203  
   Windom antenna as, 209–210, **211**  
   Yagi antennas as, 203  
 multiband fan dipole, 169–170, **170**  
 multiband top-loaded mobile antenna, 481, **482**  
 multifilar windings in transformers, 472  
 multihop communication, 23, **49, 49–50**  
 multihop interference, 36, **36**  
 multiple helical antennas, 428–431, **429, 430**  
 multiple radials from vehicle-mounted antenna, 488, **489**

## N

NAA radio station, 2  
 National Bureau of Standards (*See* National Institute of Standards and Technology)  
 national identifiers in call signs, 3  
 National Institute for Standards and Technology (NIST), 40, 47  
 Naval Research Laboratory, 2  
 near field and far field calculations in microwave waveguides and antennas, 395–396  
 NEC-2 modeling software, 330  
 NEC4Win95 modeling software, 331  
 NECWin Basic modeling software, 331–334, **332**

- NECWin Plus modeling software, 330–331
- Nittany-Scientific, 329, 330, 331
- nodes, 87, 129
- noise, 37, 68–69
- antenna noise temperature as, 417–420
  - contributors to antenna noise temperature and, 419–420, **419**
  - environmental/natural noise (QRN) in, 245
  - figure of merit and, 418
  - impedance and, 417–418
  - Johnson noise in, 417
  - man-made noise (QRM) in, 245
  - noise bridge in, 521–522, **521**, **522**, **523**
  - noise factor ( $F_n$ ) in, 418
  - noise figure (NF) in, 418
  - noise temperature ( $T_e$ ) in, 418–420
  - omnidirectionality vs., 245
  - piezoelectric, 69
  - RF noise bridge in, 521–522, **521**
  - signal-to-noise ratio (SNR) in, 246
  - sky noise sources, 419–420
  - thermal agitation noise, 417
  - thermal noise in, 417–420
  - in transmission lines, 68–69
  - in Yagi antennas, 261
- noise bridge, 521–522, **521**, **522**, **523**
- noise factor ( $F_n$ ), 418
- noise figure (NF), 418
- noise temperature ( $T_e$ ), 418–420
- nonreciprocal direction propagation, 54
- nonresonant antenna adjustment, 524–525
- nonresonant loop antennas, 234–235, **234**
- nonresonant single-wire longwire antennas, 220–220
- nonresonant vee beam longwire antenna, 223, **224**
- normalization of impedance, 95–97, **98**, 469–470
- northern lights, 54
- null steering, 229
- nulls, in loop antennas, 299–300, **300**, **301**, 303, **305**
- Numerical Electromagnetic Computation (NEC) programs, 328
- O**
- 1 $\lambda$  large loop antennas, 290–293, **290**, **291**
- octagonal loop transmitter antenna, 319, **320**, 322, **322**, 323, **324**
- off-center-fed full-wave doublet (OCFD), 163, **163**
- off-center-fed nonresonant sloper (OCFS), 164, **164**
- ohmic loss, 143–146, 369, 371, 396
- ohmic resistance, 19, 27, 140
- ohmmeter, coaxial cable test, 437–438, **437**
- Ohm's law, transmission lines, 74
- $\Omega$ -fed grounded vertical antenna, 200–201, **202**
- omnidirectionality:
- in high-frequency dipoles and doublets, 146
  - noise vs., 245
  - in shortwave reception antennas, 272, **273**
  - in vertically polarized HF antennas, 173, 179–181
- open line, 59
- Orion, 331
- orthogonal lobes, in loop antennas, 299–300, **300**, **301**
- OSCAR satellite communications, 3, 339
- out-of-phase waves, **22**
- P**
- parabolic “dish” antennas, 404–409, **407**, **408**, **409**, 421
- beamwidth in, 407
  - Cassegrain feed in, 408–409, **409**
  - cutoff frequency in, 406
  - directrix in, 404
  - effective aperture in, 407
  - feed geometries for, 408–409, **408**, **409**
  - feed illumination in, 406
  - gain in, 405–408
  - monopulse feeds, 409, **410**
  - ray tracing showing operation of, **407**
  - splash plates in, 408
  - x-y coordinates in, 404
- parallel lines, 59–63, **60**, 65–66
- parallel resonant circuits, 526
- parallel tuning in loop antennas, 307–309, **308**
- parasitic elements in Yagi antennas, 258–259
- paths, 17–18, **18**
- Patterson octagonal loop transmitter antenna, 322, **322**
- pedestals for towers, 568–570, **569**
- permits, zoning, and regulations, 547–548
- phase, 126, **128**
- phase constant, 82
- phase control circuit (PCC), 229
- phase reversal stubs, collinear “Franklin” array antenna in, 165–166, **166**
- phase shift, 19, 21–23, 129, 409, **412**, **413**, 450–451, **451**
- phase velocity ( $V_p$ ) in microwave waveguides and antennas, 377, 379, 380
- phase-shifting circuit, directional phased vertical antennas, 252, **252**
- phase-switching antenna matcher, shortwave reception antennas, 284, **285**
- phasing box, shortwave reception antennas, 284, **286**
- phasing switch, directional phased vertical antennas, 249–250, **252**
- phasing transformer, 249–250, **250**, **251**
- phasing, phased radials, directional phased vertical antennas, 246–250
- pi networks, 461, **461**
- “picket fencing,” 23
- piezoelectricity, 69
- planar windings in loop antennas, 303
- planetary terminator, 51, **53**
- plunger or short-circuit termination, microwave waveguides and antennas, 388, **388**
- polarity, 126–127, 131, 134
- reversal of, 126–127, 131, 134
- polarization of radio waves, 12–13, **13**, 27, **28**, 29, 55, 173
- in Beverage antenna, 226–227
  - in directional phased vertical antennas (*See* directional phased vertical antennas)

polarization of radio waves (*Cont.*):  
 in vertically polarized HF  
 antennas, 173–202  
 polarization diversity, 55, **58**  
 power, 11, **21**, 140, 515, 531–532  
 in RF power measurements,  
 529–539  
 in vertically polarized HF  
 antennas, 180  
 power density, 10  
 power fading, 36  
 power gain ( $G_p$ ) in microwave  
 waveguides and antennas,  
 401–402  
 power-frequency calibration in,  
 538, **538**  
 preamplifiers for loop antennas,  
 304  
 predictions of propagation, 55, **56**  
 propagation constant (B), 82,  
 384  
 propagation modes in microwave  
 waveguides and antennas,  
 373–381  
 propagation of radio waves, 5–58  
 propagation predictions, 55, **56**  
 pulley system for mounting,  
 557–558, **558**  
 pure reactance circle, Smith charts,  
 100, **103**  
 pure resistance line, Smith charts,  
 95–97, **98**

**Q**  
 Q–multipliers for loop antennas,  
 small receiver-type, 313, **315**  
 QRM noise, 245  
 QRN noise, 245  
 quad beam VHF/UHF antenna, 356,  
 358, **359**  
 quad beam antennas, 203  
 quad loop antennas, 263–267, **264**,  
**266**, **267**  
 quality factor (Q), 85, 160,  
 320–321, 483  
 quarter-wave matching (Q section)  
 sections, 85, 146, 199–200, **199**,  
 467–468, **468**, **469**  
 quarter-wave vertical polarized HF  
 antenna, 173  
 quarter-wavelength longwires, 214,  
 217  
 quarter-wavelength lines, 84–85

quarter-wavelength shorted stub  
 “looking in” impedance in, 372  
 microwave waveguides and anten-  
 nas vs., 372–373, **372**, **373**, **374**  
 transmission lines and, 86, **87**, **88**  
 quarter-wavelength vertical  
 antenna, 181, **182**, 183, 187

**R**

radar, 3, 30, 52  
 radial-to-mobile-antenna connection  
 detail, 488, **489**  
 radially scaled parameters, Smith  
 charts, 106–109, **107**  
 radials:  
 in grounding and ground systems,  
 577–579, **579**, **580**  
 in impedance matching, 460  
 in longwires, 215  
 in low-frequency operation anten-  
 nas, 502, **502**  
 multiple, from vehicle-mounted  
 antenna for, 488, **489**  
 radial-to-mobile-antenna connec-  
 tion detail in, 488, **489**  
 in vertically polarized HF anten-  
 nas, 176, 186, **189**, 560–561  
 radiated fields, 5  
 radiating vs. nonradiating slots,  
 microwave waveguides and  
 antennas, 393  
 radiation angle, 30, 48–50, **50**  
 radiation loss, 369, 372  
 radiation patterns:  
 in Adcock antenna, **449**  
 in beam antennas, 257–258, **258**  
 cardioid, 441, **442**  
 in directional phased vertical  
 antennas, 246–250, **246**, **247**  
 EZNEC for Windows modeling of,  
 336–337, **337**  
 in high-frequency dipoles and  
 doublets, 146–152, **147**  
 in isotropic antenna waveguide,  
 397, **398**  
 in  $\lambda/2$  large loop antennas, 287  
 in longwires, 219, **220**, 220–221,  
**222**  
 in loop antennas, large designs,  
 299–300, **300**, **301**  
 in loop antennas, small receiver-  
 type, 299–300, **300**, **301**, 303,  
 313  
 radiation patterns (*Cont.*):  
 in loop antennas, small transmitter-  
 type, 324, **325**  
 in loopstick antennas, 439, **440**,  
**441**  
 NECWin Basic modeling of, 332,  
**333**, **334**  
 in sense antennas, 441, **442**  
 in switched-pattern RDF antenna,  
 454–456, **454**  
 in vertically polarized HF anten-  
 nas, 178–181, **181**, **182**  
 in Watson-Watt Adcock array  
 antennas, **450**  
 in Yagi antennas, 257–258, **258**  
 radiation resistance, 140  
 in high-frequency dipoles and  
 doublets, 143–146  
 in loop antennas, small transmitter-  
 type, 319, 321  
 in microwave waveguides and  
 antennas, 396  
 in vertically polarized HF antennas,  
 183, **184**, **185**  
 radiators, high-frequency dipoles  
 and doublets, 141  
 radio astronomy antennas, 421–432  
 directional discontinuity ring  
 radiator (DDRR) antenna in,  
 424–427, **426**, 427*t*  
 helical antennas in, 427–428, **429**  
 interferometer antennas in, 431,  
**431**, **432**  
 Jupiter reception antenna,  
 421–422, **423**  
 multiple helical antennas in,  
 428–431, **429**, **430**  
 ring antenna, 422–424, **423**, **424**,  
**425**  
 summation interferometer array  
 antenna, 431, **431**  
 radio direction finding (RDF),  
 439–456  
 Adcock antennas for, 446–447,  
**448**, **449**  
 AM broadcast band (BCB) in, 439  
 birdirectionality of loopsticks in,  
 441  
 cardioid radiation pattern of sense  
 antenna in, 441, **442**  
 CONALRAD station frequencies  
 and, 439  
 Doppler antennas, 450–451, **451**

- radio direction finding (RDF)  
 (Cont.):  
 double ducky direction finders  
 (DDDF) in, 455–456, **455**  
 “fox hunting” antenna in, 444, **445**  
 Global Positioning System (GPS)  
 and, 443  
 improvisation of, using portable  
 AM radio receiver, 443, **443**  
 loop antennas in, small receiver-  
 type, 301  
 loopstick antennas in, 439–440,  
**440**  
 measurements in, 433  
 radiation patterns of loopsticks in,  
 439, **440, 441**  
 regular wire loop antennas for,  
 444, **445**  
 “rubber ducky” VHF antennas in,  
 454–456, **455**  
 S meters in, 439  
 sense antenna for, 441, **442, 446,**  
**447**  
 shortwave and AM BCB “skip” in,  
 444–446  
 switched-pattern RDF antenna in,  
 454–456, **454**  
 time difference of arrival (TDOA)  
 array antenna in, 452–454,  
**453**  
 triangulation of radio signal in,  
 439  
 Watson-Watt Adcock array anten-  
 nas in, 447–450, **449, 450**  
 wide-aperture direction finder  
 (WADF) in, 451  
 Wullenweber array antennas in,  
 451, **452**
- Radio Luxembourg effect, 54  
 radio system, 515, **516**  
 radio vectoring, 3  
 radio wave configuration, 5–6  
 random-length longwires, 209, **210,**  
 213–214, **214, 507–510, 509**  
 ray tracing, 32–33, **407**  
 reactance, 140, 524  
 (See also capacitive reactance;  
 inductive reactance)  
 in high-frequency dipoles and  
 doublets, 143  
 Smith charts, 109, 117  
 Smith charts, capacitive reactance  
 in, 100, **102**
- reactance (Cont.):  
 Smith charts, constant reactance  
 circles in, 100–105  
 Smith charts, inductive reactance  
 in, 100, **101**  
 Smith charts, pure reactance  
 circle in, 100, **103**  
 in transmission lines, 85–86, 109,  
 525  
 in vertically polarized HF antennas,  
 183, **184**
- Reber, Grote, 421  
 receivers, 123  
 reciprocity, law, 226, 271, 394  
 recombination phenomena, 18  
 rectangular waveguide using paral-  
 lel transmission lines, 370–373,  
**371**  
 reflected power ( $P_r$ ), 89, 538, **539**  
 reflected waves, 128, 134  
 reflection, 16–17, 21, 22, 50, 52,  
 161  
 reflection coefficient, 21, 22  
 Smith charts, **103, 106, 106–109,**  
 112, 114  
 transmission lines, 71–79, 89, 112,  
 114  
 reflection loss, 113–114  
 reflector antennas, 404, **405, 406**  
 reflectors, Yagi antennas,  
 258–259  
 refraction, 16–17, 16, 23–24, **24,**  
 29–30, **30, 48**  
 refractivity of atmosphere, 24–27  
 regular wire loop antennas, 444,  
**445**
- remote tuning, loop antennas, small  
 receiver-type, 307–309, **310**  
 repeaters for emergency operations  
 antennas, 487  
 resistance, 27, 140, 369, 434–437,  
 523–524  
 in high-frequency dipoles and  
 doublets, 143–146, **145**  
 in loop antennas, small transmitter-  
 type, 319, 321  
 in Smith charts, constant or isore-  
 sistance circles in, 97–100, **99**  
 in Smith charts, pure resistance  
 line in, 95–97, **98**  
 in Smith charts, unity resistance  
 circle in, 100  
 in transmission lines, 64, 525
- resistance (Cont.):  
 in vertically polarized HF  
 antennas, 183  
 resistive vane dummy load in  
 microwave waveguides and  
 antennas, 385  
 resonance, 140, 524  
 in dipoles, 237  
 in high-frequency dipoles and  
 doublets, 143, 149–151, **150**  
 in mobile HF antennas, 483  
 resonant frequency, 433  
 resonant vs. non-resonant long-  
 wires, 213  
 return loss, 108–109, 113, 114  
 RF ammeter, 530, **530**  
 RF noise bridge, 521–522, **521**  
 RF power measurements,  
 529–539  
 RF wattmeters, 529–539, **531**  
 rhombic beam longwire antenna,  
 221, 223–224, **225**  
 rhombic inverted-vee antenna, 169,  
**169**  
 right-angle flange mounting hard-  
 ware, 256  
 ring antenna, 422–424, **423, 424,**  
**425**  
 RLC networks, 59, 143  
 rotatable dipole antennas, 239,  
**242, 255–257, 257**  
 rotating joints in microwave wave-  
 guides and antennas, 390, **391**  
 “rubber ducky” VHF antennas,  
 454–456, **455**
- S**  
 S meters in RDF antennas, 439  
 safety issues, 543–548  
 burns or shocks from RF through  
 vertical antennas, 191, 195  
 hidden and limited-space  
 antennas, 233  
 longwires, 215  
 microwave waveguides and anten-  
 nas, 413–415  
 shortwave reception antennas,  
 276, **277, 279**  
 VHF/UHF transmitting/receiving  
 antennas, 340  
 “salt pipe” to apply chemical soil-  
 conductivity improvers,  
 576–577, **577**

- salt to alter soil conductivity, 575–576
- satellite communications, 3, 339, 421
- scanner band VHF/UHF antennas, 360–368
- scanner skyhook VHF/UHF antennas, 362–368, **363**
- scanner-vision antennas, 360–362, **361**
- scatter propagation, 51–54, **53**
- seasonal variations in radio propagation, 24, 40, 41, 46, 51, 178
- selective fading, 55
- self-resonant point, loop antennas, small receiver-type, 317, **317**
- semipermanent joints in microwave waveguides and antennas, 389
- sense antenna in RDF, 441, **442**, 446, **447**
- series matching section, 468–471
- series resonant tuning, loop antennas, small receiver-type, 307–309, **308**
- series tuned circuits, 526
- series-fed vertical antennas, 200
- shape constants, loop antennas, small receiver-type, 303, 303*t*
- shielded loop antennas, 309–311, **311**, **312**
- shielded parallel line, characteristic impedance of, 66
- shielded twin lead, 61, **61**
- shielding in microwave waveguides and antennas, 392
- shocks hazards from RF through antennas, 191, 195, 215, 233, 276, 279
- short path transmission, 51
- short skip, 41
- short-circuit termination, in microwave waveguides and antennas, 388, **388**
- shorted lines, 87
- shorted stubs, transmission lines, 86, **87**
- shorted stubs (*See* stub matching systems)
- shortened coil-loaded dipoles, 157–160, **158**
- shortened tee antenna, 511, **513**
- shortwave and shortwave reception antennas, 2–4, 8, 37, 41, 54–55, 271–286
- shortwave and shortwave reception antennas (*Cont.*):
- angle of radiation in, 274, **274**
  - antenna matching transformer in, 281, **285**
  - antenna matching tuner in, 281, **285**
  - connection to receiver in, 274–276, **275**, **276**
  - dipole as, 272, **273**, 274
  - directional wire antenna in, 281–286, **284**
  - directivity of, 272
  - flattop antennas for, 279, **280**
  - gain in, 272
  - grounding for, 276, **277**, 278–279, **279**
  - interference and, 272, **273**, 274
  - isotropic radiation in, 272
  - law of reciprocity in, 271
  - longwire antennas for, 277–278, **278**
  - mounting for, 277–278, **278**
  - multiband wire vertical antennas for, 281, **282**, **283**
  - omnidirectionality in, 272, **273**
  - phase-switching antenna matcher in, 284, **285**
  - phasing box for, 284, **286**
  - properties of importance in, 271–274
  - radio direction finding (RDF) antennas for, 444–446
  - safety issues in, 276, **277**, 279
  - skip propagation and, 274
  - transformers in, 281
  - vertical antennas for, 279–281, **280**, **282**, **283**
- shunt feed systems, 200, 507
- side scatter, 52
- sidelobes in microwave waveguides and antennas, 400, **400**, **401**
- signal-to-noise ratio (SNR), 37, 45, 246
- simple refraction, 29
- single-sideband (SSB), 55, 495
- skin effect, 369, 371
- skip distance, 47–48, **48**, 47
- skip propagation, 5, 29, 37, 41, 47–48, **48**, **49**, **51**, 55
- radio direction finding (RDF) antennas and, 444–446
- skip propagation (*Cont.*):
- shortwave reception antennas and, 274
  - vertically polarized HF antennas and, 176–179, **178**
  - VHF/UHF transmitting/receiving antennas and, 339
- skip zone, 47–48, **48**
- sky noise sources, 419–420
- sky wave, 47–48, **48**, **49**, 302
- skyhook VHF/UHF antennas, 362–368, **363**
- slip-up tower, 567–568, **568**, **569**
- sloping (“sloper” or “slipole”)
- dipole, 153–154, **154**, **155**, 236, **238**
  - half-delta sloper (HDS), 295–296, **296**
  - off-center-fed nonresonant sloper (OCFS), 164, **164**
  - vee-sloper antenna in, 168–169, **168**
- slot array antennas, 411–413, **413**
- Smith charts, 95–121, **96**
- absorbed power ( $P_a$ ) in, 113–114
- admittance in, 100, 112, 114
- angle of reflection coefficient
- circle in, **103**, 106, 112
- angle of transmission coefficient
- circle in, **103**, 106
- applications for, 109–121
- capacitive reactance in, 100, **102**, 117
- complex admittance in, 105
- components of, 95–109
- constant reactance circles in, 100–105
- constant resistance circles in, 97–100, **99**
- frequency in, 117–121, **119**, **120**
- impedance and impedance matching, 110–112, **111**, 115, 119–121, **119**
- impedance radius in, 100
- inductive reactance in, 100, **101**, 117
- iso-resistance circles in, 97–100, **99**
- loss coefficient scale in, 109, 112–113
- lossy circuits and, 117, **118**
- normalized impedance line in, 95–97, **98**

- Smith charts (*Cont.*):  
 pure reactance circle in, 100, **103**  
 pure resistance line in, 95–97, **98**  
 radially scaled parameters in,  
 106–109, **107**  
 reactance in, 109, 117  
 reflection coefficient in, 106–109,  
 114  
 reflection loss in, 113–114  
 return loss in, 108–109, 113, 114  
 standing wave ratio (SWR) in, 108  
 stub matching systems problems  
 solved using, 109, 114–117,  
**115, 116**  
 susceptance in, 117  
 transmission line problems solved  
 using, 109–114  
 transmission loss coefficient  
 (TLC) in, 112, 114  
 transmission loss in, 108–109  
 unity resistance circle in, 100  
 voltage standing wave ratio  
 (VSWR) circle in, 100, **104**,  
 112, 114, 115, 117  
 wavelength calculation in, 110,  
 115  
 wavelength circles in, **103**, 105  
 Smith, Phillip H., 95  
 soil conductivity, 27, 28*t*, 575–577,  
 576*t*, **576**, 577  
 solar flares vs. radio propagation,  
 37, 39, 46–47  
 solar flux index (SFI), 40  
 solar storms, 47  
 solid-state array antennas, 408–411  
 SOS, 2  
 southern lights, 54  
 space diversity, high-frequency  
 dipoles and doublets, 160  
 space quadrature fields, 127  
 space waves, 17–19, **18**  
 spark-gap transmitters, 2  
 spatial diversity, 55, **57**  
 SPC transmatch network, 463, **463**  
 “spider web” ground system,  
 vertically polarized HF antennas,  
 185, **188**  
 splash plates, parabolic “dish”  
 antennas, 408  
 split frequencies, loop antennas,  
 small receiver-type, 302  
 split-capacitor networks, 461–462,  
**462**
- spoiler loop, loop antennas, small  
 receiver-type, 313, **314**  
 sporadic E propagation, 41, 46  
 sports fan’s loop antennas, small  
 receiver-type, 309  
 spreaders, multiband and tunable-  
 wire antennas, 206  
 square loop ( $1\lambda$ ) large antenna,  
 290–293, **290**  
 square loop transmitter antenna,  
 319, **320**  
 stacked dipoles, 160–163, **162**  
 stagger tuning in high-frequency  
 dipoles and doublets, 157  
 standard refraction, 25  
 standing wave antennas, 220  
 standing wave ratio (SWR), 434  
 in high-frequency dipoles and  
 doublets, 144  
 in Smith charts, 108  
 in transmission lines, 86–91, **90**  
 standing waves, 87, 127–131, **130**,  
**131**  
 current, 134, **135**, 136, **137**, 139  
 in high-frequency dipoles and  
 doublets, 144  
 measurement of, 137, 138  
 in transmission lines, 71, **72**, **73**  
 voltage, 134–136, **135**, **136**, 139  
 station grounding, 579–580, **581**  
 stationary waves, 128  
 steady-state ac response, transmis-  
 sion lines, 70, 79–82, **81**  
 steerable notch Beverage antennas,  
 229, **229**  
 step-function response, transmis-  
 sion lines, 70–71, **70**, **71**  
 stirrup-end mounting hardware,  
 558, **559**  
 storms, ionospheric, 47  
 strain relief, 553–554, **554**  
 stratosphere, 14, **15**, 18  
 stray capacitance, 138  
 stripline or microstripline, 63, **64**,  
 66–67  
 stub matching systems, 114–117,  
**115**, 465–466, **466**, **467**  
 impedance matching in, 115  
 length of stub calculation in,  
 117  
 shorted stubs, 114  
 Smith charts for, 109, 114–117,  
**115, 116**
- stub matching systems (*Cont.*):  
 in VHF/UHF transmitting/receiving  
 antennas, 345–346, **345**  
 voltage standing wave ratio  
 (VSWR) in, 115, 117  
 wavelength calculation in, 115  
 subrefraction, 31, **32**, 36  
 sudden ionospheric disturbances  
 (SIDs), 46–47  
 summation interferometer array  
 antenna, 431, **431**  
 sunspots vs. radio propagation,  
 40–41, **40**, 43, 46, 47  
 superrefraction, 29, **31**, 26  
 support posts, 565, **565**  
 support ropes, 551  
 suburban lot installation plan, **545**  
 surface waves, 17–19, **18**, 27–29  
 surge impedance (*See* character-  
 istic impedance)  
 susceptance, Smith charts, 117  
 switched-pattern RDF antenna,  
 454–456, **454**  
 SWR analyzer, 434
- T**
- 360-degree directional array, 250,  
**253**, 254*t*  
 tapped balun, 476, **477**  
 tee antennas, 511, **512**, **513**  
 tee-network transmatch, 463, **463**  
 Teflon, 61, 62  
 telegraph, 1, 2  
 telescoping antennas, emergency  
 operations, 491  
 television, 3, 32  
 television interference/broadcast  
 interference (TVI/BCI), 214  
 television-scanner VHF/UHF  
 antennas, 360–362, **361**  
 temperature effect on propagation,  
 29–32, **32**  
 temporary antenna construction,  
 emergency operations,  
 487–488, **488**  
 temporary mast mount for emer-  
 gency antennas, 492, **493**  
 terminated longwire antennas, 220,  
**221**  
 terminations for microwave wave-  
 guides and antennas, 385–388  
 terrain effects, 28, 32, **33**  
 thermal agitation noise, 417

- thermal noise, 68, 417–420
- thermocouple for RF ammeter, 530, **530**
- Thorne array antenna, 267–269, **268, 269**
- three-dimensional (ideal) radiation pattern, dipoles and doublets, 147–148, **147**
- tilt angle of surface wave, 28, 28*t*
- tilt angle, rhombic beam longwire, 223
- tilted, center-fed, terminated, folded dipole (TCFTFD or T<sup>2</sup>FD, TTFD), 166–168, **167**
- time difference of arrival (TDOA) array antenna, 452–454, **453**
- time domain reflectometry (TDR), 75–79, **76, 77, 78, 79, 80, 81**
- time quadrature fields, 127
- time-of-day variations in radio propagation, 14, 16, 32, 39–41, 43, 46, 51, 178
- top-loaded mobile antenna, 481, **482**
- toroid broadbanded transformers, 475–477, **476**
- toroid cores, 471, **472**, 474
- toroidal transformers, 471–475, **472**
- towers, **538**, 565–572
- fold- or tilt-over, 567, **567**
- gin pole to raise, 570, **571**
- guy wires and ground stakes for, 570, **572**
- hinged baseplate, **538**, 566–567
- impedance matching in, 507, **508**
- low-frequency operation antennas, loaded tower designs for, 507
- masts and supports for, 556–559, **557, 558**
- pedestals for, 568–570, **569**
- raising of, 570, **570, 571**
- safety issues in, 543–548
- slip-up, 567–568, **568, 569**
- transducer action, microwave waveguides and antennas, 394
- transformer loop antennas, 306–307, **306**
- transformers, 199–200
- core material in, 474
- current transformer wattmeter in, 534–538, **535, 536, 537**
- transformers (*Cont.*):
- in directional phased vertical antennas, 249–250, **250, 251**
- feeding type, 476, **477**
- ferrite rods in, 475, **475**
- in high-frequency dipoles and doublets, 146
- multi-impedance balun and, 476, **477**
- in multiband and tunable-wire antennas, 210, **211**
- multifilar windings in, 472
- in shortwave reception antennas, 281
- tapped balun and, 476, **477**
- toroid core mounting in, 471, **472**, 474
- toroid broadbanded, for impedance matching, 475–477, **476**
- toroidal type, 471–475, **472**
- in transformer loop antennas, 306–307, **306**
- trifilar windings in, 472
- turns in, 474
- in VHF/UHF transmitting/receiving antennas, 341–346, **342, 343**
- windings in, 472–473, **473**
- transmatch circuits, 208–209, **209**, 462–471, **462**
- transmission lines, 59–94, 516–517
- absorbed power ( $P_a$ ) in, 113–114
- adjustments to, 433–438
- admittance in, 112, 114
- air-articulated line as, 62–63, **63**
- angle of reflection in, 75
- antinodes in, 87
- attenuation constant of, 82
- balanced transmission line and, 208
- cable TV, 67
- capacitance in, 64, 69
- characteristic impedance ( $Z_0$ ), 64–67, 89
- coaxial cable, 59, 61–62, **61, 62**, 66, 69, 208
- (*See also* coaxial cable)
- conductance in, 64
- current standing wave ratio (ISWR) in, 88
- cutoff frequency ( $F_c$ ) in, 69
- dielectric loss in, 369
- double-shielded coaxial as, 63, **63**
- transmission lines (*Cont.*):
- electrical vs. physical length of, 67–68
- equivalent circuit for, **64**, 516, **517**
- gas-filled line as, 62, **63**
- half-wavelength line as, 83
- hardline as, 62, **62**
- helical line as, 62
- impedance and impedance matching, 71, 72, 82–86, 109–112, **111**
- impedance matching network (Q section) in, 85
- incident power ( $P_i$ ) in, 89
- inductance in, 64
- installation tips for, 433–438
- insulators in, 61
- load impedance in, 109–110
- longwires and, 217–218
- “looking-in” impedance of, 82–86
- loss coefficient scale in, 112–113
- loss in, 83–84, 91–94, 369
- measurement and adjustment techniques for, 516–517, 525
- mismatch (VSWR) losses in, 91–94
- multiband and tunable-wire antennas and, feeding, 208–212
- nodes in, 87
- noise and, 68–69
- ohmic loss in, 369
- Ohm’s law and, 74
- open line as, 59
- parallel lines as, 59–63, **60**, 65–66
- phase constant of, 82
- propagation constant of, 82
- Q section in, 85
- quarter-wavelength lines as, 84–85
- quarter-wavelength shorted stubs in, 86, **87, 88**
- radiation loss in, 369
- reactance in, 85–86, 109, 525
- rectangular waveguide using parallel type, 370–373, **371**
- reflected power ( $P_r$ ) in, 89
- reflection coefficient in, 71–79, 89, 112, 114
- reflection loss in, 113–114
- resistance in, 64, 525
- responses in, 70–94
- return loss in, 113, 114
- RLC networks and, 59

- transmission lines (*Cont.*):  
 shielded parallel line as, 66  
 shielded twin lead as, 61, **61**  
 shorted lines and, 87  
 shorted stubs, 86, **87**, 114  
 (*See also* stub matching systems)  
 skin effect in, 369  
 Smith charts for, 109–114  
 standing wave ratio (SWR) in,  
 86–91, **90**  
 standing waves in, 71, **72**, **73**, 87  
 steady-state ac response in, 70,  
 79–82, **81**  
 step–function response in, 70–71,  
**70**, **71**  
 stripline or microstripline as, 63,  
**64**, 66–67  
 stub matching systems (*See* stub  
 matching systems)  
 surge impedance ( $Z_0$ ) in, 516  
 time domain reflectometry (TDR)  
 measurements in, 75–79, **76**,  
**77**, **78**, **79**, **80**, **81**  
 transmission loss coefficient  
 (TLC) in, 112, 114  
 troubleshooting tips for,  
 433–438  
 twin lead as, 59, **60**  
 velocity factor of, 67–68, 68*t*, 525  
 voltage minima and maxima in, 87  
 voltage standing wave ratio  
 (VSWR) in, 88–94, **90**, 112,  
 114, 517, **518**  
 waveguides vs., 86  
 wavelength calculation in, 110  
 transmission loss, using Smith  
 charts, 108–109  
 transmission loss coefficient (TLC),  
 112, 114  
 transmitters, 123  
 transmitting antennas, small loops  
 (*See* loop antennas, small  
 transmitter-type)  
 transverse electric mode (TE  
 mode) in microwave wave-  
 guides and antennas, 375  
 transverse electromagnetic (TEM)  
 fields, 9, 69, 373–375  
 transverse magnetic mode (TM  
 mode) in microwave wave-  
 guides and antennas, 375  
 trap dipoles, 204–206, **204**, **205**  
 traveling wave antennas, 220, 223  
 traveling waves, 127, 128–129, **129**,  
 134  
 trenching to apply chemical soil-  
 conductivity improvers, 576,  
**576**  
 triangle loop ( $1\lambda$ ) large antenna,  
 290–293, **291**  
 triangulation of radio signal, 302,  
 439  
 trifilar windings in transformers, 472  
 tropical band, Beverage or “wave”  
 antenna in, 224–228, **226**, **228**,  
**229**  
 troposphere, 14, **15**, 29, 51  
 tropospheric propagation, 17–19,  
**18**, 29–32  
 troubleshooting tips, 433–438  
 true longwire antenna, 217–219,  
**217**  
 tubing construction of vertical  
 dipole, 192–194, **193**  
 tunability, multiband and tunable-  
 wire antennas, 203  
 tunable dipoles, 160  
 tunable-wire antennas, 203–212  
 tuned circuit measurements, 526  
 tuned feeder antennas, 206, **207**  
 tuning, 524  
 antenna matching tuner in, 281,  
**285**  
 of ground wires, 580–581  
 of high–frequency dipoles and  
 doublets, 149–152  
 of longwires, 214, **215**, 215  
 of loop antennas, small receiver-  
 type, 304, 307–309, **308**,  
 312–313  
 of marine radio antennas,  
 497–500, **499**, **500**  
 of mobile HF antennas, 483–485  
 of multiband and tunable-wire  
 antennas, 203–205  
 of shortwave reception antennas,  
 281  
 tuning antennas, 203  
 tuning capacitance and capacitors,  
 loop antennas, small transmitter-  
 type, 320–321  
 turns in transformers, 474  
 TV receive-only (TVRO) satellite  
 dishes, 421  
 twenty–seven day cycle of radio  
 propagation, 46  
 twin lead, 59, **60**  
 two-meter Yagi VHF/UHF antenna,  
 352–353, **352**  
 two-port phasing transformer for  
 directional phased vertical  
 antennas, 249–250, **250**, **251**
- ## U
- U-bolt mounting hardware, 558,  
**559**  
 ultrahigh frequency (UHF), 13,  
 17–19, 23, 29, 32, 39, 61  
 in loop antennas, small receiver-  
 type, 302  
 in VHF/UHF transmitting/  
 receiving antennas, 339–368  
 ultraviolet (UV) radiation, 9, 39  
 umbra region, 17  
 unity resistance circle, Smith  
 charts, 100  
 Universal Coordinated Time (UTC),  
 55  
 upper sideband (USB), 55
- ## V
- varactor tuning, in loop antennas,  
 small receiver-type, 309, **310**  
 variation, ionospheric, 45–47  
 vee beam longwire antennas,  
 221–223, **222**  
 vee dipole, 236, **239**  
 vee-sloper antenna, 168–169, **168**  
 velocity factor ( $V_F$ ), 14, 525  
 of coax, 248  
 of microwave waveguides and  
 antennas, 377–381, **378**  
 of transmission lines, 67–68, 68*t*  
 velocity of propagation, 8, 13–14,  
 137–139, **139**  
 vertical antennas (*See* vertically  
 polarized HF antennas)  
 vertical coax VHF/UHF antenna,  
 346–349, **347**, **348**  
 vertical dipole scanner antenna,  
 364, **364**, **365**  
 vertical extent radiation pattern,  
 vertically polarized HF anten-  
 nas, 179, **180**  
 vertical plane radiation pattern,  
 high–frequency dipoles and  
 doublets, 148–149, **148**  
 vertical polarization, 12–13, **13**, 27,  
**28**, 55, 173

- vertical polarization (*Cont.*):  
 Beverage antenna, 226–227  
 helically wound antennas, **239**,  
 242, **243**  
 vertically polarized HF antennas,  
 173–202  
 vertically polarized HF antennas,  
 173–202, **174**  
 above-ground mounted, 197  
 angle of radiation in, 176–179, **178**  
 antenna tuning units (ATU) in, 173  
 coaxial vertical antennas in,  
 191–194, **192**, **193**  
 coaxial vertical scanner antenna  
 in, 365–368, **367**, **368**  
 coaxial-fed, **175**  
 collinear vertical VHF/UHF  
 antenna, 349, **350**  
 collinear vertical antennas in,  
 191–194, **192**  
 construction of, 195–197  
 current in, 173, **174**  
 delta-fed grounded vertical  
 antenna in, 200–201, **200**  
 directionally phased (*See*  
 directional phased vertical  
 antennas)  
 “drooping radials” in, 176, **177**  
 electric fields in, 173  
 electrical connections for, 197,  
**198**, 560–561  
 feedpoint circuits for, 197, **198**  
 fence posts as supports for verti-  
 cal antennas, 561–565, **563**,  
**564**  
 five-eighths-wavelength antenna,  
 181, **182**, 183, 197–200  
 four/4x4 support posts for, 565,  
**565**  
 gain in, 179–181  
 gamma-fed grounded vertical  
 antenna in, 200–201, **201**  
 ghost radiator in, 173  
 ground losses in, 183  
 ground plane vertical antennas in,  
 176, **177**  
 ground-mounted, 176, 195–197,  
**196**, 561–565, **562**  
 grounding and ground system for,  
 176, 185–187, **188**, 195–197,  
**196**, 200–201  
 half-slope vertical antenna, 238,  
**240**
- vertically polarized HF antennas  
 (*Cont.*):  
 half-wave vertical dipole in,  
 187–191, **190**  
 half-wavelength antenna, 181, **182**  
 height of antenna in, 181, 187  
 helically wound antennas, **239**,  
 242, **243**  
 image antenna in, 173  
 impedance and impedance match-  
 ing, 173, 175–176, 183,  
**184**, **186**, 195–197, 199–200,  
**199**  
 installation of, 560–572  
 insulators in, 175, 195  
 isotropic radiation in, 179–181,  
**181**  
 J-pole antennas in, 201  
 length of, 181, 183, **184**  
 limited-space designs, 238, **240**  
 losses in, 183  
 low-frequency operation antennas,  
 502–505, **503**  
 mast-mounted, **175**, 194–195,  
**194**, **546**, 560–561, **561**, **562**  
 mounting, 186, 189–190, 194–197,  
**194**, **196**, 560–565, **561**  
 non-ground-mounted, 195, 197  
 non-quarter-wavelength verticals  
 in, 181–185  
 omega-fed grounded vertical  
 antenna in, 200–201, **202**  
 omnidirectionality in, 173,  
 179–181  
 power calculation in, 180  
 Q-sections in, 199–200, **199**  
 quarter-wavelength antenna, 173,  
 181, **182**, 183, 187  
 radials used in, 176, 186, **189**,  
 560–561  
 radiation patterns in, 178–181,  
**181**, **182**  
 radiation resistance in, 183, **184**,  
**185**  
 reactance in, 183, **184**  
 resistance in, 183  
 safety issues, burns or shocks  
 from RF through vertical  
 antennas, 191, 195  
 series-fed vertical antennas in,  
 200  
 shortwave reception antennas,  
 279–281, **280**, **282**, **283**
- vertically polarized HF antennas  
 (*Cont.*):  
 shunt-fed vertical antennas in, 200  
 skip propagation and, 176–179,  
**178**  
 “spider web” ground system for,  
 185, **188**  
 tower installation, **538**, 565–572  
 transformers in, 199–200  
 tubing construction of vertical  
 dipole, 192–194, **193**  
 vertical coax VHF/UHF antenna,  
 346–349, **347**, **348**  
 vertical dipole scanner antenna in,  
 364, **364**, **365**  
 vertical extent radiation pattern  
 in, 179, 180  
 VHF/UHF transmitting/receiving  
 antennas, 340–341  
 voltage in, 173, **174**  
 voltage standing wave ratio  
 (VSWR) in, 176, 195  
 window-sill antennas, 238, **241**  
 Yagi antennas and, 201  
 very high frequency (VHF), 13,  
 17–19, 23, 29, 30, 32, 39, 41, 46,  
 54, 55, 170–171  
 emergency operations antennas  
 and, 487  
 loop antennas, small receiver-type  
 and, 302  
 marine radio antennas and, 495  
 vertically polarized HF antennas  
 and, 191  
 VHF/UHF transmitting/receiving  
 antennas and, 339–368  
 very low frequency (VLF), 29, 55,  
 61  
 Beverage or “wave” antenna in,  
 224–228, **226**, **228**, **229**  
 VHF/UHF transmitting/receiving  
 antennas, 339–368  
 baluns in, 341–346, **342**, **343**  
 bandwidth of, 339  
 beam antennas in, 349–352, **351**,  
**352**  
 “bowtie” dipoles in, **361**, 362  
 citizens’ band antennas as, 341  
 coaxial vertical antenna for,  
 346–349, **347**, **348**  
 coaxial vertical scanner antenna  
 in, 365–368, **367**, **368**  
 collinear vertical antenna, 349, **350**

- VHF/UHF transmitting/receiving antennas (*Cont.*):  
 delta feed match for, 343, **344**, 345  
 design limitations in, 340  
 directivity in, 349  
 drooping dipole scanner antenna in, 363, **363**  
 effective radiated power (ERP) in, 349  
 five-eighths-wavelength 2-meter antenna for, 353, **354**  
 folded dipoles in, 362, **362**  
 frequency spectrum of, 339, 341  
 gain in, 349  
 gamma match system for, 346, **346**  
 groundplane antenna for, 356, **357**  
 groundplane scanner antenna in, 364, **366**  
 halo antenna for, 356, **358**  
 impedance matching in, 341–346, **342**  
 J-pole antenna for, 353–355, **355**  
 length of, 339  
 line-of-sight communications and, 339  
 longwires as, 340–341  
 low frequency (LF) vs., 339  
 lower band antennas on VHF/UHF, 340–341  
 mounting, 347  
 OSCAR satellite communications and, 339  
 quad beam antenna for, 356, 358, **359**  
 safety issues in, 340  
 satellite communications and, 339  
 scanner band antennas, 360–368  
 scanner skyhook antennas in, 362–368, **363**  
 scanner-vision antennas in, 360–362, **361**  
 size of, as limiting factor, 340  
 skip propagation in, 339  
 stub matching system for, 345–346, **345**  
 transformers in, 341–346, **342**, **343**  
 two-meter Yagi antenna, 352–353, **352**  
 vertical antennas as, 340–341  
 vertical dipole scanner antenna in, 364, **364**, **365**
- VHF/UHF transmitting/receiving antennas (*Cont.*):  
 voltage standing wave ratio (VSWR) in, 341  
 wavelength in, 339  
 Yagi antennas in, 349–352, **351**, **352**  
 virtual height, 43, **44**, **45**  
 voltage, 123–120, 131–137, **132**, **133**, 531  
 (*See also* voltage standing wave ratio)  
 in high-frequency dipoles and doublets, 143–146, **143**  
 lag, 126  
 in loop antennas, small receiver-type, 303, 304  
 in loop antennas, small transmitter-type, 320, 323  
 minima and maxima of (*See* antinodes; nodes)  
 standing waves of, 134–136, **135**, **136**, 139  
 in vertically polarized HF antennas in, 173, **174**  
 voltage standing wave ratio (VSWR), 88–91, **90**, 515, 517–519, 529, 532  
 calculation of, 517–519  
 EZNEC for Windows modeling of, 337, **338**  
 frequency vs., 434–437, **435**  
 in hidden and limited-space antennas, 238  
 in high-frequency dipoles and doublets in, 144, 146, 149–151, **150**, 160  
 in longwires, 215  
 in loop antennas, small transmitter-type, 320  
 measurement of, 433, 435–437, **436**, 529, 532  
 in microwave waveguides and antennas, 388  
 in mobile HF antennas, 483  
 in multiband and tunable-wire antennas, 203  
 NECWin Basic modeling of, 333, **335**  
 using Smith charts circle, 100, **104**, 112, 114, 115, 117  
 in stub matching systems, 115, 117  
 in transmission lines, 91–94, 112, 114, 517, **518**
- voltage (*Cont.*):  
 in vertically polarized HF antennas, 195  
 in VHF/UHF transmitting/receiving antennas, 341  
 in Yagi antennas, 261
- W**  
 water analogy of radio wave propagation, 6–7  
 Watson-Watt Adcock array antennas, 447–450, **449**, **450**  
 wattmeters, 529–539, **531**, **533**, **534**, **535**, **536**, **537**, **538**  
 waveguides, 86, 369–415  
 (*See also* microwave waveguides and antennas)  
 wavelength calculation, using Smith charts, **103**, 105, 110, 115  
 wavelength, 7–8, 28  
 high-frequency dipoles and doublets, 141–142  
 longwires, 217, 218–219  
 microwave waveguides and antennas, 377–381, **378**  
 VHF/UHF transmitting/receiving antennas, 339  
 “wave” antenna, 224–228, **226**, **228**, **229**  
 weather-related propagation phenomena, 28, 36  
 wedge dummy loads, in microwave waveguides and antennas, 385, **386**, **387**  
 Wheatstone bridge, 519–521, **520**  
 whip antennas for marine radio, 496–497, **497**  
 wide-aperture direction finder (WADF), 451  
 windings in transformers, 472–473, **473**  
 Windom antenna, 209–210, **211**  
 Windom, Loren, 209  
 window-sill antennas, 238, **241**  
 wire antenna, basic design, 548, **548**  
 wire used in longwires, 214  
 wireless telegraphy, 1–2  
 Wolf Number, 40  
 World Administrative Radio Conferences (WARC), 2

Wullenweber array antennas, 451,  
**452**  
WWV radio station, 40, 47  
WWVH radio station, 40

## **X**

X bands, 69  
x-ray radiation, 39

## **Y**

Yagi antennas, 203, 257–262, **259**,  
421  
capacitors in, 261  
directors in, 258–259  
feedpoints, 260–261, **261**

Yagi antennas (*Cont.*):  
gamma match feed system for,  
260–261, **261**  
impedance matching in, 260–261,  
**261**  
length of, 259–260  
noise and VSWR in, 261  
parasitic elements in, 258–259  
quad beam antennas vs., 265  
radiation pattern of, 257–258, **258**  
reflectors in, 258–259  
two-meter Yagi VHF/UHF antenna,  
352–353, **352**  
vertically polarized HF antennas,  
201

Yagi antennas (*Cont.*):  
VHF/UHF transmitting/receiving  
antennas, 349–352, **351**,  
**352**  
Yagi-Uda antenna, 258

## **Z**

Zepp antenna, 164–165, **165**,  
171  
end-fed multiband, 208, **208**  
zipcord, 275  
ZL-special beam antenna, 20-m,  
262–263, **262**  
Zurich Smoothed Sunspot  
Number, 40

# About the Author

Joe Carr was a military electronics technician and author. Along with the best-selling *Practical Antenna Handbook*, his other books with McGraw-Hill include *Secrets of RF Circuit Design*. Carr also contributed a monthly column to *Nuts & Volts* magazine entitled “Open Channel.”

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