This book is dedicated to my wife Beverly Benson Long, for our enduring love and her world-wide work in the promotion of mental health and the prevention of mental disorders.
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The publisher has posted a list of supplemental materials (SM) to an accompanying website at:

www.scitechpub.com/blakelong3.htm

Within the text of the book you will find references to the specific SM sections that relate to the material being covered. A computer icon ( ) is used in the margin to further identify sections that refer to the SM.

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SM 5.0 Three-Dimensional Pattern Construction, by Aaron Loggins
This is a senior undergraduate or first-year graduate level textbook on antenna fundamentals, design, performance analysis, and measurements. In addition to its use as a formal course textbook, it is well-suited for professional training and self-study by practicing engineers, scientists, and technologists who desire to expand their knowledge of antennas. The book provides a broad coverage of antenna types and phenomena, for operations at very low radio frequencies, as well as frequencies up to those of submillimeter wavelengths. Unlike most university-level antenna textbooks, reading it does not require prior skills in electromagnetic theory, sophisticated mathematics, or computer programming. An additional feature is the downloadable collection of computer solutions in both Mathcad® and MATLAB® to numerous antenna radiation examples, which can be easily implemented and revised by persons not having prior programming experience.

Evolution of the Third Edition

This new edition was prepared for use in a one-semester first-year graduate night class at the Southern Polytechnic State University located in Marietta, Georgia, where student backgrounds vary widely. At least half the students are from overseas and hold a bachelor’s degree in electrical engineering or electronics. Of those from the United States, about half have degrees in electrical engineering, and the remainder hold degrees in electrical engineering technology. Generally, the students are older than a typical first-year graduate student, being 25 to 35 years of age. A few of the students have excellent backgrounds in vector calculus and the use of Maxwell’s equations, while some of the older ones may need to refresh their abilities with the phasor calculations of electrical circuits.

The computer capabilities of the students also vary widely. Many of the entering graduate students are proficient in use of MATLAB or Mathcad software, but sometimes they are experienced with neither and rarely both. Additionally, some students who have been away from academics a few years may have little computer proficiency. Thus, to expedite the class learning on antennas with computer analyses, Mathcad, which is easily read without prior experience, is used in classroom lectures. For class assignments, students are allowed to use whatever computer software they choose. Mathcad software
is used most often, with MATLAB being the second most popular choice. Often, original
MATLAB users, after switching to Mathcad, have expressed appreciation for their intro-
duction to Mathcad for its relative ease of use and intuitive qualities.

Because of the wide differences in student backgrounds, and after considering the
available textbooks, Blake’s Antennas, Second Edition, was selected for adoption because
of its superior readability. Due to the book’s age, the selection required preparing
and distributing materials for updating and expanding the text and adding appendices.
Consequently, the present book evolved into one that retains the benefits of Blake’s
second edition but expands the subject material suitably for a senior or graduate level
textbook.

Background Assumed

Most antenna textbooks are written for students proficient with vector calculus and begin
with the use of Maxwell’s equations in the development of antenna theory. Such books
often do not meet the needs of many students and practicing engineers who, because of
their backgrounds or personal interests, desire a more direct path for assimilating antenna
fundamentals and their connection to application topics of antenna engineering. Although
antenna theory is founded on Maxwell’s equations, understanding their concepts does
not require advanced mathematics. At the beginning of each antenna course, the revising
author (MWL) uses Appendix A, Maxwell’s Equations, to address the key “postulates”
of Maxwell and provide a brief introduction to, or review of, the essential equations.
Thus, Maxwell equations are discussed with the goal of expressing their meaning in
words. Then, the concepts of displacement current, interdependence of changing electric
and magnetic fields, and wave propagation are described, and thus Maxwell’s equations
are underscored as “the ultimate truth” but thereafter considered outside the scopes of
antenna design, performance analysis, and measurements.

Organization

This book was prepared with the intention of providing a comprehensive antenna text
that can be readily understood by persons with undergraduate educations in engineering,
science, or technology. The chapter titles follow:

  Chapter 1. Electromagnetic Waves
  Chapter 2. Transmission Lines
  Chapter 3. Antenna Parameters
  Chapter 4. Basic Radiators and Feed Methods
  Chapter 5. Arrays
  Chapter 6. Reflectors and Lenses
  Chapter 7. Antennas with Special Properties
  Chapter 8. Electronically Steered Arrays
  Chapter 9. Antenna Measurements
Chapters 1 through 6 cover, generally, the physics and technology of antennas and include such subjects as wave propagation, reflection, refraction, diffraction, transmission and reception, basic radiators, antenna arrays, reflector antennas, and lenses.

Chapter 7 discusses antenna properties and analysis techniques not addressed in other chapters. Its range of topics is wide, and includes techniques for providing wide bandwidths, multiple polarizations, low receiver noise, and extremely low sidelobes. In addition, direction-finding antennas and mechanical beam scanners are addressed. Finally discussed are synthetic-aperture antennas, geometrical theory of diffraction (GTD), method of moments (MoM), and fractals.

Chapter 8 treats electronically steered arrays, whereas Chapter 5 is focused on fixed beam arrays. In other words, chapter 8 stresses array concepts specific to beam movement made possible with fast, wide-dynamic-range digital components and cheap computer memory, along with continued improvements in high-speed switches and phase shifters.

Chapter 9 includes a broad coverage of antenna measurement techniques and equipment. Subjects include radiating near fields as well as far field patterns and pattern statistics, compact ranges, and near-field measurements. Included also is a comprehensive treatment of antenna noise, noise temperature, noise figure, and system signal-to-noise ratios.

There are problems at the end of each chapter, and answers to the odd numbered problems are included in a section near the book’s end. Appendices provide technical depth to the chapters, appropriate for a senior or first graduate level antenna course. The appendix titles follow:

Appendix A. Maxwell’s Equations
Appendix B. Polarization Theory
Appendix C. Review of Complex-Variable Algebra
Appendix D. Complex Reflection Coefficients and Multipath Effects
Appendix E. Radomes
Appendix F. Far-Zone Range-Approximation and Phase Error
Appendix G. Radiating Near and Far Fields, and the Obliquity Factor
Appendix H. Path Length Differences from a Planar Aperture
Appendix I. Effects of Random Aperture Phase Errors

It is to be noted that Appendix C discusses complex-variable algebra. Although its contents will be familiar to most readers, it is included because some may find parts of it useful for review.

Data files of computer scripts

Where appropriate, the appendices and the downloadable data files are referenced in the chapters for providing a more complete treatment of antennas. In the Deluxe Edition, a full-featured copy of Mathcad 14.0 is included so that readers can easily create their own computer analyses. The downloadable data files provide computer solutions in both Mathcad and MATLAB to problems in the areas that follow:
- reflection coefficients for surfaces versus dielectric properties, conductivity, polarization, surface roughness, and incidence angle
- earth’s multipath effects on antenna patterns versus surface properties, antenna and observation heights and separation distance, and polarization for flat and spherical earth models.
- radiating near and far fields from arrays and continuous aperture antennas, as functions of aperture phase and amplitude distributions and random aperture errors.

The files also include a supplemental chapter in PDF on the creation of antenna radiation field graphics using Mathcad. It was prepared by student Aaron Loggins as one of three project assignments in a one-semester antenna course.

Files can be downloaded from the publisher’s web page for this book: www.scitechpub.com/blakelong3.htm

Acknowledgements

Permissions to use the contents of Antennas, 2nd Edition, by Lamont V. Blake, now deceased, were provided by Barbara Blake, Lamont Blake’s daughter, and other Blake family members and are gratefully appreciated. This third edition could not have been written otherwise because it was built upon an easily read, well-written text based on a solid technical foundation. Therefore, it could be readily expanded to provide a senior or graduate level textbook suitable for students with widely different academic backgrounds, including persons with limited or no computer programming experience.

Two important and closely related tasks were accomplished by Dr. Donald G. Bodnar in connection with “Chapter 9 – Measurements.” First, he completed a technical review of an early version of the chapter, and he then wrote Sec. 9.4, a major section titled “Near Field Antenna Measurements.” That section is copyrighted by MI Technologies, Inc., Don Bodnar’s company.

Appreciation is acknowledged to Aaron Loggins for letting me use his classroom project paper as a PDF file that discusses the creation of 3-D graphics with Mathcad.

A major and generally thankless task of pursuing a penetrating technical edit of each chapter and appendix was accomplished by Dr. Edward B. Joy, and it was performed with record-breaking speed. Ed found and corrected not only accidental and careless errors, but he also underscored and made suggestions for correcting more substantive oversights.

Special thanks are due to Dr. Anatoliy Boryssenko of the University of Massachusetts for his expertise in checking the Mathcad files, offering helpful suggestions, and then rewriting them into MATLAB scripts. He did so under very tight deadlines. Dr. Boryssenko has also graciously offered additional files from his personal collection to further enhance the data set of the publisher’s web page.

A major contribution to this book was made by Dr. Randy J. Jost of Utah State University, as an advisor to SciTech Publishing, by reviewing and suggesting additions to my early book writing plans. One of those suggestions was to include the files that contain
a number of antenna radiation problems and their computer solutions. Inclusion of the CD in the Deluxe Edition that contains Mathcad, version 14, software results from the initiative of Dudley Kay, Founder of SciTech Publishing, and the cooperation of Parametric Technology Corporation, the owner of Mathcad. Permission by Parametric Technology Corporation to use screenshots of computer images from Mathcad software, included in the file on creating 3-D graphics is gratefully acknowledged.

There have been a number of persons who have made significant editorial improvements and others who have simply expressed an interest in an updated edition of Lamont Blake’s *Antennas* becoming available. Some of these include Gerald Oortman of Lockheed Martin, Marietta, Georgia; Professor Charles Bachman of the Southern Polytechnic State University; Dr. Andrew Peterson of the School of Electrical and Computer Engineering, Georgia Institute of Technology; James Gitre of Motorola; Michael Havrilla of the Air Force Institute of Technology; and Rickey Cotton (deceased), Mark Mitchell, and Dr. Charles Ryan (retired) of Georgia Tech Research Institute.

I thank Phyllis Hinton of Georgia Tech Research Institute, who has, over the years, brightened my days when she sketches a figure I need or somehow helps me find my way through the ends and outs of Microsoft Word.

Appreciation is expressed here to Dudley Kay, Susan Manning, and Robert Lawless of SciTech Publishing who, during the preparation of this book, have demonstrated an enthusiasm for producing quality textbooks.

Maurice W. Long
Atlanta, Georgia
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CHAPTER 4

Basic Radiators and Feed Methods

The preceding chapters have been largely groundwork, covering principles, concepts, and terminology, as well as some basic antenna theory. It is now possible to discuss specific antennas.

Initially the basic forms of radiating structures will be discussed. They are sometimes used by themselves as simple antennas; they may be very effective in some applications. They are also used in combination with each other and with other components to form more complicated antennas having special properties such as high directivity, large bandwidth, omnidirectionality, steerable beams, low noise, and so on. These more complicated antennas are described in chapters 5 through 8. The properties of the basic radiators will be discussed in terms of their behavior in free space, except for specific types that require the presence of the earth for their operation, as will be indicated in the discussion. It will also be assumed, except where specifically otherwise stated, that current-carrying portions of the antennas are perfect conductors. The conductors of actual antennas are usually very good so that this assumption is reasonably well approximated in practice. When some conductor losses do occur, the principal effect is to reduce the efficiency, as defined by equation (3-30), chapter 3. Ordinarily the pattern (and hence the directivity and beamwidth) are not seriously affected, although appreciable conductor resistance would affect these parameters also.

4.1. Short Dipoles

Perhaps the simplest and most important radiating structure, from the viewpoint of antenna theory, is the electrically short dipole with uniform current along its length. "Electrically short" means "short compared to a half wavelength."

Throughout this book, the term short dipole is understood to mean simply a dipole that is much shorter than a half wavelength, but not necessarily one with a uniform current. The term is generally applied to any dipole that is no longer than about a tenth of a wavelength (0.1λ). Isolated short dipoles do not ordinarily have uniform current throughout their length, but approximate uniformity of the current may be achieved by capacitive end loading, as will be described later in this section.
A short dipole that does have a uniform current will be called an elemental dipole. Such a dipole will usually be considerably shorter than the tenth-wavelength maximum specified for a short dipole. The term infinitesimal dipole may be used to imply extreme shortness, as required in certain mathematical analyses. Other terms sometimes used for an elemental dipole are elementary dipole, elementary doublet, and Hertzian dipole. Part of the importance of this type of dipole is that many more complicated antennas can be analyzed by considering them to be assemblages of many elemental dipoles. For example, a long-wire antenna may be regarded as composed of many elemental dipoles connected end-to-end. Although the current is considered constant (at any instant of time) along the length of each elemental dipole, the currents in different dipoles may be different, both in magnitude and phase; therefore, a nonuniform current in the long wire can be approximated by this representation.

If a center-fed short dipole is initially in a neutral condition and then a current starts to flow in one direction, one half of the dipole will acquire an excess of charge and the other a deficit (since a current is a flow of electrical charge). There will then be a voltage between the two halves of the dipole. If the current then reverses its direction, this charge unbalance will be first neutralized and then reversed. Therefore an oscillating current will result in an oscillating voltage as well (or vice versa). If the current oscillation is sinusoidal, the voltage oscillation will also be sinusoidal and approximately 90 degrees lagging the current in phase angle; that is, the short dipole is capacitive in nature, from the viewpoint of its current-voltage relationship.

Since such a dipole can be regarded as one in which electric charge oscillates, it is an oscillating electric dipole. It is distinguished from an oscillating magnetic dipole, which is equivalent to a bar magnet whose magnetic strength and polarity oscillate. An example of a magnetic dipole is discussed in sec. 4.5.1.

The current, though uniform throughout the length of the elemental dipole at any instant, is assumed to vary sinusoidally in time, according to (4–1) that follows:

\[ I(t) = I_0 \sin(2\pi ft + \alpha) \]  

where \( I(t) \) is the current at any time \( t \), \( I_0 \) is the amplitude of the current (the rf peak value), \( f \) is the frequency in Hertz, \( t \) is the time in seconds, and \( \alpha \) is the phase angle (which simply means that when \( t = 0 \), \( I(t) = I_0 \sin \alpha \)).

4.1.1. Dipole Radiation

It is a well-known fact of elementary electrical theory that current in a wire is accompanied by a magnetic field surrounding the wire, and also that a voltage existing between two conductors, or different parts of a conductor, is associated with an electric field in the intervening and adjoining space. A dipole antenna, therefore, will be surrounded by an electric field and a magnetic field. The nature of these fields at an instant of time in the vicinity of the dipole is indicated in Fig. 4–1.

The current and voltage of an rf dipole change direction (polarity) at a rate determined by the frequency \( f \) and undergo a sinusoidal variation of their magnitudes as shown (for
the current) by (4–1). At a given instant, however, they have a particular direction, and the field lines have a corresponding direction, as shown in Fig. 4–1. When the current direction reverses, the direction of the magnetic field lines reverses also; and when the voltage polarity reverses, the electric field reverses direction.

The field lines are shown only in the immediate vicinity of the dipole, and only a few lines are shown. In principle, however, they fill the entire space around the dipole and would extend to an infinite distance from it if the dipole were located in empty space and if the current had been flowing in it for an infinite time. But the fields do not exist at an infinite distance as soon as the current starts flowing. They travel outward from the dipole at the finite speed $c = 3 \times 10^8$ meters per second (in empty space and at practically the same speed in air).

Therefore, by the time the field lines corresponding to a given polarity of the dipole have reached a distance from the dipole equal to $\frac{1}{2}(c/f)$, which is a half wavelength as defined by equation (1–1) of chapter 1, the dipole polarity has reversed, and the new field lines being set up in the region immediately adjoining the dipole have directions opposite to those at the distance $\frac{1}{2}(c/f) = \frac{\lambda}{2}$. As the oscillation of current and voltage in the dipole continues, the outward-traveling field lines will evidently have opposite directions at half-wavelength intervals along the direction of travel. This oscillating outward-traveling field is an electromagnetic wave, which has been radiated by the dipole. The approximate configuration of the electric-field component of this wave, in a region extending several half wavelengths from the dipole, is shown in Fig. 4–2a, and the magnetic-field component is shown in Fig. 4–2b.

It is noteworthy that some of the electric field lines, rather than terminating on charges in the conductors, form closed loops in space—something that cannot occur in the electrostatic case (where the field is due to a steady or d-c voltage). This field configuration, resulting in radiation, is an electrodynamic phenomenon; it happens only with varying currents and voltages, and with moving fields. At much greater distances than those shown in Fig. 4–2, however, the “closed ends” of the loops virtually disappear, leaving only field lines that are transverse to the direction of propagation of the waves, as in Fig. 1–1, chapter 1.

Not all the fields in the vicinity of the dipole represent radiation. When the dipole polarity reverses, some of these field components “collapse” upon the conductor, causing induced voltage and current associated with the inductance and capacitance of the dipole conductor. These are the field components that decrease rapidly with distance from the dipole, as discussed in sec. 3.2.5. One component, the static field, decreases as the inverse cube ($1/r^3$) of the distance $r$. A second component, the time varying reactive field, decreases as the inverse square ($1/r^2$) of the distance. These two components comprise the
near field of the dipole. The third component, which decreases as the inverse first power of the distance (1/r), is the radiation field. Because of their more rapid decrease with distance, the static and reactive fields become very much smaller than the radiation field at great distances from the dipole; in fact, at a distance of a few wavelengths, the radiation field is so very much the strongest component that the near-field components are virtually negligible. Close to the dipole, however, the near-field components are much stronger than the radiation components.

The existence of a radiation field was unnoticed in the earliest experiments with electricity, because of the dominating reactive fields at short distances. Its existence was first demonstrated about 1887 by the classical experiments of the German physicist Heinrich Hertz, though it had been predicted theoretically about 25 years earlier by the English mathematical physicist James Clerk Maxwell (see Appendix A).

Electric and magnetic fields represent energy. The static and reactive fields represent stored energy that is returned to the conductor, just as the fields associated with coils and condensers of an electric circuit return energy to the circuit when the current or voltage maintaining the field is removed. The radiated fields, on the other hand, represent electromagnetic energy flowing outward into space. From the viewpoint of the dipole conductor, radiated energy is “lost” in the same sense that electrical energy is lost when it is converted into heat energy in a resistance. As mentioned in sec. 3.7, this energy loss is ascribed to a hypothetical “radiation resistance,” which is a helpful concept for theoretical considerations involving the antenna input impedance, efficiency, and radiated power.

FIGURE 4–2.
Configuration of electric and magnetic field lines in radiation field of an elemental-dipole radiator. (a) Electric field. (b) Magnetic field. Electric field lines are shown in a plane containing the dipole axis, and magnetic field lines are shown in the plane through the center of the dipole perpendicular to its axis. (Dipole is perpendicular to plane of paper in Figure 4–2b.)
In the radiation field there is a continual exchange of energy between the electric and magnetic field components. As part of his theory, Maxwell postulated that an electric field acting in empty space causes a “displacement current” just as it causes an electron current in a conductor. He further hypothesized that this displacement current has the same ability to set up a magnetic field as does an electron current. Many experiments confirm this theory, which is now unquestioned. A varying electric field therefore causes a varying displacement current, which in turn results in a varying magnetic field. The magnetic field, in its turn, creates a varying electric field in accordance with Faraday’s law. The cycle is thus complete; each field component sustains the other, and in regions remote from conductors or charges (empty space), one cannot exist without the other, in the fixed ratio given by equation (1-10), chapter 1.

That electric and magnetic fields represent energy may be demonstrated by placing a charged object in an electromagnetic field. The field will exert a force upon such an object and cause it to move if there are no restraining forces. This effect occurs, for example, when radio waves encounter the earth’s ionosphere, in which there are free electrons. These electrons are accelerated by the field, which means that kinetic energy is imparted to them. The energy is supplied by the electromagnetic field. Similarly, an electromagnetic field impinging on a receiving antenna transfers energy (power) to it, which may be amplified by the receiver and eventually converted into sound variations in a loudspeaker, light variations on a television screen, or some other form of intelligence.

### 4.1.2. Pattern of an Elemental Dipole

If an elemental dipole (length < 0.1λ) is placed at the origin of a spherical-coordinate system (Fig. 3-1) with its axis (direction of its length) parallel to the θ = 0 direction (the z-axis of the corresponding cartesian-coordinate system), the radiated field may be considered for any point in space whose position is given by the coordinates r, θ, φ; and the electric field intensity will then be designated \( E(r, \theta, \phi) \). The geometry of this situation is shown in Fig. 4-3.

Analysis based on Maxwell’s equations shows that the radiation field (far field) of this radiator is given by

\[
E(r, \theta, \phi) = \frac{60\pi l \sin \theta}{\lambda r} \quad (4-2)
\]

where \( l \) is the length of the elemental dipole and \( I \) is the dipole current, in amperes. If \( l \), \( \lambda \), and \( r \) are in meters, and \( I \) is in rms amperes, \( E \) is obtained from this formula in rms volts per meter.

The fact that the angle \( \phi \) does not appear in this expression means that for

**FIGURE 4-3.** Relationship of elemental dipole radiator to spherical coordinate system.
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The magnitude of \( E \), given by (4–2), and its direction in space (polarization) are two properties of a radiation field that must be specified for a description of the field. A third quantity also to be specified is the relative phase of \( E \) as a function of \( \theta \) and \( \phi \) for a fixed value of \( r \). For the elemental dipole field the phase is constant for a fixed value of \( r \); that is, it does not vary as \( \theta \) and \( \phi \) are varied. The absolute phase of \( E \) may be expressed as a function of the distance \( r \) and the phase of the current \( I \). Relative to the phase of \( I \), the phase angle of the electric field \( E \) at distance \( r \) is

\[
\gamma = \frac{\pi}{2} - \frac{2\pi r}{\lambda} = \frac{\pi}{2} \left(1 - \frac{4r}{\lambda}\right) \text{ radians} \quad (4-3)
\]

Note that \(-2\pi/\lambda\) is the delay of phase, in radians, created by the separation distance \( r \). Equation (4–3), as well as (4–2), applies only in the far field, which means at values of \( r \) are of the order of \( \lambda \) or greater.

**FIGURE 4–4.** Relative electric-intensity pattern of elemental (or short) dipole in a plane perpendicular to the dipole axis. \( (E = |\sin \theta|) \). (Note: The range of the angle \( \theta \) in this diagram is \( 0–360^\circ \), to indicate that the pattern is shown in a whole plane. Strictly, however, since the spherical colatitude coordinate \( \theta \) has a maximum value of \( 180^\circ \), the angles on the left side of the diagram should go from \( 0 \) to \( 180^\circ \) like those on the right side.)

The dipole center perpendicular to the dipole axis. At any general field point the polarization is by definition the direction of the electric field lines. These lines, in the radiation field, are always perpendicular to a line from the dipole center to the field point, and they lie in a plane containing this line and the dipole axis.
The relative-power-density pattern may be obtained from (4–2) by making use of the relation \( p = \frac{E^2}{377} \) from (1–9). (It is useful here to note that the number 377 actually represents the quantity \( 120\pi \).) The result is

\[
p(r, \theta, \phi) = \frac{30\pi l^2 |l^2\sin^2 \theta}{\lambda^2 r^2} \tag{4–4}
\]

This formula gives \( p \) in watts per square meter when \( I \) is in rms amperes and all lengths are in meters.

4.1.3. Radiation Resistance

The total power radiated \( P_{\text{total}} \) by an elemental dipole is found by integrating \( p(r, \theta, \phi) \) over the surface of an imaginary sphere at a fixed value of \( r \). Then, in accordance with equation (3–15), chapter 3

\[
P_{\text{total}} = \int_0^{2\pi} \int_0^\pi [p(r, \theta, \phi)] r^2 \sin \theta \, d\theta \, d\phi = \frac{30\pi l^2 |l^2|}{\lambda^2} \int_0^{2\pi} \int_0^\pi \left[ \frac{\sin^2 \theta}{r^2} \right] r^2 \sin \theta \, d\theta \, d\phi
\]

\[
= \frac{60\pi^2 l^2 |l^2|}{\lambda^2} \int_0^\pi \sin^3 \theta \, d\theta = \frac{790 |l^2|}{\lambda^2} \text{ watts} \tag{4–5}
\]

The radiation resistance may now be calculated, from (3–28), by dividing \( P_{\text{total}} \) by \( l^2 \). The result is

\[
R_r = 790 \frac{|l^2|}{\lambda^2} \text{ ohms} \tag{4–6}
\]

For example, if \( l/\lambda = 0.1 \), \( R_r = 7.9 \) ohms. (It will be recalled that this value of \( l/\lambda \) was stipulated at the beginning of this section to be approximately the maximum permissible length for an antenna to qualify as an elemental dipole.) The dipole current required for one kilowatt of power with this value of radiation resistance is 11.3 amp (a rather large current). This result indicates a major disadvantage of short dipoles, namely, the very large currents required for radiation of appreciable power. Therefore, radiators with larger values of radiation resistance are preferred when they can be used; but this is not always possible, and short dipoles are very useful radiators under these circumstances.

4.1.4. Directivity

The directivity of an elemental dipole can be computed by using foregoing results and the fundamental definition of directivity, equation (3–16), that follows:

\[
D = U_{\text{max}} / U_{\text{av}} = p(r, \theta, \phi)_{\text{max}} / p(r, \theta, \phi)_{\text{av}}
\]
providing \( p(r, \theta, \phi)_{\text{max}} \) and \( p(r, \theta, \phi)_{\text{av}} \) are evaluated at the same distance \( r \). The numerator of (3–16), the equation above, is available from (4–4) with \( \theta = 90^\circ \), so that \( \sin^2 \theta = 1 \) (the maximum value). The denominator is \( P_{\text{total}}/4\pi r^2 \), where \( 4\pi r^2 \) is the area of a sphere of radius \( r \) and \( P_{\text{total}} \) is given by (4–5). Then, complete expression for the directivity (with 790 written as \( 80\pi^2 \)) is

\[
D_{\text{Il}}(r) = \left( \frac{30\pi l^2}{\lambda^2 r^2} \right) \left( \frac{4\pi r^2 \lambda^2}{80\pi^2 l^2} \right) = 1.5
\]

(4–7)

Thus the directive is 1.5 regardless of the exact length \( l \) in relation to the wavelength \( \lambda \), as long as the radiator qualifies as an elemental dipole. In fact, this result applies to any short dipole.

4.1.5. Beamwidth

The pattern of Fig. 4–4 (the two circles formed by heavy lines) is created by a “slice” through the three-dimensional pattern. However, the total pattern in space is shaped like a doughnut (with a pin-sized hole). The two lobes of Fig. 4–4 are really cross sections of the same lobe, and the angular width of this doughnut-shaped lobe will now be discussed.

The half-power beamwidth may be determined either by measuring the angular width of the pattern, as plotted in Fig. 4–4, between the points of electric intensity equal to 0.707 \( E_{\text{max}} \), or from analysis of (4–4). The only angle-dependent term in this expression is \( \sin^2 \theta \), and it determines the pattern and the beamwidth. The quantity \( \sin^2 \theta \) has its maximum value of unity at \( \theta = 90^\circ \). Therefore, the half-power beamwidth is determined by finding the values of \( \theta \) where \( \sin^2 \theta \) equals \( \frac{1}{2} \). The values are \( \theta = 45^\circ \) and \( \theta = 135^\circ \) and thus the half-power beamwidth is \( 135^\circ - 45^\circ = 90^\circ \).

4.1.6. Input Impedance

The input impedance of a short dipole, when it is fed at a small gap near its center, consists of the radiation resistance in series with a large value of capacitive reactance (equivalent to a small series capacitance). There is also, in series, an equivalent loss resistance \( R_0 \), included previously in (3–26), that may be large enough to be significant in relation to \( R_r \). An equivalent circuit of the dipole, as it looks from the viewpoint of the transmission line, is shown in Fig. 4–5. Of course there is not actually a “series condenser” in the dipole; this circuit merely represents the impedance in a lumped-circuit equivalent form. Because the value of the equivalent series capacitance is very small, it represents a high value of capacitive reactance, which must be “tuned out” by including an inductive reactance of equal value in the feed circuit. Then the voltage supplied by the transmission line need not be extremely high. The current must be high, however, in order to radiate appreciable power, as indicated previously by (3–25). (In effect this equation shows that the radiated power is \( P = I^2 R_r \); so if \( R_r \) is small, \( I \) must be large to make \( P \) large.)
This high current, flowing through the inductive and capacitive reactances, produces very high voltages across the feed-circuit inductor and across the antenna input terminals, even though the transmitter itself does not have to supply a high rf voltage (because the inductive and capacitive reactances cancel, leaving only $R_r$ and $R_0$ as the effective transmitter load). If the input resistance $R_i (= R_0 + R_r)$ is very small, a more complicated arrangement than that shown in Fig. 4–6 may be required to provide an impedance transformation in addition to reactance cancellation. Therefore feeding a short dipole antenna is somewhat difficult. In particular, the inductance in the feed circuit, when high power is being radiated, must be very large both to withstand the high voltage and to carry the heavy current without excessive loss. These are expensive requirements.

4.1.7. Short Dipole with Nonuniform Current Distribution

An isolated elemental dipole cannot actually be achieved as a physical reality; it is more of a concept, like the isotrope, than a practical radiator. Some types of dipole radiators do exist that approximate the properties of the elemental dipole. If an actual short dipole is fed in the manner shown in Fig. 4–5b, current will flow in the dipole conductor, and radiation will occur. The current in the conductor, however, will not be uniform as assumed for an elemental dipole, and therefore equations (4–2) to (4–6) will not apply to its radiation without modification. The necessary modification turns out to be very simple.

Because the short dipole conductor is “open” at both ends, that is, not connected to anything, the current at these points must be zero. (Physical theory exists to support this statement, although here it will simply be assumed to be intuitively obvious that a current
cannot exist where it has “no place to go.”) At the same time, current can exist elsewhere in the dipole. This statement would seem to violate the physical principle of continuity for electric current, which implies that a current must have the same value everywhere along a continuous conductor. It will be recalled that this is not true in a transmission line on which a standing wave exists (sec. 2.1). The current in the dipole is a standing wave. It may be thought of as two equal and opposite currents (the reverse current is due to reflection from the open end) that, at the ends of the dipole, have exactly opposite phases and thus cancel, resulting in zero current. The cancellation is incomplete (because of the changing phase relationship) at points a short distance from the end and becomes progressively less complete going from the ends to the center. The principle of continuity is satisfied for the currents traveling in each direction, separately.

When the dipole is very short compared with a wavelength, as assumed, the current will vary approximately linearly along the dipole from end to center. This means that a graph representing the current as ordinate $I$, and distance from the end to the center of the dipole as abscissa $x$, is a pair of sloping straight lines forming a triangle, as shown in Fig. 4–7, with zero value at the left end (end of the dipole, $x = -l/2$), maximum value, $I_{\text{max}}$, at the center, $x = 0$, and zero value again at the right end ($x = +l/2$, $l$ being the total length of the dipole). It can be shown that the short dipole with linear current distribution is like an elemental dipole with an “effective length” equal to half its actual length, and a uniform current equal to the current at the center of the actual dipole. Equations (4–2) to (4–6) may be used to describe the behavior of the actual short dipole if $l$ in these equations is replaced by $(l/2)$. With this modification, the entire discussion of the
elemental dipole, including the feedpoint impedance considerations, applies to the actual short dipole. The directivity and the beamwidth—in fact, the total pattern—are the same for the open-ended short dipole and for the elemental dipole.

4.1.8. Short Vertical Antenna with Ground Image

At very low frequencies, below 500 kHz, for example, the earth in most localities is a nearly perfect reflector of radio waves. Since a wavelength at these frequencies is physically quite long (600 m or about 2,000 ft. at 500 kHz, for example, and 30 km or about 19 miles at 10 kHz!) it is difficult to get a horizontal antenna high enough above the earth to produce appreciable low-angle radiation (based on the principles illustrated in Figs. 1–14 and 1–15). Moreover, antenna lengths that are an appreciable fraction of a wavelength are quite long, and expensive.

If a vertical antenna is erected with its base at the ground, it will be imaged in the earth in accordance with the principle of images (see sec. 1.2). The phase of the equivalent current in the image conductor is such that the antenna-plus-image may be considered a single antenna in free space. Since the height of the vertical antenna-plus-image, 2h, will usually be a small fraction of a wavelength, the radiation is like that of a short dipole in free space. This combination of a half dipole, of height h, and its image in a reflecting surface is known as a monopole.

The pattern, however, is actually only half the free-space pattern, since the earth “cuts off” the other half. For a given current at the base of the antenna, the total radiated power is only half as great as it would be for the antenna-plus-image in (actual) free space with the same maximum current, as is found by substituting π/2 for π in the θ-integration of (4–5). Therefore, the radiated power and radiation resistance are only half as great as the values calculated on the free-space antenna-plus-image basis. Then, for a fixed current I, because of a smaller radiation resistance, both the input power and the radiated power are reduced.

As already discussed in connection with Fig. 4–7, the effective length of a short dipole being one-half its overall length may be used. Thus, although the length of the antenna-plus-image is 2h, the effective monopole length is half as much, that is, h.

The total radiated power $P_{\text{total}}$ of a short monopole is determined as follows:

(1) replace the length l of the elemental dipole of (4–5) with h, the effective length of the short monopole (i.e., the half-dipole plus image in ground plane), and
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(2) substitute $\pi/2$ for $\pi$ in the $\theta$-integration of (4–5), because the earth’s blockage; this causes the factor 790 to be replaced by 395.

Then, by using the $P_{\text{total}}$ for the monopole, its radiation resistance becomes

$$R_r = \frac{P_{\text{total}}}{I^2} = \frac{395h^2}{\lambda^2} \quad (4–8)$$

It is to be recalled that $h$ is the actual height of the half-dipole of the monopole, which is the same as the effective height of the half-dipole plus its image. Alternatively, $R_r$ may be expressed in terms of the effective height (i.e., length) of the half-dipole $h_e$, which equals $h/2$. Then, $R_r = 1580h_e^2/\lambda^2$.

Recall that, for a monopole, the total power radiated is effectively concentrated into half the solid angle of the free-space case. Then, by the reasoning given in chapter 3, e.g., equation (3–21), the directivity will be twice that of a free-space short dipole. Thus for the monopole, $D = 3$, as compared to $D = 1.5$ given in (4–7) for the elemental dipole. The pattern in the horizontal plane is uniform (circular), that is, equal signal strength is radiated in all horizontal directions. The radiation at the peak of the beam is vertically polarized. The half-power vertical beamwidth is half that of the free-space short dipole (i.e., 45 degrees) since the earth eliminates half of the pattern. However, calculations of field strength at a distance in the actual earth environment cannot be made on the basis of these “semi-free-space” results because the propagation of the vertically polarized waves depends on the semi-guiding effect of the earth’s surface, and at very great distances the ionosphere plays a part, as discussed in sec. 1.4.

A vertical radiator of this type may take the form of a steel tower with its base insulated from earth; then it is fed by connecting the source (transmitter) between the tower base and the ground, with an inductance in the feed line to compensate for the capacitive reactance of the antenna. The ground is a source of appreciable loss unless care is taken to minimize the resistance of the ground connection. In high-power transmitting installations an elaborate network of buried wires is used to make a good connection.

4.1.9. Top-Loaded Antenna

As has been indicated in the discussion of input impedance, in this section, the low radiation resistance of a short dipole together with the high capacitive reactance component of the input impedance create a difficult feed problem. As (4–8) above shows, the radiation resistance can be increased by increasing the effective height. This can be done either by increasing the actual height, or by changing the current distribution in the antenna so as to increase the effective height. As has been explained in the discussion of a short dipole with nonuniform current distribution, with a linearly decreasing current that has zero value at the end of the radiator the effective height is only half the actual height. If the current can somehow be made uniform, the effective height becomes equal to the actual height, and the radiation resistance is increased by a factor of 4 (because the effective height is squared in the radiation-resistance formula, i.e., the numerical constant 395 in (4–8) becomes 1580 if $h$ is interpreted as the effective height).
Current uniformity can be either totally or partially accomplished by “top loading”* the antenna. This procedure consists of running wires approximately horizontally from the upper end of the dipole. These wires do not radiate, but their capacitance to ground results in larger current at the upper end of the vertical conductor. If this “flat top” of horizontal wires is sufficiently extensive, and the vertical section is of “short dipole” length, the vertical current distribution may be made practically uniform (as assumed in analysis of elemental dipoles) so that the effective height is equal to the actual height. When the top loading does not produce approximately uniform current in the vertical section, the result is an effective height less than the actual height but greater than the effective height with no top loading.

The top loading also has the effect of reducing the magnitude of the capacitive-reactance component of the input impedance. This means that much less inductance is needed in series with the feed line, and also that for a given radiated power the voltage across this inductance and at the antenna base will be greatly reduced. This effect is perhaps the greatest benefit of top loading in high-power installations, since voltage breakdown and corona losses on the antenna are serious problems.

Top-loaded short vertical radiators at very low frequencies take many physical forms. A typical one is illustrated in Fig. 4–8. The flat top, as shown, is strung between two vertical towers, and the vertical radiating portion is supported at the center of the flat-top span. The flat top is insulated from the towers, and the antenna is fed at the bottom of the down lead through an inductance; the other side of the feed line is grounded through a network of buried conductors.

A dramatic form of this type of antenna is the United States Navy’s Jim Creek VLF installation (in the state of Washington), which operates in the 10 to 30 kHz frequency region. This flat top is a conductor supported between two mountains about 2 kilometers

* The equivalent procedure for a dipole would be called *end loading*. The term *top loading* is used simply because the end of the monopole is at the top. On a dipole, both ends would be “loaded,” symmetrically.
apart, and the vertical radiating portion is close to 300 meters in height. Another United States Navy installation, in Cutler, Maine, has towers nearly 300 meters high; but even these heights are short compared to the wavelength at these frequencies ($\lambda = 10,000$ meters at $f = 30$ kHz).

To illustrate the loss problem with electrically short vertical antennas using ground images, typical radiation efficiency factors ($k_r$, equation 3–30) range from 0.05 to 0.5 (often expressed as 5 to 50 percent). The majority of the loss occurs in the ground resistance, though some occurs in the feed-line tuning coil and, in very high-power installations, in insulator leakage losses and corona. (Efficiencies of high-frequency antennas are often close to 100 percent.)

### 4.2. Current and Voltage in Longer Antennas

Electrically short antennas have been discussed in some detail because they are of both theoretical and practical importance. At higher frequencies, beginning in the MHz region and especially in the HF region and above, longer electrical lengths become feasible, and their use results in certain advantages. In particular, the difficulties encountered in feeding short dipoles are eliminated, and higher directivities may be obtained. The input impedance of longer antennas may be made nonreactive, and the radiation resistance is usually much higher than that of a short dipole.

The variation of current and voltage along the length of longer antenna wires is more complicated than for short or elemental dipoles. For an elemental dipole it was assumed that the current is uniform (constant) along the conductor at any instant of time, although varying in time according to (4–1). For a short dipole without capacitive end loading, the current varies in a linear (straightline) fashion, as shown in Fig. 4–7. The voltage, though not shown in Fig. 4–7, is of opposite polarity on either side of the feed-point gap and has virtually constant amplitude along the conductor length.

The patterns of variation of the current and voltage along the conductor are called the current and voltage distributions of the antenna. These distributions are important in understanding the radiation properties of various antenna lengths and feed arrangements.

The current and voltage distributions on open-ended long antenna wires are basically similar to the standing waves of current and voltage on an open-ended two-wire transmission line. This standing-wave pattern is shown in Fig. 4–9a. (The voltage distribution is the same as shown in Fig. 2–5, chapter 2, for $Z_L \to \infty$.) The voltage maxima occur at the end of the line, and at other points an integral number of half wavelengths from the end. The voltage has zero values (nulls) at points an odd number of quarter wavelengths from the end. The current has maxima at the voltage minima, and the current minima are at the voltage maxima.

These patterns are plotted with both positive and negative “amplitudes” ($I_0$ and $V_0$) to emphasize the phase reversals. The same phase information is conveyed by the phase distribution patterns of Fig. 4–9b, which show that the phases of the voltage and current are constant in the intervals between the nulls and that a sudden change of 180 degrees occurs at the nulls. It is also shown that the current and voltage are everywhere 90 degrees
out of phase with each other. (A 270-degree phase difference is equivalent to 90 degrees.)

If now the wires in the end quarter-wavelength section of the open-ended two-wire line are bent outwards at right angles to their original directions in the plane of the line, as shown in Fig. 4–10a, the result is a half-wave center-fed dipole. Figure 4–10b shows the current and voltage distribution on the dipole. Essentially it is the same pattern that existed on these wires before they were bent to form the dipole, with zero current at the ends and maximum current at the center. The voltage is maximum at the ends and zero at the center. These features are the same as those of the short open-ended dipole, whose current distribution is shown in Fig. 4–7. The difference is that the variation is no longer linear; it is sinusoidal. If the current and voltage are expressed as functions of the distance \(x\) measured along the wire from one of the open ends of the antenna, the phasor* amplitudes are given by the equations:

\[
I_0(x) = jI_{0(m)} \sin \left(\frac{2\pi x}{\lambda}\right) \tag{4–9}
\]

\[
V_0(x) = V_{0(m)} \cos \left(\frac{2\pi x}{\lambda}\right) \tag{4–10}
\]

where \(I_{0(m)}\) and \(V_{0(m)}\) are the maximum amplitudes. These equations do not show the time variation, or instantaneous values; they are obtained by multiplying the amplitudes by the factor \(\sin (2\pi ft + \alpha)\), as in (4–1).

The space–time relationships of current and voltage on a half-wave dipole are somewhat difficult to visualize at first. As an aid in this effort, Fig. 4–11 shows the instantaneous patterns of current and voltage on the dipole at several instants during a single rf cycle. In these diagrams, the rf period \(T\) is the time for completion of a single cycle;

* In these expressions, the \(j\) factor indicates that the current is 90 degrees out of phase with the voltage, and the positive and negative sign changes resulting from the sine and cosine functions of \(x\) indicate the 180-degree phase changes that occur at the standing-wave nulls. \(I_{0(m)}\) is the value of \(I_0(x)\) that occurs at \(x = \lambda/4\), and \(V_{0(m)}\) is the value of \(V_0(x)\) that occurs at \(x = 0\) and \(x = \lambda/2\).
it is equal to $1/f$, where $f$ is the frequency in Hertz. The patterns are drawn assuming that at zero time ($t = 0$) the current is at its instantaneous maximum value, corresponding to $\alpha = \pi/2$ in (4–1).

Linear antennas longer than a half wavelength are also used. They may or may not be fed at their centers. On each side of the feed point, the current and voltage distributions are determined by (4–9) and (4–10) above. For correct operation of such an antenna,
with a balanced two-wire line, the lengths of wire on either side of the feed point must be the same or must differ by an integral number of half wavelengths. Linear antennas of this type, on which standing waves of current and voltage exist owing to reflections from the end of an open-ended wire, are called resonant antennas.

At the gap in the antenna wire across which the feed line is connected, the antenna voltage distribution undergoes a 180-degree phase reversal, but the current phase is the same on either side of the gap. Therefore, if the feed point is at a current maximum (voltage minimum), the distributions over the entire antenna length will be the same as they would be in an unbroken wire of the same length (case 1). This situation exists when the wire lengths on each side of the feed point are odd integral multiples of a quarter wavelength. But if the feed point is at a voltage maximum, which will be the case if the wire lengths on each side are integral multiples of a half wavelength, the antenna phase pattern is not the same as it would be on an unbroken wire (case 2). The voltage-current-phase patterns for these two cases are shown in Fig. 4–12, for a total antenna length of one wavelength. Antennas of the first type, having the unbroken-wire type of distribution (obtained by feeding either at one end or at a current maximum), are properly termed long-wire resonant antennas; those of the second type, in which the feed point is a voltage maximum, are actually two-element collinear arrays. (If the total antenna length is one wavelength, as in Fig. 4–12b, each element of this collinear array is a half-wave dipole, and for longer total lengths each element may itself be a long-wire antenna.) Because arrays are discussed in chapter 5, only the true long-wire types will be discussed in this chapter, although some types of collinear and other long-wire arrays are sometimes loosely referred to as long-wire antennas.

Actually these sinusoidal distributions of current and voltage are approximations rather than exact descriptions. They are slightly modified, on an actual antenna, by the
radiation resistance of the antenna and by the fact that the antenna wires are not equivalent to a uniform transmission line. The radiation resistance, as well as any actual resistance in the antenna wire, results in a small component of current that is in phase with the voltage, rather than 90 degrees out of phase. But the sinusoidal approximation is quite good for linear antennas whose conductors are very thin compared to their length, and of high conductivity. (It is also assumed that the antenna wire is not close to any large irregular conducting bodies or dielectric material that would disturb the uniformity of the electrical environment. In fact, a free-space environment is assumed, but the assumed distributions apply reasonably well in practical situations.)

Antennas may also be designed to have uniform current and voltage amplitudes along their lengths, that is, no standing waves. This result is achieved by terminating the end of the antenna wire in a resistive load so that no reflection occurs. In one form of such an antenna (Beverage or wave antenna), the wire runs approximately horizontally above the earth, and the input terminals consist of one end of the wire and the ground. The terminating resistor is connected between the other end of the wire and the ground. In another form (rhombic antenna) long wires form an array in the shape of a diamond (rhombus) in a horizontal plane. The two sides of the diamond are fed at one vertex, and the terminating resistor is connected between them at the other vertex. The current and voltage are approximately constant along the wires, but there is a gradual decrease of both with increasing distance from the feed point, owing to the radiation losses and the ohmic loss in the wire. The current and voltage are in phase with each other everywhere, rather than approximately 90 degrees out of phase as with standing-wave distributions, but their phases change linearly with distance along the wire in the amount of \(2\pi\) radians or 360 degrees for every wavelength. This description is characteristic of traveling waves, as described by (1–2) and (1–3) of chapter 1 for waves in space, and by (2–3) of chapter 2 for waves on wires. Antennas having traveling-wave current and voltage distribution are called nonresonant antennas or traveling wave antennas.

### 4.3. The Half-Wave Dipole

The radiation patterns of linear antennas that do not qualify as “short dipoles” may be found by considering them to be composed of a number of elemental dipoles placed end to end. For example, a dipole a half wavelength long might be approximated by five tenth-wavelength elemental dipoles end to end. The current in each elemental dipole would (by definition) be constant and equal to the average current in the corresponding section of the half-wave dipole, as indicated in Fig. 4–13. The current distribution is a half cycle of a sinusoid with the maximum at the dipole center, as in Fig. 4–10. The current has a constant phase angle everywhere on a half-wave dipole so that all the elemental dipoles are assumed to be in phase.

The radiated field intensity at a distant point (field point) due to each elemental dipole was given previously by (4–2), and its phase angle was given by (4–3), with the distance \(r\) taken to be the distance to the field point from the center of the elemental dipole; that is, \(r\) will have a slightly different value in computing the field-point contribution of each elemental dipole. These slight distance differences will not significantly affect the relative
intensities of the individual elemental dipole fields, but they will affect the relative phases significantly. The total field at a distant point is the phasor sum of the contributions of the individual elemental dipoles, in accordance with the principle of interference in sec. 1.2.

This method of analysis, as described thus far, is obviously an approximation and a rather crude one, when the half-wave dipole is dissected into only five elemental dipoles. The accuracy of the approximation may be increased by dissecting it into more elemental dipoles of shorter individual lengths. But the labor of calculation is also thereby increased, if the phasor summation process is employed as described.

The approximation may be made exact, however, and the tedious summation process avoided, by applying the methods of calculus. The half-wave dipole is then considered to be composed of an infinite number of infinitesimal dipoles, and the phasor summation of their fields at a distant point is expressed as an integral. The resulting expression for the magnitude of the electric field of the half-wave dipole, at a distance \( r \) in the direction \( \theta, \phi \), obtained by solving the integral expression, is

\[
E (r, \theta, \phi) = \frac{60 I}{r} \left[ \cos\left(\frac{\pi}{2} \cos \theta\right) \right] \sin \theta \tag{4-11}
\]

where \( E \) is in rms volts per meter if \( r \) is in meters and \( I \) is the rms current in amperes at the center of the dipole. This pattern is seen to be a slightly more complicated mathematical expression than that of the short or elemental dipole of (4-2), but the patterns are only slightly different. They are compared in Fig. 4-14. The half-wave dipole has a slightly narrower beamwidth—78 degrees compared to 90 degrees for the short dipole. Consequently its directivity is slightly greater—1.64 compared to 1.5 for the short dipole. The power-density ratio is \( 1.64/1.5 = 1.093 \), and the field-strength ratio 1.047.

The slightly greater directivity of the half-wave dipole is thus almost insignificant. Its advantage lies primarily in its increased radiation resistance and reduced or nonexistent feed-point reactance. The radiation resistance for an exactly half-wavelength dipole is found, by the method illustrated in the case of the elemental dipole, to be 73.1 ohms, referred to the maximum current point (dipole center). Therefore this is also the resistive component of the input impedance when the dipole is fed at the center. There is also a small reactive component of 42.5 ohms, inductive. This small inductive reactance may
be eliminated by shortening the dipole to about 95 percent of a half-wavelength (i.e., about 0.475\(\lambda\)). The radiation resistance (and input impedance) is then about 65 ohms (Kraus and Marhefka 2002, p. 182). The pattern (beamwidth and gain) is not significantly affected by this slight shortening.

These properties make the half-wave dipole especially attractive as a radiator for many purposes, at frequencies for which its length is not excessively large or minutely small. The results given for the radiation resistance and input impedance are free-space values for a conductor very thin compared to the length, with no ohmic resistance. Consequently they become somewhat modified for conductors of appreciable diameter or when the dipole is close enough to the ground or other conductors to result in “coupling.” Usually ohmic loss is small enough to be disregarded. Dipoles of larger diameter, and of special shapes and configurations, are discussed in sec. 7.1.

When a vertical quarter-wavelength radiator is erected with its base at or just above the ground, it is imaged in the earth so that its radiation may be analyzed as if it were a half-wave dipole in free space, subject to the same modifications as discussed for the short vertical dipole imaged in the ground. When the quarter-wave vertical antenna (monopole) is fed at its base with the other side of the feed line connected to ground, its radiation resistance and input impedance are just half the values for the half-wave dipole in free space, and the directivity is twice as great. The radiation is vertically polarized at the peak of the beam. Vertical radiators of other lengths may be similarly analyzed, that is, by use of the image principle. Vertical-tower radiators of heights up to about 5/8 wavelength are much used for broadcasting and other applications in the medium-frequency range of about 500 to 3,000 kHz. (Even higher vertical antennas may be used if they are “sectionalized” so that they become, in effect, collinear-array antennas, described in sec. 5.2). A monopole too short to be a quarter-wavelength high, yet too long to be classed as a “short monopole,” may be capacitively top loaded (in the same general manner as described for short vertical monopoles) to result in virtually quarter-wave performance.

### 4.4. Long-Wire Antennas

Antennas consisting of a single straight wire, either unbroken or with a feed-point gap at a current maximum when the antenna has standing-wave current distribution, are classed as long-wire antennas if their length is substantially greater than a half wave-
length. Such antennas are not properly called dipoles. On the other hand, the half-wave antenna is commonly called a half-wave dipole, even though a true electric dipole is equivalent to two equal and opposite polarity point charges separated by a definite distance. The elemental dipole and the short dipole are, in essence, equivalent to oscillating electric dipoles, but longer antennas are not. However, usage sanctions the term for the half-wave dipole.

The radiation patterns of long-wire antennas may be determined by the method described for the half-wave dipole by considering them to be composed of end-to-end infinitesimal dipoles. The current amplitude and phase in each infinitesimal dipole are taken to be the values indicated by the current distributions calculated from (4–9), and as shown for a one-wavelength wire in Fig. 4–12 for a resonant antenna. For a nonresonant antenna a constant current amplitude along the wire is assumed, with a linear phase change corresponding to a traveling wave of current ($2\pi$ radians or 360 degrees per wavelength). These current-distribution assumptions are valid for a thin wire of perfect conductivity, ignoring the effect on current distribution of the radiation losses. Therefore, the results are approximate, but useful in that they indicate the general nature of the radiation patterns.

4.4.1. Patterns of Resonant Antennas

As shown in Fig. 4–14, the pattern of a short or half-wave dipole consists of a single doughnut-shaped lobe of radiation (appearing as two oppositely directed lobes in a “slice” or plane pattern containing the dipole axis). Long-wire radiators have more than one three-dimensional lobe, taking the form of cones of radiation. The axes of the cones coincide with the axis of the wire, and the sides of the cones are inclined at various angles with respect to the wire. As for the short dipole, the patterns are uniform (circles) in the plane perpendicular to the axis of the wire.

There will be one cone-shaped lobe for each half wavelength of wire length, for both the standing-wave and traveling-wave antennas. The lobes are symmetrically disposed with respect to the plane that bisects the wire. Therefore, if there is an odd number of half wave-lengths, one lobe will be perpendicular to the wire, like the short-dipole lobe except that it is thinner (narrower beamwidth), more like a pancake than a doughnut when the wire is many wavelengths long. When the wire length is an even number of half-wavelengths, there is no perpendicular lobe.

For the standing-wave or resonant antenna, the amplitude of the electric field strength (pattern) is given by

$$ E(r, \theta, \phi) = \frac{60I}{r} \left[ \cos \left( \frac{n\pi \cos \theta}{2} \right) \right] $$  

(4–12a)

where $n$ is the number of half wavelengths in the wire length, assumed to be an odd number, and as usual $E$ is in rms volts per meter if $r$ is in meters, and $I$ is the rms current.
at a current maximum, in amperes. For a wire an even number of half-wavelengths long, the equation becomes

\[
E(r, \theta, \phi) = \frac{60I}{r} \left[ \sin \left( \frac{n\pi}{2} \cos \theta \right) \right] \frac{\sin \theta}{\sin \theta} \tag{4–12b}
\]

The nature of these patterns is shown in Fig. 4–15 for an odd and an even number of half-wavelengths.

In these formulas, it is assumed as usual that the antenna is located at the origin of a spherical coordinate system with the axis of the wire along the \( \theta = 0^\circ \) axis (z-axis, Fig. 3–1) and that \( I \) is the rms current at a current-maximum point of the sinusoidal standing wave. The patterns shown are for long wires of modest length \((n = 3 \text{ and } n = 4)\). As the number of half wavelengths is made larger, the number of lobes increases proportionately and the lobes of maximum radiation lie closer to the wire. Because of the factor \( \sin \theta \) in the denominator of (4–11) and (4–12), an envelope of the lobe pattern is, in three dimensions, a circular cylinder parallel to the axis of the wire. In a plane containing the wire axis, the edges of the envelope are straight lines parallel to the wire. The effect is shown in Fig. 4–16 for a many-lobed pattern.

These patterns, and equations (4–11) and (4–12), are for antennas an integral number \((n)\) of half wavelengths long. The total length, however, should be shortened by about 5
percent of one half wavelength to eliminate a reactive component of input impedance. This will not affect the pattern appreciably.

### 4.4.2. Radiation Resistance and Directivity

The radiation resistance of a long-wire resonant antenna \( n \) half wavelengths long, in free space, is given approximately by

\[
R_r = 73 + 69 \log_{10} n \tag{4-13}
\]

for values of \( n \) greater than 2 (Brainerd et al. 1942). The equation is also approximately correct for \( n = 1 \).

The angle of maximum radiation, that is, the angle that the strongest lobe makes with the wire axis (this is also the lobe closest to the axis, as Figs. 4–15 and 4–16 show), is given approximately by (4–14) below. This formula is quite accurate for small values of \( n \) and gives a result close enough for most purposes, even for large values of \( n \).

\[
\cos \theta_{\text{max}} = \frac{n - 1}{n} \tag{4-14}
\]

The maximum directivity may be attained through knowledge of the maximum field strength \( E_{\text{max}} \), the radiated (transmitted) power \( P_t \), and other previously developed relationships. By substituting \( \cos \theta_{\text{max}} \) into (4–12a) or (4–12b), according to whether \( n \) is odd or even, the value \( 60I/\text{rsin} \theta_{\text{max}} \) for the maximum field strength \( E_{\text{max}} \) is obtained.* The power \( P_t \) is \( I^2R_r \) from equation (3–28). Furthermore, the maximum and average power densities \( p_{\text{max}} \) and \( p_{\text{av}} \) are \( (E_{\text{max}})^2/377 \) and \( P_t/4\pi^2 \), respectively. Then, the maximum directivity \( D_{\text{max}} \) becomes

\[
D_{\text{max}} = \frac{U_{\text{max}}}{U_{\text{av}}} = \frac{p_{\text{max}}}{p_{\text{av}}} = \frac{120}{R_r \sin^2 \theta_{\text{max}}} \tag{4-15}
\]

It is interesting to note that this formula gives the correct result for a half-wave dipole, \( D_{\text{max}} = 1.64 \), when \( R_r \) is taken as 73 ohms and \( \theta_{\text{max}} = 90^\circ \); also, (4–13) and (4–14) are correct for \( n = 1 \).

* When (4–14) is substituted into (4–12a) and (4–12b), their numerators become, respectively, \( 60|\cos \{(n - 1)\pi/2\}| \) and \( 60|\sin\{(n - 1)\pi/2\}| \). For \( n \) odd, \( |\cos\{(n - 1)\pi/2\}| = 1 \) and for \( n \) even, \( |\sin\{(n - 1)\pi/2\}| = 1 \). The denominators become \( \text{rsin} \theta_{\text{max}} \).
4.4.3. Patterns of Nonresonant Antennas

A long wire with a traveling-wave current of uniform amplitude \( I \) has an electric-intensity pattern that is given by

\[
E(r, \theta, \phi) = \frac{60I \sin \theta}{r(1 - \cos \theta)} \sin \left[ \frac{\pi L}{\lambda} (1 - \cos \theta) \right]
\]

(4–16)

where \( L \) is the length of the wire. This pattern has the same number of lobes as a resonant wire of the same length, and the maxima and minima occur at approximately the same positions. Their magnitudes, however, are quite different, as shown by the pattern of a 3/2-wavelength nonresonant wire in Fig. 4–17. As seen there, the lobes directed toward one end of the wire are much larger than those at the other end of the pattern. The lobe nearest the axis of the wire and pointed in the direction of the traveling wave of current on the wire is the largest. The smallest lobe is the one at the other end of the pattern, the magnitudes increasing progressively toward the large-lobe end.

This type of pattern has an advantage when it is desired to radiate or receive in predominantly one direction, rather than two. Suppression of the pattern in one direction is accomplished by eliminating the reflected current at the end of the wire by means of a resistive termination. This usually takes the form of a resistor connected from the end of the wire to ground. Such termination can be successful, however, only if the height of the antenna above ground is a very small fraction of a wavelength (otherwise the connection would have reactance as well as resistance and would not be a reflection-free termination). The correct value of the resistor, being connected between the end of the wire and ground, is half the value that matches the impedance of a transmission line consisting of the antenna wire and its earth image. If the antenna height is \( h \) and the wire diameter is \( d \), the resistance is from equation (2–62), chapter 2:

\[
R = 138 \log_{10} \left( \frac{4h}{d} \right) \text{ ohms}
\]

(4–17)

However, this formula should be used only as a rough guide. The usual practice is to adjust the resistance until no standing wave exists on the antenna wire.

4.4.4. Polarization

The radiation from a long-wire antenna is linearly polarized, but the polarization (electric field) direction is not the same in all parts of the pattern. (This is true even of short and half-wave dipoles, but for them the variation is not as great because there is only one lobe perpendicular to the wire; in the perpendicular plane the polarization is simply parallel to the wire.) The polarization in one of the oblique
lobes of a long-wire antenna, or in fact in any particular part of the pattern, may be
determined by the following procedure (assuming free-space propagation):

(i) Draw a line from the center of the antenna in the direction of interest.
(ii) Form the plane that contains both the antenna wire and this direction line.
(iii) At any point on this direction line the polarization (electric field) vector is
    perpendicular to the direction line and lies in the plane thus formed.

(This procedure is in fact applicable to the radiation from any straight-wire radiator,
including short dipoles and half-wave dipoles.)

4.4.5. Effect of the Ground and Other Factors

The foregoing discussion of long-wire antennas has assumed perfect thin-wire conductors
and a free-space environment. Ground reflection affects the vertical-plane pattern in the
manner discussed in sec. 1.4.3 and illustrated for a special case by Fig. 1–15. The presence
of the ground also affects the radiation resistance and the input impedance because
of mutual coupling between the antenna and its image. Resistance of the wire itself is
usually small, but in a very long wire the total resistance may be appreciable. In addition
to its direct effect on the input impedance, this resistance also changes the form of the
current distribution on the wire, in both the resonant and nonresonant antennas, and so
affects the radiation pattern. Therefore the behavior calculated for the free-space perfect-
conductor antenna serves primarily as a general guide to what will be observed with
practical antennas. When correction for ground effects is made, the theoretical patterns
agree quite well with those that actually occur.

4.4.6. Uses of Long-Wire Antennas

Both resonant and nonresonant long-wire antennas are used for transmitting and receiv-
ing in the MF and HF range, from perhaps 500 kHz to 30 MHz. They provide a simple
and effective method of obtaining a directional pattern and gain. As will be described in
Ch. 5, these properties can be further enhanced when long-wire antennas are used as
elements in an array.

The single terminated wire used as a nonresonant antenna will not be effective for
horizontal polarization, as already discussed, because of its small-fraction-of-a-wave-
length height. As the discussion of polarization indicates, however, a lobe that makes a
small angle with the axis of a horizontal wire in a vertical plane will radiate or receive
waves whose electric field vector has an appreciable vertical component. Such an antenna
is sometimes used as a rather highly unidirectional receiving antenna for vertically polar-
ized waves. In this use, the long-wire nonresonant antenna is known as a Beverage or
"wave" antenna (Beverage, Rice, and Kellog 1923).

The functioning of a Beverage antenna is now described. For a wave with propagation
direction slanted between horizontal and vertical, there will be both a vertically- and a
horizontally-polarized (V- and H-POL) component. Since the earth is an imperfect con-
ductor, the H-POL component is not totally diminished. Thus, the incoming H-polarized
fields, from the ground, induce voltages along the antenna that add in phase at the receiving end. Waves propagating along the wire from the opposite direction are, ideally, absorbed by the terminating resistor. Therefore, the Beverage antenna provides a highly directive pattern in the horizontal plane for vertical polarization. Beverage antennas are not ordinarily used for transmitting, because the power absorbed in the terminating resistor results in poor radiation efficiency.

4.5. Loop Antennas

Another basic form of radiator is the loop, which in its fundamental form is a single-turn coil of wire. A current can be made to flow in the loop by breaking it at some point and connecting the terminals of a transmission line (or other source) at the gap in the loop, as indicated in Fig. 4–18.

4.5.1. The Small Loop

The radius of the small loop, \( a \), is assumed to be very small compared to the wavelength \( \lambda \), so that the current in all parts of the loop will be of the same amplitude and phase at any instant. An analysis of the radiation from such a loop may be made by considering it to be made up of many elemental dipoles connected together. Since dipoles are straight rather than curved, the figure thus formed will be a polygon rather than a circle. However, the approximation to a circle can be made as good as desired by taking the elemental dipoles to be sufficiently short or, ideally, infinitesimal. The fields of the individual dipoles are then superposed in the manner described for analyzing the half-wave dipole, sec. 4.3. Here not all the electric vectors of the separate field components are parallel. Although this complicates the mathematics, the principle is the same. The superposition of nonparallel fields was discussed in sec. 1.2.

From such an analysis it is found that the field pattern of a loop has exactly the same shape as that of a single elemental dipole oriented with its axis coincident with the loop axis (i.e., with its axis perpendicular to the plane of the loop) (Fig. 4–4). However, the vector directions of the electric (\( E \)) and magnetic (\( H \)) components of this field are interchanged, relative to the \( E \) and \( H \) directions of the elemental-dipole field. The polarization is linear but perpendicular to that of the corresponding electric dipole. Therefore a loop with its axis horizontal radiates maximum field intensity in the plane of the loop, the pattern being doughnut-shaped; but with axis horizontal the polarization is vertical, rather than horizontal. If the plane of the loop is horizontal, the radiation pattern in the horizontal plane is uniform (a circle), like that of a vertical dipole, but the polarization is

![Basic form of loop antenna.](image-url)
horizontal. The small loop is an oscillating magnetic dipole, which is equivalent to a bar magnet whose magnetic strength and polarity oscillate. Because the pattern has the same shape as the elemental electric dipole, the directivity is the same \( D = 1.5 \) and so is the half-power beamwidth \( BW_{3dB} = 90^\circ \).

Now consider Fig. 4–18 and assume an observer is on the \( z \)-axis, which extends outward from the page, and where angle \( \theta = 0^\circ \). Recall that the current \( I \) is assumed constant and of the same phase everywhere on the loop. Then, the radiation reaching the observer from any two diametrically opposite located elemental dipoles are equal in magnitude and opposite in phase, and thus cancel. In other words, at \( \theta = 0^\circ \) the pattern has zero amplitude, as indicated in Fig. 4–4 for the dipole. Now notice that as \( \theta \) is increased by being removed from the \( z \)-axis, the path length difference (and thus the phase difference) increases between the radiation from diametrically opposite elemental dipoles. Therefore, the pattern’s amplitude increases with increases in \( \theta \). Furthermore, because of the symmetry of the loop about the \( z \)-axis, it is clear that the amplitude is constant and independent of \( \theta \). Therefore, analogous to the pattern of a short dipole, the \( E \)-field pattern of a loop antenna is donut shaped.

The polarization can also be determined from considering Fig. 4–18. Let an observer be anywhere on the \( x-y \) plane and far removed from the loop. Then, recalling that the polarization from each elemental dipole within the loop is aligned with the direction of its current, one can discern that the vectorial sum of the radiation at each angle \( \theta \) is perpendicular to the \( z \)-axis. But, because of loop symmetry, the electric field lines are necessarily concentric about the \( z \)-axis. Thus, if the \( z \)-axis is vertical, the polarization is horizontal.

The relationship between the loop current and the radiated power is quite different for a loop, since the radiation resistance of the elemental dipole depends on the ratio of its length to the wavelength, and the geometry of a loop is not comparable. The formula for the radiation resistance of a small loop is

\[
R_r \approx \frac{31,200A^2}{\lambda^4} \text{ohms} \quad (4–18a)
\]

where \( A \) is the area of the loop and \( \lambda \) is the wavelength. Since \( A \) can be expressed in terms of circumference \( C \) as \( C^2/4\pi \), (4–18a) can also be expressed as

\[
R_r \approx 197\left(\frac{C}{\lambda}\right)^4 \quad (4–18b)
\]

In these formulas, \( A \), \( C \), and \( \lambda \) must be expressed in the same units of length.

The elemental-dipole pattern (with \( E \) and \( H \) directions interchanged) and these radiation-resistance formulas apply only to loops that are small compared to a wavelength. This criterion is considered to be met if the loop diameter is less than 0.1 wavelength. For this maximum size of loop, the radiation resistance is about 2 ohms. A very large current, therefore, is needed for radiation of appreciable power, as was also found to be true of short dipoles (sec. 4.1).
It turns out that the results found for a circular loop apply equally well for a loop of any shape as long as its dimensions are sufficiently small compared to a wavelength; the radiation resistance depends only upon its area, following (4–18a). The loop may be square, triangular, or even irregular in shape.

At very low frequencies it is common to make a loop with more than one turn of wire. If the number of turns is $N$, the resulting radiation resistance is found by multiplying the one-turn value of (4–18a) or (4–18b) by $N^2$. It is necessary that the total length of wire be small compared to the wavelength if the small-loop behavior is to apply, but if the frequency is very low (wavelength very long) this requirement is not difficult to meet. Because of the large wavelength, however, the radiation resistance may still be a very small value in spite of the number of turns, and the loop may have a significant ohmic resistance also. Therefore the radiation efficiency will be poor.

When such loops are used for receiving, the terminals of the loop may be connected to a very high-impedance receiver input circuit, so that the quantity of primary interest is the voltage induced in the loop rather than the power delivered. When the loop has its plane in the direction of a properly polarized incoming signal, if the incident-wave field intensity is $E$ volts per meter, the induced voltage will be

$$V = \frac{2\pi NAE}{\lambda} \text{ volts}$$ \hspace{1cm} (4–19)

where $N$ is the number of turns of the loop and $A$ is its area.

Small loops are often used for receiving as direction finders when the received signals are vertically polarized. The direction of the received signal is determined by orienting the loop with its axis toward the signal direction. The “null” (minimum value) of the pattern is in this direction and is very sharp (corresponding to the pattern in the direction of the axis of a dipole, Fig. 4–2). Thus, when the orientation of the loop is adjusted for minimum signal, the direction of the signal is accurately indicated. There is a twofold ambiguity in the direction, however, because a null exists in the pattern on both sides of the loop. This ambiguity may be resolved in various ways. A common way is to combine (in the receiver input circuit) the loop output with the output of a small vertical dipole, with a 90 degree phase difference (produced by the circuit arrangement). If the loop and dipole signal amplitudes are equal, the resulting combined pattern has only one null and is therefore unambiguous. However, loop direction finding is successful only when the loop can be located in an environment free from nearby large reflecting objects that may result in signals arriving at the loop position from more than one direction. This destroys the null effect, or at the least it destroys the sharpness of the null.

4.5.2. Other Loop Antennas

This section describes some other types of loops. Smith (2007) discusses these and other configurations, and includes an extensive annotated list of references.

Loops are fed by coaxial cables or two-wire lines. An interesting coax-fed loop configuration has some features common with the short vertical monopoles on automobiles.
In both cases, the outer conductor of the coaxial cable is attached to the base plate. Also in both cases, the center conductor passes through a hole in the conducting base plate, from which an image is formed. In the monopole arrangement, an extension of the center conductor serves as the half-dipole above the base plate. However, for the loop, the extension is bent into a half circle, whereby its image in the base plate forms the other half of the loop.

For light-weight receiver applications, a low-loss magnetic ceramic (ferrite) is commonly used as the core within a multiturn loop to improve efficiency. A reflecting backplane (parallel with the plane of the loop) can be used for providing a unidirectional pattern and increased directivity. Resonant loops (discussed below) are also used as the elements of phased arrays to increase directivity: an example being coaxially positioned loops to function as the reflector, driven element, and directors of a Yagi-Uda array (Fig. 5–11).

Small-loop analysis is applicable to loop wire-lengths of roughly $0.1\lambda$ or less, so that the current distribution around the loop is approximately uniform. Larger loops are also used, especially at the higher frequencies. Ordinarily the patterns of larger loops have multiple lobes, and the current distributions on these loops affect the patterns considerably.

The Alford loop, shown in Fig. 4–19, is an example of a larger loop. It is more efficient than, and has a pattern similar to, that of a small loop. It consists of a square one-turn loop with quarter-wavelength sides, and it is fed with opposite phases at opposite corners. The other two corners are capacitively connected. The out-of-phase feed is achieved by transposing one branch of the feed line as shown. The capacitors are commonly open-end sections of transmission line; with the reactance given in chapter 2 by (2–24). The radiation resistance of the Alford loop is about 80 ohms. Its radiation pattern is similar (though not identical) to that of a small loop, but it is much more efficient.

As the circumference $C$ of a loop increases and approaches a wavelength, the peak of the beam moves toward the loop axis. Near this resonant length, the far field pattern is nearly the same as two parallel dipoles separated by approximately the loop diameter. The input impedance varies significantly with $C$, having large peaks in both the resistance and reactance for circumferences near odd multiples of $\lambda/2$. Near the resonant loop length of $\lambda$, the input resistance is about 100 ohms and the reactance component is relatively small. Thus, the input impedance for a loop with a one $\lambda$ circumference can be readily matched to a transmission line.

The pattern of the resonant $C = \lambda$ loop can be made unidirectional and the directivity increased an axial direction by placing the loop over a planar reflector. With spacings between the reflector and

![FIGURE 4–19. The Alford loop antenna.](image_url)
loop in the range of 0.05\(\lambda\) to 0.2\(\lambda\), measurements have given directivities of approximately 10, and input impedances that can be readily matched (resistance \(R \leq 135\ \Omega\)). The reflector used was square reflector and had different side dimensions between 0.6\(\lambda\) and 1.9\(\lambda\).

As already noted, when the loop circumference is increased to value near one wavelength, the maximum of the far field pattern is along the loop axis. Resonant loops of this type are used as the elements of a Yagi-Uda array to form a unidirectional beam along the axes of the loops and of the array. Details for choosing the dimensions of the loops and their spacings for a Yagi-Uda array are given by Balanis (2005, p. 599).

### 4.6. Helical Antennas

Another basic form of radiator is the helix, which is a wire (conductor) wound in the shape of a screw thread and used as an antenna in conjunction with a flat metal plate called a ground plane. Helixes are mostly used at relatively high frequencies so that their dimensions are appreciable compared to the wavelength. Theory and practice in the art of helical antennas has been developed largely by John Kraus and his associates at the Ohio State University (Kraus and Marhefka 2002, pp. 222–42).

As shown in Fig. 4–20, the helix is fed at one end, usually being connected to the center conductor of a coaxial transmission line whose outer conductor is connected to the ground plane. The basic geometry of the helix is described in terms of its diameter \(D\) and its turn spacing \(S\). For an \(N\)-turn helix the total length of the antenna is equal to

\[
\text{length} = (N - 1) \cdot S + \frac{\pi D}{2}.
\]

---

**FIGURE 4–20.**

Helical antenna.
NS, and the circumference \( C = \pi D \). The length of the wire per turn of the helix is
\[ L = \sqrt{S^2 + C^2} = \sqrt{S^2 + \pi D^2}. \]
The pitch angle \( \alpha \), an important parameter of the helix, is the angle that a line tangent to the helix wire makes with the plane perpendicular to the axis of the helix; and it can be found from this relation:
\[ \sin \alpha = S/L \quad \text{or} \quad \tan \alpha = S/\pi D = S/C. \]
The properties of helical antennas are described in terms of these geometric parameters. Many different radiation characteristics may be obtained by varying their magnitudes in relation to the wavelength \( \lambda \).

The feed wire (Fig. 4–20), which connects the terminus of the co-axial center conductor to the beginning of the actual helix, lies in a plane through the helix axis and is inclined with respect to the ground plane at approximately the pitch angle of the helix. Variation of its geometry affects the input impedance of the antenna.

When the dimensions of the helix are very small compared with the wavelength, the maximum radiation is in the plane perpendicular to the helix axis, and the radiation pattern is a combination of the equivalent radiation from a short dipole positioned on the helix axis and a small loop also coaxial with the helix. This type of helix is known as a normal mode helical antenna (NMHA), and it is widely used in wireless handsets. To reduce cost of manufacture, NMHAs have been fabricated by printing a conducting path into a groove in a small, cylindrical plastic post or mast (see Fig. 36–2, Vardaxoglou and James 2007).

For a NMHA, the patterns of the two equivalent radiators are of course the same, but the linearly polarized components are at right angles, and the phase angles at a given point in space are 90 degrees apart. Therefore, as explained in sec. 1.1.3, the resultant field is elliptically polarized or circularly polarized, depending on the field-strength ratio of the two components. This ratio depends on the pitch angle \( \alpha \). When \( \alpha \) is very small, the loop type of radiation predominates; when it becomes very large, the helix becomes essentially a short dipole. In these two limiting cases the radiated polarization is linear, in one having loop polarization, and in the other, dipole polarization. For intermediate values of \( \alpha \) the polarization is elliptical, and at a particular value of \( \alpha \) it will be circular. Wheeler (1947) showed that this result is obtained when
\[ S = \frac{\pi D^2}{2\lambda}, \]
which corresponds to a value of \( \alpha \) given by
\[ \tan \alpha = \frac{\pi D}{2\lambda}. \]

The analysis of the helix leading to these conclusions may be made by considering it to be equivalent to a number of small loops having the same diameter as the turns of the helix, with their planes parallel and their axes in line with the helix axis and spaced the same as the helix turn spacing. Then these loops are considered to be connected by short dipoles parallel to the helix axis and of length equal to the helix turn spacing. The radiation field of the helix is equivalent to that obtained by superposition of the fields of these elemental radiators.

When the diameter and spacing (D and S) are appreciable fractions of a wavelength, an entirely different radiation pattern is obtained. The maximum intensity is radiated in the direction of the helix axis, in the form of a directional beam with minor lobes at
oblique angles. The radiation in the main lobe is circularly polarized. It is this feature of
the helix, in this mode of radiation (the axial mode), that probably accounts for most of
the practical applications of this type of antenna.

A typical helical antenna operating in this mode has a circumference \( C \) of approxi-
mately one wavelength and spacing \( S \) approximately a quarter wavelength. The antenna
will operate quite well over a range of frequency so that these dimensions are noncritical.
The ground plane (which may be either a solid sheet or a wire grid or mesh) should be
at least \( \frac{3}{4} \) of a wavelength in diameter. The pitch angle may range from about 12 to 18
degrees; approximately 14 degrees is optimum. The gain and beamwidth depend on the
helix length (equal to \( NS \), Fig. 4–20, where \( N \) is the number of turns). The feed-point
impedance is resistive and of the order of 150 ohms at the frequency for which \( C = \lambda \).
At higher and lower frequencies the resistive component varies, and a reactive component
appears. Detailed design information are available (Kraus and Marhefka 2002, ch. 8).

In terms of a three-dimensional spherical coordinate system (Fig. 3–1) with the \( \theta = 0^\circ \)
axis coincident with the helix axis, the beam (pattern) has axial symmetry, that is, it
is the same in any plane containing the axis (does not depend on the longitude angle \( \phi \)).
The 3-dB beamwidth, obtained imperically, is given approximately by the formula

\[
\theta_{3\text{db}} = \frac{52}{C} \sqrt{\frac{\lambda^3}{NS}} \text{ degrees}
\]

(4–21)

The formula assumes that the pitch angle is between 12 and 15 degrees, that \( N \) is equal
to or greater than 3, that \( NS \) (the helix length) is not greater than 10 wavelengths, and
that \( C \) is between 0.75\( \lambda \) and 1.33\( \lambda \).

The directivity, subject to the same assumptions, is given by

\[
D_{\text{max}} = \frac{12NSC^2}{\lambda^3}
\]

(4–22)

In some applications it is necessary to pay attention to whether the helix is wound
with a right-hand or a left-hand pitch (analogous to right-hand and left-hand screw
threads). This determines whether the wave will be right- or left-hand circularly polar-
ized. (A receiving antenna designed to receive right-hand circular polarization cannot
receive left-hand circular, and vice versa.)

Helical antennas have found considerable application in space telemetry applications
at the ground end of the telemetry link with ballistic missiles, satellites, and space probes,
at HF and VHF. The circular polarization is useful in this application because of the
polarization rotation of waves produced by the ionosphere (Faraday effect, sec. 1.4.6).

4.7. Horn Radiators

The radiators thus far discussed are based on the concept of fields set up by alternating
currents in wires (in the generalized sense of the term “wire,” which includes tubing,
pipes, and bars). The most basic of them is the elemental dipole, since in principle all other current-carrying conductors can be regarded (mathematically and conceptually) as an assemblage of elemental dipoles, and the radiation field is then deduced by applying the principle of linear superposition to the fields of the individual dipoles.

Another class of radiators is based on the existence over a surface of a specific electromagnetic field configuration. The intensity, phase, and polarization of the field over this surface, or aperture, are analogous to the current amplitude, phase, and direction in antennas represented by an assemblage of dipoles. The description of the variation of these field quantities over the aperture is called the aperture distribution. When this distribution is known, it is possible in principle to calculate the radiation pattern, just as it is possible for a wire or arrangement of wires in which the current distribution is known. As for radiation due to currents, the analysis of radiation due to the field distribution of an aperture is based on Maxwell’s equations.

An example of an aperture over which the field distribution is known is the cross section of a waveguide, in which a particular known mode is propagating, as described in sec. 2.5. It is a well-known experimental fact that if such a guide is “sawed off” in a plane perpendicular to the axis of the guide, leaving an open end, radiation will occur from this open mouth. The resulting field pattern can be calculated from the known configuration of the field for the particular waveguide mode. This is the simplest case of a waveguide horn radiator.

It should not be surprising that radiation occurs under these circumstances because the fields inside a waveguide are propagating in essentially the same way that fields propagate in free space; the only difference is that they are constrained from spreading spherically by the walls of the guide. When this propagating field reaches the guide mouth it continues to propagate in the same general direction except that, in accordance with Huygen’s principle, it also spreads laterally, and the wavefront eventually becomes spherical, although there is a “near field” region in the vicinity of the mouth of the guide in which the wavefront is more complicated. It can be thought of as a transition region in which the changeover from guided propagation to free-space propagation takes place. This changeover involves a change of phase velocity and a change in the characteristic wave impedance, from those of the guide to the free-space values, $c = 3 \times 10^8$ meters per second and $Z_c = 377$ ohms.

Because the waveguide impedance is ordinarily different from this free-space value, the radiating open end does not usually present a matched-impedance load to the guide, resulting in an undesirable standing wave. This can be eliminated by some form of transformer matching device, such as those described in sec. 2.5. A better method, however, is to flare the walls of the guide. If this is done properly, it results not only in a matched impedance but also in a more concentrated radiation pattern, that is, narrower beamwidth and higher directivity. This flared structure is what is ordinarily meant by the term horn radiator.

Various possible flaring arrangements, resulting in different types of horns, are shown in Fig. 4–21. As shown, a rectangular guide may be flared on the narrow walls, the wide walls, or both. A sectoral horn is flared in only one dimension. If the flare is in the direction of the electric vector, as when the broad walls are flared with the $TE_{10}$ mode in
rectangular guide, the result is an E-plane sectoral horn; when the narrow walls are flared, the radiator is an H-plane sectoral horn. Flaring both walls results in a pyramidal horn. A conical horn is formed by uniform flaring of the walls of a circular waveguide.

If the flare angle $\phi$ is too great, the wavefront at the mouth of the horn will be curved rather than plane. This means that the phase distribution over the aperture will be non-uniform, resulting in decreased directivity and increased beamwidth. On the other hand, too small a flare angle results in a small aperture area for a given length $L$ of the horn. The directivity is proportional to the aperture size for a given aperture distribution. Thus, there is an optimum flare angle that provides the maximum gain for a given horn length, and therefore the optimum flare angle is intermediate between slight and abrupt. In other words, an optimum horn or optimum-gain horn is one that compromises in aperture phase error to maximize its gain for a given length $L$, and therefore its flare angle is intermediate between slight and abrupt.

Graphs of radiation patterns are available for a variety of horn types and dimensions (Balanis, ch. 13, 2005; Love 1993). These graphs, known as universal patterns, were generated by aperture theory and have been validated, generally, by measurements. The “universal” patterns provide details of the major lobe shapes for large as well as small horns, with aperture widths as small as 1.5 or 2$\lambda$. Even so, designs for center-fed reflectors and lenses often require feed-horn patterns with dimensions even smaller than 1.5$\lambda$. (Section 6.3.3 provides guidance on design for horn aperture widths less than 2$\lambda$).
It is apparent that the flare angle must be made smaller as the length is increased to maintain a given maximum phase variation across the aperture. Therefore there is a practical limit to the gain that can be obtained with a horn radiator; very high gain requires an excessive horn length. For moderate gains, however, horn radiators are very useful. They are of course especially appropriate when the feed line is a waveguide. Their bandwidth is then essentially the bandwidth of the guide—2:1 typically, for the $\text{TE}_{10}$ mode in rectangular guide and a sectoral or pyramidal horn.

The published literature on optimum-gain horns includes somewhat different results for beamwidths versus horn dimensions, depending on how the horn dimensions are specified and the details of analysis. Kraus and Marhefka (2002, p. 339) provide simple equations for the beamwidths of optimum horns, that were obtained from analyses of measured beamwidths versus horn flare angle, for various horn lengths. Those equations for the 3 dB beamwidths of “optimally” tapered horns follow:

$$\theta_E = \frac{56}{d_E} \text{ degrees} \quad (4-23a)$$

and

$$\theta_H = \frac{67}{d_H} \text{ degrees} \quad (4-23b)$$

where $E$ and $H$ refer to the horn’s E and H plane patterns. The symbols $d_E$ and $d_H$ are the aperture dimensions (widths), expressed in wavelengths along the E and H planes.

Use of (4-23a) and (4-23b) does not require detailed horn dimensions, yet they provide good “first” beamwidth estimates for an “optimally” tapered horn. As already mentioned, the universal patterns are considered accurate when detailed horn dimensions are available. It is to be noted that optimum gain horns are often called standard gain horns, because of their wide laboratory usage. When detailed horn dimensions are available and more accurate calculated beamwidths are desired, the reader is referred to Bird and Love (2007, pp. 14–17). These authors also include an equation for the gain of an optimum-gain antenna, which follows

$$G = \frac{6.5d_E d_H}{\lambda^2} = \frac{6.5A}{\lambda^2} \quad (4-24)$$

where $A = d_E d_H$ is the area of the horn-mouth opening (aperture). From (3-39) of chapter 3, a recommended formula for estimating gain is

$$G = \frac{26,000}{\theta_E \theta_H} \quad (4-25)$$

By using (4-23a) and (4-23b) with (4-24), one finds a difference of only 0.3 dB in gain from that provided by (4-25).
Incidentally an unflared open-mouthed waveguide provides a slightly greater directivity than (4–24) indicates, because of its more nearly constant phase distribution. For such a radiator the factor 6.5 in (4–24) becomes approximately 10. For a typical rectangular guide $A$ is approximately equal to $\lambda^2/4$; hence $G$ is about 2.5 Thus this simple radiator has a somewhat greater gain than 1.64, the gain of a half-wave dipole.

Many special forms of horns are used. There are too many to describe in detail, but their basic principle is the same as those of the commoner types that have been discussed. One special horn that deserves a brief discussion, however, is the biconical horn, pictured in Fig. 4–22. It consists of two conical metal surfaces with their axes collinear (in line with each other) and their vertexes opposed. As with sectoral horns, there is an optimum flare angle that depends on the mode of excitation employed (Terman 1943). With the coaxial-line feed shown, the polarization will be vertical, and the optimum-flare criterion is $h = \sqrt{2s\lambda}$. Since $\sin(\phi/2) = h/2s$, the optimum flare angle $\phi$ is the one that satisfies the relation

$$\sin\left(\frac{\phi}{2}\right) = \frac{\sqrt{\lambda/2s}}{2} \tag{4–26}$$

Horizontal polarization may be excited in this horn by means of a small loop antenna with its plane perpendicular to the cone axes and lying between the vertexes, and the loop axis collinear with the cone axes. The optimum flare angle is then given by

$$\sin\left(\frac{\phi}{2}\right) = \frac{\sqrt{3\lambda/4s}}{2} \tag{4–27}$$

The radiation pattern is omnidirectional in the horizontal plane with the cone axes vertical, the vertical beamwidth and directivity depending on the dimension $h$. For optimum flare angles the directivity is given by

$$D = m\left(\frac{2h}{\lambda}\right) \tag{4–28}$$

where for vertical polarization (assuming vertical cone axes) the factor $m = 0.8$, and for horizontal polarization $m = 0.6$. The vertical beamwidths may be estimated using (4–23a) for vertical polarization and (4–23b) for horizontal polarization, with $d_E$ and $d_H$ replaced by $h$. 
The omnidirectional horizontal pattern is the attractive feature of the biconical horn, in the VHF and UHF bands, especially with horizontal polarization. Few antennas provide a truly omnidirectional horizontal-plane pattern and horizontal polarization.

4.8. Slot Radiators

If a narrow slot-like opening is cut in a large flat sheet of metal, and properly connected to a source of rf power, it will radiate in a manner that bears a certain resemblance to the radiation by a dipole of the same dimensions as the slot. In fact, if the conducting sheet is a plane of infinite extent, the radiation pattern will have exactly the same shape as that of the corresponding dipole except that the electric and magnetic vectors are interchanged. (The same relationship exists, incidentally, between the patterns of a short dipole and a small loop.) Also, the impedance properties of the slot are somewhat different. These complementary properties of a slot in an infinite plane conductor, and a thin flat dipole of exactly the same dimensions as the slot, are predicted by an important result of electromagnetic theory known as Babinet's principle. It also relates the impedance properties of the two kinds of radiators.

Figure 4–23 shows such a slot and a two-wire transmission-line feed connected to it. For the half-wavelength-long slot shown, the radiation pattern will have the same angle dependence as that of a half-wave dipole, as given by (4–11) and plotted in Fig. 4–14. Note that only the term in square brackets of (4–11) applies; i.e., the relative patterns are the same. Also, since the E and H vectors are interchanged, the polarization is opposite to that of the corresponding (complementary) dipole. It is this fact that allows the three-dimensional “doughnut” pattern to exist unaffected by the presence of the conducting plane, and on both sides of it, except for the infinitesimally thin slice eliminated by the metal sheet. That is, the required electric field intensity can exist in the space at the surface of the conducting plane because the electric field lines are perpendicular to the plane. (It is a well-known principle of electromagnetic theory that they could not so exist if they were parallel to the surface.)

Thus beamwidth and directivity are the same as for the half-wave dipole; \( \theta_{3\text{dB}} = 78^\circ \) and \( D = 1.64 \). The input impedance \( Z_s \) for the connection shown in Fig. 4–23 is related to the input impedance of a center-fed half-wave dipole, \( Z_d \), by the formula

\[
Z_s = \frac{(120\pi)^2}{4Z_d} = \frac{35,530}{Z_d}
\]  

(4–29)
Since $Z_d = 73.1 + j42.5$ ohms for a thin half-wave dipole, a thin (narrow) half-wave slot is found, from the above formula, to have impedance $Z_s = 363 - j211$ ohms. A thin slot of length $0.475\lambda$, the complement of a dipole having nonreactive input impedance of 67 ohms, will have $Z_s = 530$ ohms, nonreactive.

There is no such thing in the real world as an infinite plane conducting sheet, but if a slot is cut into a sheet that is very large compared to the slot, then the behavior predicted above will be realized to a high degree of approximation. Slots cut into sheets of even moderate size will radiate effectively, but their exact behavior is not as readily predictable.

The slot as described will radiate on both sides of the sheet. If radiation on one side only is desired, the “back” side of the slot may be enclosed by a box, or cavity. The field distribution along the slot is then affected by the dimensions of the cavity. The problem of theoretical design is quite complicated, and design is often determined by experiment.

A unidirectionally radiating slot may also be obtained by cutting it in a proper position and orientation in the wall of a waveguide. Figure 4–24 shows the appearance of several slots in a $TE_{10}$-mode rectangular waveguide that will radiate, and two that will not radiate. A waveguide slot, if it is to radiate, must be positioned so that it interrupts currents that would otherwise flow across its length in the inner walls of the guide. This is equivalent to saying that there must be a component of magnetic field ($H$) parallel to the slot at the

![FIGURE 4–24.](image)
Radiating and nonradiating waveguide slots in rectangular guide, $TE_{10}$ mode.
inner surface of the guide. The field configurations in waveguide for various modes are given in advanced engineering textbooks and handbooks. The examples of radiating slots shown in Fig. 4–24 by no means exhaust the possibilities. To be most effective the waveguide slot should be resonant, which it will be if the length is approximately half a wavelength. However, the exact length for resonance depends on the position of the slot in the guide.

A waveguide slot does not radiate the total power flowing in the guide; it "extracts" or "couples out" some fraction of it. The remaining power continues on down the guide to whatever additional load there may be further on. Or if it encounters a "short circuit" (end cap), it is reflected. By impedance matching devices it may then be possible to couple all the power to the slot. Waveguide slots, however, are seldom used singly; usually an “array” of them is cut along the length of the guide, as will be described in sec. 5.9.

Slot radiators need not be complements of half-wave dipoles. Many other radiating shapes are possible, but all may be analyzed in terms of their complementary radiators. For example, an annular-ring slot in an infinite plane sheet will have a radiation pattern exactly like that of a loop of the same size, with E-fields and H-fields interchanged. Slot radiators are practically restricted to the higher frequencies by the requirement of a conducting sheet considerably larger than the slot. They are very useful when a radiator must be devised that will not project from a surface, for example, in an airplane wing or fuselage. (Here the opening may be covered by a protective cover of low-loss dielectric material.)

4.9. Patch or Microstrip Antennas

The geometry of a microstrip is well adapted to construction by printed-circuit techniques, in which the strip conductor and adjoining ground planes can be deposited on low-loss substrate material by mass-production methods. An unbalanced microstrip geometry is illustrated in chapter 2, Fig. 2–9 (lower right). The balanced form has the strip enclosed both below and above by a metallic ground plane. This transmission-line technology can be combined with ordinary printed-circuit techniques to permit construction of monolithic phased-array modules containing not only the transmission line components, but also phase shifters, couplers, control circuitry, radiating elements, and solid-state transmitting and receiving amplifiers. This fabrication technique greatly reduces the cost as well as the bulk of large phased arrays. Moreover, the "sheet" construction need not be completely planar; it can be made to conform to the surface of a vehicle such as the fuselage of an airplane. An array thus constructed in nonplanar form is known as a conformal array.

A patch or microstrip antenna is commonly a metal rectangular or circular metal surface (the patch) on a dielectric ground plane, such as a printed circuit board. The patch or microstrip antenna is useful as a single element antenna and as the element-type within multielement arrays. Their features include lightweight and ruggedness with low profile, and bandwidths typically less than a few percent. Patch antennas can be made that radiate either a linear or a circular polarization, and the use of two operating frequencies is
Because of their ruggedness and lightweight, patches are especially useful for airborne or space applications (Howell 1975; Munson 1974). Practical patch antennas have been made for a wide range of wavelengths, covering at least from UHF into the millimeter wavelengths.

The top and side views of a rectangular patch or microstrip antenna are illustrated in Fig. 4–25. The patch is energized by a microstrip transmission line at the patch’s left side, where the dipole launches radiation. Instead, the patch can be energized by a coaxial line in a configuration called a pin fed patch. Then, the outer coaxial conductor is connected to the ground plane of the printed circuit board and the center conductor is routed through a hole in the printed circuit board and connected to the patch. Sometimes proximity coupling is used, where the connection is electromagnetic instead of direct.

Figure 4–25 depicts a patch antenna, where a wave propagates in a dielectric substrate within parallel plates between the patch and the ground plane. The parallel plate medium is a resonant structure. With a relative dielectric constant $\varepsilon_r$, the between-plates wavelength is $\frac{\lambda}{\sqrt{\varepsilon_r}}$. The length, electrically, from the left to the right patch sides is approximately one-half $\lambda$, and thus the physical length is $\approx \frac{\lambda}{2\sqrt{\varepsilon_r}}$. This produces a phase difference between the left and right edges (designated as slots 1 and 2 in Fig. 4–25) of approximately 180°, making the vertical components of the edge $E$-fields of opposite polarity. However, the edge horizontal components are in time phase. The terminology underscores the fact that the fringing of the electric fields at the patch edges is similar to the electric fields that emanate from narrow slits. Therefore, the pattern has maximum amplitude in the direction normal to the patch (i.e., upward) and it is horizontally polarized (pointed to the right).

Patch antenna design and development became an active field in the early 1970s. By January 1981, a special issue of the IEEE Transactions on Antennas and Propagation was published that contained a number of especially significant papers (see, e.g., Carver and Mink 1981; M ailloux, M cIlvenna, and K ernweis 1981). At that time, the perceived needs included, for example, increases in impedance bandwidth beyond the then available few percent, improved techniques to allow multifrequency operation, and small microstrip phase shifters for use in electronic scanning patch arrays.

Robust developments continued in the patch antenna field, resulting in reliable fabrication techniques, dramatically increased bandwidths, and reduced sizes. Novel designs now include patches that are stacked, stacked patches that are probe fed, and multiple shorting posts on stacked patches. For example, the literature describes stacked-patch antennas having bandwidths up to octave widths (Targonski, Waterhouse, and Pozar...
1998; Waterhouse 1999) and electrical sizes small enough for hand-held mobile communications at frequencies less than 2 GHz (Waterhouse, Targonski, and Kokotoff 1998).

4.10. Surface-Wave and Leaky-Wave Antennas

In the years during and since World War II some rather exotic forms of antennas have been devised, primarily in the category of VHF, UHF, and microwave antennas. Two such types of antennas that may be in the class of basic radiators are the surface-wave and leaky-wave antennas. Their theory and applications are too complex and too extensive for a full discussion here (see, e.g., Kraus and Marhefka 2002, pp. 734–41).

These antenna types are analyzed in terms of guided electromagnetic waves propagated along a surface or other guiding structure that does not fully confine them. Radiation may take place at discontinuities of the structure, or gradually along an aperture. Prominent examples of such antennas are the polyrod antenna, the cigar antenna, the zigzag, the holey plate, and the mushroom antenna. Some of these antennas find use in applications requiring a high-gain radiator with a low silhouette, as on a streamlined aircraft. They are primarily microwave or near-microwave devices, but they are by no means limited to the microwave region. In fact, the Yagi-Uda antenna, usually discussed as an array antenna (sec. 5.3), can also be regarded as a surface-wave antenna. This antenna is often used at frequencies as low as the 20 meter (14 MHz) amateur radio band.

4.11. Basic Feed Methods

The usual arrangement for feeding power to an antenna from a transmitter, or to a receiver from the antenna, is indicated in block-diagram form in Fig. 4–26. The transmission line can be any of the types discussed in secs. 2–4 and 2–5. The function of transformer A is to match the antenna feed-point impedance to the characteristic impedance of the line and also to make a transition, if necessary, from one form of line to another (e.g., from balanced to unbalanced, or from coaxial line to waveguide). Transformer B serves the same functions between the transmission line and the internal impedance of the transmitter or receiver. The two transformers insure not only that the transmitter will be correctly "loaded" by the antenna and that the antenna will be correctly loaded by the receiver, but also that there will be no standing waves on the transmission line. Elimination of standing waves is desirable, but not essential. If standing waves are to be permitted, one transformer can be omitted, and the other transformer adjusted for correct loading of the

![FIGURE 4–26.](image)

Block diagram of basic antenna feeding arrangement.
transmitter (or, in reception, optimum transfer of received signal power from the antenna to the receiver).

Sometimes the transmission line can be eliminated altogether by making a direct connection from the antenna to the transmitter output terminals or the receiver input terminals. Here also only one transformer is required. This situation can exist at very low frequencies, where a “line” of even a few hundred feet in length may be in effect a direct connection. Any conductor of length less than about a hundredth of a wavelength may be so regarded, since no appreciable standing wave pattern can exist on a conductor so electrically short. The direct connection can also exist at higher frequencies when the antenna is “built into” a receiver or transceiver.

The transformers at low frequencies are inductor-and-capacitor devices. In addition to impedance step-up or step-down, they also can provide reactance cancellation; both effects may occur in the same circuit elements or they may be separated. At the higher frequencies the transformers may be composed of transmission-line or waveguide elements, as described in secs. 2–3 and 2–5.

At frequencies up to about 30 MHz, two-wire balanced transmission lines can be used, since at these frequencies radiation due to the line spacing being a significant fraction of a wavelength is not a serious problem. Line impedances range from slightly less than 100 ohms to perhaps 800 ohms. The lower impedances are achieved with close-spaced wires embedded in low-loss polyethylene plastic. The higher impedances result with air-insulated wires or tubing. The wire lines have spacing bars of porcelain or other insulating material at intervals of a few feet or more, depending on the spacing and the wire stiffness.

When the higher-impedance lines are used to feed an ordinary half-wave dipole at its center, or a long-wire antenna at a current maximum, a transformer must be used if the impedances are to be matched. One possible type of transformer is shown in Fig. 4–27.

**FIGURE 4–27.**

An antenna installation for the HF range (3–30 MHz), showing method of center feeding a half-wave dipole with two-wire balanced line and matching stub. Line section above stub has standing wave, but with proper adjustment there is no standing wave on the feed line to left of the stub.
It is the shorted-stub arrangement whose principle is explained in chapter 2. (See Fig. 2–6.) An alternative method, which can be used when the antenna feed-point impedance is purely resistive, is the quarter-wave transmission-line transformer, whose transformation ratio is given by equation (2–22). (These methods are of course not limited to dipole and long-wire antennas; they are of general applicability.)

Another method, which can be used for feeding a wire antenna at a current maximum, is the “delta match,” illustrated in Fig. 4–28. In this method there is no gap in the antenna at the feed point; the line spacing is gradually increased until it spans a section of the antenna of length denoted as $s$ in Fig. 4–28. The effective impedance presented to the transmission line increases as $s$ is increased. This spacing is adjusted until an impedance match results, as indicated by absence of standing waves on the line. Methods of measuring the voltage standing-wave ratio (VSWR), and of thereby determining the antenna input impedance, are described in chapter 9.

Another method of obtaining an impedance match to a dipole with a two-wire line of moderate impedance is to use a folded dipole, as described in sec. 7.1. In its simplest form the folded dipole has an input impedance that approximately matches a 300-ohm line.

Although it is not absolutely essential to operate the transmission line without standing waves, it may be important to do so in high-power applications to minimize line losses due to the high currents at the current maxima and to avoid excessive voltages at the voltage maxima. Radiation losses are also greater when there are standing waves, and the loading adjustment is more critical. A line operating with an appreciable standing wave is called a resonant line; one with little or no standing wave is called nonresonant.

A half-wave dipole or a long-wire antenna can be fed at one end (maximum-voltage point) with a two-wire resonant line. The antenna is connected to one side of the line, and the other side is simply left open, as indicated in Fig. 4–29. A line used in this way is called a “Zepp” feeder. An advantage of this arrangement is that the antenna can be operated at any integral multiple of the frequency for which the antenna length is a half wavelength. The antenna pattern will be different for each such frequency, in accordance with (4–12a) and (4–12b). The value of $n$ in these equations is of course 1 for half-wavelength operation, 2 for full-wave operation, and so on. The frequency for which $n = 1$ is called the fundamental frequency of the antenna, and those for successively higher values of $n$ are called harmonic frequencies. (For example, the frequency for $n = 3$ is called the third-harmonic frequency.)
The pure Zepp feeder arrangement, shown in Fig. 4–29, omits transformer A in Fig. 4–26. The line is operated with a standing wave. When the antenna is operated at more than one of the possible frequencies (values of $n$) transformer B must in general be adjusted differently for the different frequencies. In the HF band, where such antennas are most commonly used, this transformer typically consists of a variable inductance-capacitance circuit (ARRL Antenna Book 2007). This reference and its earlier editions contain much practical information on antennas and feed systems of the type described in this chapter and on some of those described in chapter 5.

A matching circuit (i.e., transformer A) can also be incorporated at the antenna end of this type of feeder, to eliminate standing waves on the line. The stub arrangement of Fig. 4–27 (and Fig. 2–6, chapter 2) is used for this purpose. When a half-wave vertical dipole is fed in this way, the resulting arrangement is called a “J” antenna. This matching will in general be effective at only one frequency.

As the Zepp feeder indicates, “balanced” two-wire lines may be used to feed an unbalanced antenna, though at the sacrifice of perfect balance of the line currents. (Therefore such lines will radiate somewhat more than would a perfectly balanced line.) In general, however, balanced two-wire lines are preferred when the antenna is fed at a point of symmetry—where the structure is electrically balanced with respect to the ground on both sides of a center feed point. But when one side of the feed point is the ground, or a metallic ground plane, the favored transmission line for feeding the antenna is coaxial line. Examples of such antennas are monopoles, antennas whose axes are perpendicular to a ground plane in which they are imaged. All the feed methods described for two-wire lines are applicable in these cases, except that the center conductor of the coaxial line connects in the manner indicated for one of the conductors of a two-wire line, and the coaxial outer conductor connects to the base of the antenna (ground).

Transmission-line feed is appropriate to such radiators as dipoles, long wires, loops, and helixes, whose radiation is based on currents flowing in wires. Waveguide feed is more appropriate for horns and waveguide-slot antennas. However, these “rules” have
exceptions; it may at times be convenient to use a coaxial line to feed a horn—for example, when the frequency is low enough so that a waveguide would be very large and expensive, but yet high enough to make a horn radiator feasible. (This might be the case at a frequency in the region of 500 MHz.) Then a line-to-waveguide transition (described in sec. 2.5) is used, as indicated in Fig. 4–30.

Coaxial lines are also often used when the ultimate load is balanced because of their nonradiating properties and the protection that the outer conductor affords against weather and physical damage, and because they have lower losses and higher voltage breakdown rating for the same conductor spacing. A wide variety of methods for accomplishing the connection of an unbalanced (coaxial) line to a balanced (two parallel-conductor) load are possible. The device used to make an unbalanced-to-balance connection is called a balun, an abbreviation of the words balance and unbalance. Examples of baluns are shown in Fig. 4–31.

At low frequencies inductor-and-capacitor arrangements are used, as suggested by Fig. 4–31(a). At frequencies above about 100 MHz, transmission-line transformers are customary. Typically, as now briefly addressed, the basic configurations include either a $\lambda/4$- or and a $\lambda/2$-length transmission line.

Figure 4–31(b) shows “Bazooka” balun that improves the coupling between a coaxial (coax) line and a dipole (or the balanced two-conductor load). Note that, in feeding a dipole in Fig. 4–31(b), the inner and outer coax conductors are connected to the right- and left-hand arms of the dipole. Ordinarily, current within a coax at high frequency travels along the outer surface of the inner conductor and the inner surface of the outer conductor. However, because of radiation from the dipole elements and from the coax-dipole junction discontinuity, and without the bazooka balun present, additional

![FIGURE 4–30.](image-url)  
Arrangement for feeding a horn radiator with a coaxial transmission line.

![FIGURE 4–31.](image-url)  
Examples of balanced-to-unbalanced line coupling devices (baluns). (a) Coupled coils. (b) “Bazooka” balun. (c) Half-wave-line balun.
Current flows vertically along the outer surface of the outer coax conductor. This vertical current creates vertically polarized radiation, yet the radiation of the dipole, per se, is horizontal. Therefore, without the $\lambda/4$-length “bazooka” sleeve, the effective antenna gain is reduced because the antenna will simultaneously radiate horizontally and vertically polarized waves. The essence of this balun is the $\lambda/4$ length conducting coaxial section, formed by the larger diameter conducting sleeve and the outer conductor of the coax line, with the $\lambda/4$ section open at its top and connected its bottom to the outer conductor of the coaxial line. In other words, looking from the top and downwards, the impedance of the $\lambda/4$ coax (formed by the sleeve and the outer conductor of the coax line) appears as an open circuit. Thus, the vertically moving outer coaxial-line surface current, and hence the vertically polarized radiation, is appreciably reduced.

Figure 4–31c shows a half-wave-line balun for coupling between the bottom coaxial (coax) line and the balanced two-conductor line. To recognize how this balun functions, it may be helpful to recall that the voltages between points on a transmission line are equal if separated by a distance $\lambda/2$, but opposite in phase. Then, one will recognize that, at the ends of the $\lambda/2$-length coaxial line, the voltages are of equal amplitude but of opposite phase. Therefore, the voltage across the balanced line is twice that at the input coaxial line. This balun may provide a 4-to-1 impedance transformation. First, note that the two coax lines formed by the $\lambda/2$ length line are in parallel across the input coax. Thus, an impedance-matched condition may exist if the $\lambda/2$ length coax characteristic impedance is $2Z_0$, where $Z_0$ is the characteristic impedance of the input coax line. Note that the double voltage across the two output conductors is consistent with the ends of the $\lambda/2$ length line being in series with one another. Thus, an impedance matched condition exists if the two-conductor output line is terminated with an impedance of $4Z_0$, where $Z_0$ is the characteristic impedance of the input coaxial line.

Figures 4–31(b) and (c) show only simple baluns that provide connections between coaxial and two-conductor balanced lines. Munk (2002) discusses numerous other balun configurations; including those that use multiple transmission lines for increased bandwidth, as well as baluns where connections are made between coax and balanced printed lines, and between unbalanced and balanced printed lines.

References
Brainerd, J. G., ed., Ultra-High-Frequency Techniques, Van Nostrand, New York, 1942, p. 415. (The formula is there given as $R_r = 72.5 + 30\log_{10}n$.)
1. At a distance of $r = 1$ meter from a particular short-dipole antenna, the reactive component of the total electric field is equal in strength to the radiation component. At a distance $r = 10$ meters in the same direction, which of these two components is stronger? How much stronger is it? That is, what is the ratio of their strengths?

2. A thin, perfectly conducting wire is fed as a short dipole at its center; its total end-to-end length is 0.04 wavelength. The wire has no capacitive end loading so that its effective length (in terms of an equivalent elemental dipole) is less than its actual length. The rf current at the feed point (center) is 10 amp, rms. (a) What is the radiation resistance? (b) How much total power is radiated?

3. The dipole described in Problem 2 is connected to another transmitter of the same frequency (and wavelength), but of higher power output, so that the feed-point current is now 25 amp. (a) At a point half way between the center and one end of the dipole, what is the current? (b) What is it at a point at either end,
which is at a distance from the center equally 90 percent of the distance from
the center to the end? (c) What is the current at the end of the dipole?

4. A half-wave dipole in free space is center-fed with a current of 10 amp (rms).
At a distance of 1,000 meters from the dipole, which is in the far field, and in a
direction that is 60 degrees from the dipole axis, what is the field strength in
volts per meter?

5. A medium-frequency (MF) radio station (for which a long-wire horizontal
antenna is practical) is required to communicate with only four other stations.
These stations are in four different directions; one is due north, one due south,
one east, and one west. It is desired to utilize a single horizontal long-wire reso-
nant antenna that will have four major lobes, one directed at each of the four
other stations. (The major lobes are those nearest the wire axis.) (a) How many
half wavelengths long should this antenna be? (b) What are the two possible
directions of the antenna wire? Suggestions: Assume that equation (4–14) is valid
for the value of \( n \) involved. Draw a diagram showing the relative positions and
directions of the stations, and consider possible orientations of the antenna wire.
Determine from this the required value of \( \theta_{\text{max}} \) in equation (4–14). Then, using
equation (4–14), solve for \( n \). The value obtained for \( n \) will not be an exact integer;
the correct value is the integer nearest to the value found.)

6. A helical antenna with a ground plane (as in Fig. 4–20) has a turn diameter and
spacing that are appreciable fractions of a wavelength, so that it radiates in the
axial mode. The circumference of a turn, \( C \), is equal to the wavelength, \( \lambda \), and
the turn spacing \( S \) is equal to 0.25\( \lambda \). The number of turns is \( N = 16 \). (a) What is
the 3-dB beamwidth of this antenna? (b) What is its directivity?

7. A rectangular waveguide pyramidal horn has aperture dimensions, \( d_E \) and \( d_H \)
of Fig. 4–21, of one wavelength. (a) Determine approximate E- and H-plane half
power beamwidths. (b) What is the approximate directivity?

8. A slot in an infinite plane sheet of metal is in the form of an annular ring so that
it is the complement of a small loop antenna. (The central portion of the sheet,
inside the annulus, is supported by low-loss insulating material.) (a) Is the
maximum radiation of this annular slot in the direction perpendicular to the plane
of the sheet, or is it parallel to it? (b) What is the polarization of the field near
the metal sheet at a point distant from the slot?

9. A thin vertical monopole antenna is 95 percent of a quarter wavelength in height
above a ground plane of infinite extent and perfect conductivity. It is fed at a
gap at its base by a coaxial line. No transformer is used at the feed point, yet
there are no apparent standing waves on this line. What is the approximate
characteristic impedance of the line?

10. Several means of delivering power to an antenna from a transmitter are listed
below, and identified by capital letters:

(A) Coaxial line.

(B) Balanced two-wire line followed by an impedance transformer to provide
nonresonant operation (no standing waves).
(C) Waveguide.
(D) Direct connection.
(E) Resonant two-wire line.

After each of the antenna types below, write in one of the above capital letters to indicate which form of line or connection you consider most appropriate. (Use each letter once and only once.)

(i) Automobile radio antenna for broadcast-band reception (535–1,605 kHz)
(ii) Helical antenna with ground plane
(iii) Long-wire antenna fed at one end
(iv) Horn radiator
(v) Half-wave dipole having 20-meter length, fed at gap in center
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