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Faculty of Engineering

Department of Electrical and Electronic Engineering

TV AND FM TRANSMITTING ANTENNAS

Graduation Project

EE - 400

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ABSTRACT

We had thought to do our work on the antenna, and then we search for the important parts on this subject since the antenna is one of the most common and important parts in the communication system.

The term antenna is defined by the dictionary as a usually metallic device (as a rod or wire) for radiating or receiving radio waves. The official definition of the Institute of Electrical and Electronics Engineers (IEEE) is simply as a means for radiating or receiving radio waves. The ideal antenna is, in most application, one that will radiate all the power delivered to it by a transmitter in the desired direction or directions and with the desired polarization.

The objective in this chapter is to provide the reader with specifications and descriptions of television and FM radiobroadcast antennas. The various antennas to be described also may be used for other applications in the frequency range of 10 MHz to 10 GHz. Broadcast antennas have frequency, pattern, power capacity, impedance, and environmental requirements which are imposed by regulatory agencies such as the Federal Communications Commission (FCC) or by system specifications. For instance, frequency and pattern are regulated by the FCC, while impedance, power capacity, and environmental requirements are system-related.

TABLE OF CONTENTS

INTRODUCTION

The first antennas were employed by Hertz in 1887 in his classic demonstrations of the electromagnetic waves that had been predicted earlier by Maxwell. Hertz's receiving antenna was a circular loop of wire broken by a microscopic gap. The radius of the loop was 35 cm, a dimension that had been found by experiment to put the loop into resonance with the transmitter. Hertz later placed his rod antenna in the focal plane of a cylindrical mirror. About 8 years after the early work of Hertz, a Professor Popoff of Kronstadt was engaged in a study atmospheric electricity, and in connection with this study, he placed a receiving antenna, consisting of a metallic rod, above this housetop. The receiver ("coherer") was connected between this rod and the earth.

In 1897, Marconi described a complete system for wireless telegraphy. In this system, one terminal of the spark transmitter was connected to an elevated wire and the other terminal was connected to the earth. Apparently, Marconi was the first to realize the importance of elevating the transmitting antenna. The transmitting antenna for the first wireless transatlantic communication (from Poldhu in Cornwall to Newfoundland, in 1901) was a vertical fanlike structure consisting of 50 vertical copper wires supported by a horizontal wire. The horizontal wire was stretched between two masts about 150 ft high and 200 ft apart. The receiving antenna in Newfoundland was supported by kites. By 1907, commercial telegraph services had been established, and the advantage of top-loaded antennas was widely recognized. The frequency of operation of these early communications systems was usually in the range from 50 to 100 kHz, and consequently the antennas were small compared to the wavelength. Unlike the transmitters of today, the antennas that were employed in the early systems usually strongly influenced the operating frequency.

One of the earliest papers on the subject was that of Abraham, who, in extending some earlier work on spheres by J.J. Thomson, studied the natural oscillations of a conducting prolate spheroid. He determined the natural frequencies and calculated the fields of a half-wave dipole. Hertz himself had studied the fields of point dipoles. His work was carried on further by Sommerfeld and others, and by 1914, Hertz potentials and vector potentials had been employed extensively in calculations of the radiation patterns of known current. Further, Poynting's theorem had been employed to calculate the total power radiated from antennas together with their radiation resistances. Interest in resonant length antennas (half-wavelength dipoles or quarter-wavelength monopoles above ground) began to grow about 1920, after the discovery that the De Forest triode tube could be made to produce continuous wave oscillations at the higher frequencies (hundreds or even thousands of kilohertz). A these higher frequencies, it become practical to construct resonant length antennas or even arrays of these. By about 1930, the theory and practice of simple linear arrays had been developed and applied to broadcast transmitters for interference control.

As antennas that were of the order of a wavelength came into use, the need for a better understanding of the interaction of the antennas with transmission lines and transmitters grew more pressing. Outstanding contributions on this subject were made by J. R. Carson, who presented a generalization of the reciprocity theorem, and by P. S. Carter, who published antenna terminal-impedance definitions and calculations. Later in the thirties the treatment of the antenna as an electromagnetic boundary-value problem was revived. King and Hallen formulated the linear-antenna problem as an integral equation. Stratton and Chu employed a prolate spheroidal model for the linear antenna and deduced some of its properties by means of spheroidal wave functions. Perhaps the most illuminating model was introduced by Schelkunoff. According to his initial model, the straight-wire antenna was regarded as a limiting case of a biconical horn antenna. With such a model, the solutions may be expressed in spherical coordinates.

In the mid-thirties, a new branch of antenna technology began to develop. The developments went hand in hand with the development of generators in the microwave frequency range and the use of metallic pipes as waveguides. These waveguides were flared out into horns, a rather natural step by analogy with the corresponding acoustic problem. Later, radiating slots were introduced into the walls of the waveguides. Still later, the World War II requirements for special high-gain antennas led to the development of large parabolic reflectors and lenses. Schelkunoff also published a beautiful generalization of linear array theory.

The commonly used "whip" antennas on cars "rabbit ears" on TV receivers single turn loop antennas for UHF TV reception roof mounted log-periodic TV antennas and satellite paraboloidal reflector receiving antennas are so prevalent that most people are clearly aware of need for antennas in the support of our daily communication needs. These commonly occurring antennas represent only a small segment of the antenna systems that have been developed. For specialized and high performance communication links, radar systems, navigational systems, and scientific studies highly complex antenna systems are needed.

The objective in this project is to provide the reader with specifications and descriptions of TV and FM radiobroadcast antennas. The various antennas to be described also may be used for other applications in the frequency range of 10 MHz to 10 GHz. Broadcast antennas have frequency, pattern, power capacity, impedance, and environmental requirements which are imposed by regulatory agencies such as the Federal Communications Commission (FCC) or by system specifications. For instance, frequency and pattern are regulated by the FCC, while impedance, power capacity, and environmental requirements are system-related.

Broadcast frequencies in the United States are allocated and regulated by the FCC. The following frequency bands are assigned to television broadcasting:

Table 1. Frequency bands in TV broadcasting.

Low VHF	Channels 2-4	54-72 MHz
	Channels 5-6	76-88 MHz
High VHF	Channels 7-13	174-216 MHz
UHF	Channels 14-83	470-890 MHz

Each channel is assigned 6 MHz of bandwidth, with visual carrier and color subcarrier at 1.25 and 4.83 MHz above the lower edge of the channel, respectively, and with aural carrier at 0.25 MHz below the upper edge of the channel. The power levels of the visual subcarrier and aural carrier usually are within 20 percent of the visual carrier.

FM radio frequencies are limited to the band between 88 and 108 MHz. There are 100 channels, each with a 200-kHz bandwidth. Pattern requirements are functions of coverage goals, site location, local terrain, and the available options on mounting structures. Coverage goals are regulated and limited by FCC specifications as spelled out in the *Code of Federal Regulations (CFR* 47). For instance, a TV station's coverage is specified by the distance to the "city grade," "grade A," and "grade B" contours. Table 28-1 shows the minimum levels of the field strength (present at 50 percent of locations 50 percent of the time) assigned to these contours. The transmitter location and power and the antenna height and gain are chosen such that the city-grade contour covers the entire principal community to be served.

For an FM station, the coverage limits are defined by two contours. These are the 70-dBu contour (3.16 mV/m), for city grade, and the 60-dBu contour (1.0 mV/m). These limits, along with the path-loss curves (also known as the 50-50 curves), which are docu mented in *CFR* 47, are used to determine the ERP* and/or the gain of the antenna for a given antenna height and location. The FCC manual specifies the maximum ERP for TV and FM stations. The maximum power[2] varies with regions or zones of the United States and with the antenna height above the average terrain (HAAT). Towers that support broadcast antennas are either self-supported or guyed, and their heights range from 100 to 2000 ft.
Since broadcast antenna structures usually consist of vertical arrays of radiating elements, antenna 'directivity is approximately equal to the product of azimuth-pattern directivity and the elevation directivity (the number of bays in some antennas). The antenna gain is always referenced to the gain of a half-wave dipole (2.15 dB above the isotropic element) and is equal to the directivity less the losses, such as impedance loss and/or polarization-mismatch loss.The majority of applications call for omnidirectional azimuth patterns. The circularity of the pattern depends on the type of antenna when top-mounted and also on the

Table 2. Grade contours of TV broadcast stations.

Channel	City grade (dbu)	Grade A (dbu)	Grade B (dbu)
2-6	74	68	47
7-13	77	71	56
14-69	80	74	64

configuration of the support structure when side-mounted. Other requirements call for various types of azimuth patterns, such as cardioid, skull-shaped, peanut-shaped, etc., to protect other stations or reduce radiation into low-population areas. An azimuth pattern with a circularity of ±2 dB is considered omnidirectional. The maximum to minimum ratio of directional patterns should not exceed 15 dB for FM and UHF TV (with the ERP greater than 1 kW). For Channels 2 to 13 this ratio is 10 dB.

TV and FM broadcasting was originally limited to horizontal polarization. In the 1960s, the FCC allowed circular polarization (CP) for FM broadcasting. This provided improved reception, especially for vehicles with whip antennas, which are predominantly vertically polarized. In 1977, the FCC permitted TV broadcasting in right-hand CP as well. In going from horizontal polarization to CP, the stations were allowed to maintain their maximum ERP in horizontal polarization so as to maintain the field strength existing before the conversion. By allowing the same ERP for vertical polarization, the FCC actually allowed doubling the radiated power. This has provided improved reception for receivers with indoor antennas such as monopoles and rabbit ears.

In some instances, use of CP has reduced ghosting because reflections from buildings and other objects tend to have the opposite sense of CP. The acceptable axial ratio for CP antennas is 3 dB or less. The receiving antennas are almost all linearly polarized, and because of this and the unfriendly propagation path of most broadcast environments, the importance of axial ratio is somewhat superfluous. For instance, the majority of FM antennas are omni directional CP antennas side mounted on towers without regard to the effect of the tower on the axial ratio of the radiated CP wave. FCC regulations are limited to the shape and directivity of the vertical and horizontal polarization patterns and do not specify the axial ratio of the radiated CP wave

Chapter one is primary concerned with definitions and releated terminology. There is an explanation of antenna parameters and structure with some equations and the figures. Chapter two gives an information about the types of broadcasting antennas such as circularly polarized antennas. Chapter three presents an explanation of horizontally polarized antennas and FM antennas. Finally, in chapter four we give a conclusion of the project.

CHAPTER ONE

ANTENNA FUNDAMENTALS AND STRUCTURES

1.1Antenna Structure

The structure of the antennas depends upon the type and the destination, but in general, all antennas have the following structure.

1.1.1 Size

The size of antenna range from microminiature to gigantic, and it depends on the wavelength, which has proportionality with the operations frequency, and this relationship is simple and fast.

The large antennas are used for low frequencies (high wavelength), and vice versa, small antennas are used for high frequencies (low wavelength), but sometimes-large antennas are used at short wavelength (high frequencies) to obtain a highly directional radiation pattern and high gain in a preferred direction.

In practice field, the increasing of the size is limited, because at determining size, there is no point in increasing this size because it produce a little or no additional gain and the required precision of construction or maintenance of phase relationship is not attainable. Moreover, very small antennas can be used at long wavelength, when efficiency is not important. In general, the largest antennas are used at the VLF, especially for transmitting, where radiation efficiency is important. As an example of the extremely large VLF antenna is Navy's installation that has tower 1000 feet high, extends over an area of 2 square miles. In contrast, a half wave dipole at the microwave frequencies may be considerably less than an inch long.

1.1.2 Supports

There must often be some supporting structure to place the radiating element or elements in a clear location (with often is synonymous with a high location). Such devices as towers, masts, and pedestals support antennas.

Towers are used when great height is required. Masts may be quite high, but they are often as short as a few feet. Pedestals are the base structures of antennas such as reflectors and lenses, for which height is not important as strength. Sometimes an antenna may be mounted directly on a vehicle, such as an automobile, ship, aircraft, or spacecraft, where no intermediate support is required. Moreover, towers and masts are sometimes themselves used as antennas rather than as supports. In the standard broadcast band (550-1600KHz). As an example, vertical towers of heights up to several hundred feet are used as transmitting antennas.

1.1.3 Feed Line

We can simply define the feed lines as the transmission lines. These lines are used to connect the transmitter or receiver to the antenna. The design of the feed lines and any necessary impedance matching or power-dividing devices associated with it is one of the most important problems in the calculation of antenna design. At the very lowest frequencies the earth (ground) is a part of the antenna electrical system. Therefore, one terminal of the antenna input is a rod driven into the ground or a wire leading to a system of buried conductors, especially if the earth is dry in the vicinity of the antenna. The other terminal is then usually the base of a tower or other vertically rising conductor. Towers used in this way are usually supported at the base by a heavy insulator or insulators (series feed), but occasionally they are directly grounded and fed by connecting the feed wire a short distance up from the ground (shunt feed).

At somewhat higher frequencies, up to (up to 30MHz), the antenna may be a horizontal wire strung between towers, or other supports (from which it is insulated). The feed line is then often a two-wire balanced line connected at the center of the antenna, either to the two terminals provided by a gap in the antenna wire (series feed), or to two points somewhat separated on the unbroken antenna *wire* (shunt feed). Sometimes the feed line is connected at the end of the horizontal span, or elsewhere of center, but center feed is preferred because it results in better balance of the currents in the feed wires. The spacing between the two-wire-line is range from less than an inch to 12 inches or more. The last method is used for high frequencies. But coaxial feed lines are commonly used for upper high frequencies UHF (up to 1 GHz), because the two-wire-line spacing becomes too great a fraction of the wavelength to prevent appreciable radiation and because waveguides below 1000MHz are quite large and expensive. Coaxial line diameters range from a fraction of an inch up to 9 inches or more. Above 1000MHz, waveguides are commonly use, with some use of mall-diameter coaxial lines in low-power noncritical applications.

We should mention that, when the antenna rotates on a pedestal, or has other motion with respect to its support, the feed line must contain flexing sections or rotating joints, this require is quite important on the antenna measurement operations, as we will see later.

1.1.4 Conductors

Metals are the usual conducting materials of antennas. Metals of high conductivity, such as copper and aluminum (and its alloys), are naturally preferred. Brass may be used for machined parts. Magnesium is sometimes used where ultralight weight is important, usually in an alloy and with a protective coating or treatment. The steel may be used, when the strength is of primary importance, either with or without a coating or plating of copper. The conductivity of unplated steel is adequate when it is used in the form of sheets or other large-surface-area forms (as for the surface of a paraboloidal reflector). Antenna ware is sometimes made with a steel core for strength and to minimize stretching and with a copper coating to increase the conductivity. Such wire is virtually as good a conductor as solid copper. Since the radio frequency RF currents are concentrated near the surfaces of conductors (skin effect). For this reason brass and other metals are sometimes silver plated when exceptionally high conductivity is required. For the same reason large-diameter conductors may be hollow tubes without loss of conductivity. At low radio frequencies the conductivity of large-diameter conductors may be increased, compared to a solid conductor, by interweaving strands of small-diameter insulated wires; the resulting conductor is called Litz wire. This technique is most effective below about 500KHz. At higher frequencies it is not effective because the currents tend to flow only in the outer strands.

Conductor size in antenna design is determined by many factors, principally the permissible ohmic losses and resultant heating effects in some cases, mechanical strength requirements, permissible weight, electrical inductance and capacitance effects, and corona considerations in high-voltage portions of transmitting antennas. Large-diameter conductors minimize the Corona, by avoidance of sharp or highly curved edges, and by using insulators with metal end caps bonded to the insulating material, so that small air gaps between wires and insulators do not exist. Corona can occur on metal supports of the antenna as well as on the antenna conductor itself, as a result of induced voltages.

1.1.5 Insulators

The conducting portions of an antenna not only carry RF currents but also have RF voltages between their different parts and between the conductors and ground. So that, to avoid the short circuiting these voltages, insulators must sometimes be used between the antenna and its supports, or between different parts of the antenna. The insulators are also used as spacer supports for two-wire and coaxial lines and to break up guy wires with masts and towers to prevent the resonant or near-resonant lengths. The maximum permissible uninterrupted length of guy wire sections is about 1/8 wavelength. Also, the insulators are used to support long heavy spans of wire, so that it must be high strength. Typical insulating materials for such insulators are glass and ceramics, other (low loss) materials such as polystyrene and other plastics are used where less strength is required. Very large and heavy insulators are necessary in high-power transmitting applications to prevent flashover. Coaxial lines and waveguides in high-power applications may be filled with an inert gas, or dry air, at a pressure of several atmospheres, to increase the voltage-breakdown.

1.1.6 Weather Protection

The antennas are ordinarily out doors, so that, it must withstand wind, ice, snow, lightning, and sometimes corrosive gases or salt-laden air. Protection against wind and ice loads is primarily a matter of mechanical strength and bracing. Guy wires are used with tall structures or towers, to prevent their overturning in high winds. In the heavy current networks, the ice is sometimes melted from the heating that is produced from the current. Sometimes an antenna is totally enclosed in a protective housing of low-loss insulating material, which is practically transparent to the electromagnetic radiation. Such housing is called radome. Radomes are commonly used on some types of aircraft antenna for aerodynamic reasons. The protection against lightning-induced currents, and static-charge buildup is necessary for some types of antennas such as broadcasting towers, or any structure that stands high above its surrounding, if the conducting path to ground is not heavy, and direct. Insulators may be protected by horn or ball gaps, and static may be drained by connecting high-ohmage resistors across insulators.

1.2Antenna Parameters

The most fundamental properties of antennas are the following

1.2.1 Polarization

The wave polarization refers to the instantaneous component direction on a surface perpendicular to the direction of energy propagation. In the communication system only sinusoidal varying fields are ordinary used. The radiation of an antenna may be linearly, elliptically, or circularly polarized. Polarization in one part of the total pattern may be different from polarization in anther. As an example, in the case of a directional antenna with a main beam and minor lobes, the polarization may be different in the minor lobes and in the main lobe, or may even vary in different parts of the main lobe.

The simplest antennas radiate (and receive) linearly polarized wave. They are usually oriented so that the polarization (direction of the electric vector) is either horizontal or vertical. But sometimes the choice is dictated by the necessity, at other times by preference based on technical advantages, and sometimes there is no basis for choice one is as good and as easily achieved as the other. For example at the very low frequencies it is practically difficult to radiate a horizontally polarized wave successfully because it will be virtually cancelled by radiation from the image of the antenna in the earth, also vertically polarized waves propagate much more successfully at these frequencies (e.g., below 1000KHz). Therefore vertical polarization is practically required at these frequencies.

At the frequencies of television broadcasting (54 to 890MHz) horizontal polarization has been adopted as standard. The standard frequency is very important to determine the type of polarization. Otherwise, we have to design an antenna such has both polarizations, thus greatly complicating design problem and increasing the received noise level.

At the microwave frequencies (above 1GHz) there is little basis for a choice of horizontal or vertical polarization. Also in specific applications there may be some possible advantages in one or the other. Of course in communication it is essential that the transmitting and receiving antennas have the same polarization.

Circular Polarization has advantages in some VHF, UHF, and microwave applications. As an example, in transmission of VHF and low-UHF signals through the ionosphere, rotation of polarization vector occurs, the amount of rotation being generally unpredictable. Therefore if a linear polarization is transmitted it is advantageous to have a circularly polarized receiving antenna which can receive either polarization, or vice versa. The maximum efficiency is realized if both antennas are circularly polarized.

From the above explanation. It is obvious that in communication circuits it is essential that transmission and receiving antennas have the same polarization. Also it is apparent that the polarization properties of any antenna are an important part of its technical description (parameter of its performance). Sometimes it may be desirable to provide polarization pattern of the antenna, that is, a description of the polarization radiated as a function of the direction angles of a spherical coordinate system, although such a complete picture of the polarization is not ordinarily.

1.2.2 Radiation Pattern

The radiation pattern of an antenna is one of its most fundamental properties, and many of its performance parameters pertain to various aspects of the pattern.

We should mention that antennas have a reciprocal relationship between the processes of radiation and reception, so, it is customary to speak of the antenna pattern as radiation pattern, and a reception pattern as well because it also describes the receiving properties of the antenna. The radiation pattern describes the relative strength of the radiated field in various directions from the antenna, at a fixed or a constant distance.

Because the antenna pattern is three dimensional, a three-dimensional coordinate system is required. So, either Cartesian (rectangular) coordinates (x, y, z) or spherical coordinates (R, θ, ø) is used. The spherical coordinate system is an appropriate coordinate system to describe the antenna pattern because the radiation pattern may be expressed in terms of the electric field intensity, (for example, at some fixed distance R from the antenna), at all points on the spherical surface at that distance. Spherical points on the surface are then defined by the direction angles θ and Φ. The pattern then becomes a function of only two independent variables, since R is a constant, and this fact greatly simplifies the matter.

Figure 1-1 illustrates the relationship between the Cartesian and spherical coordinates. The projection of this distance r onto the xy-plane is designated θ, Φ, this means that changing r causes changing on θ, Φ.

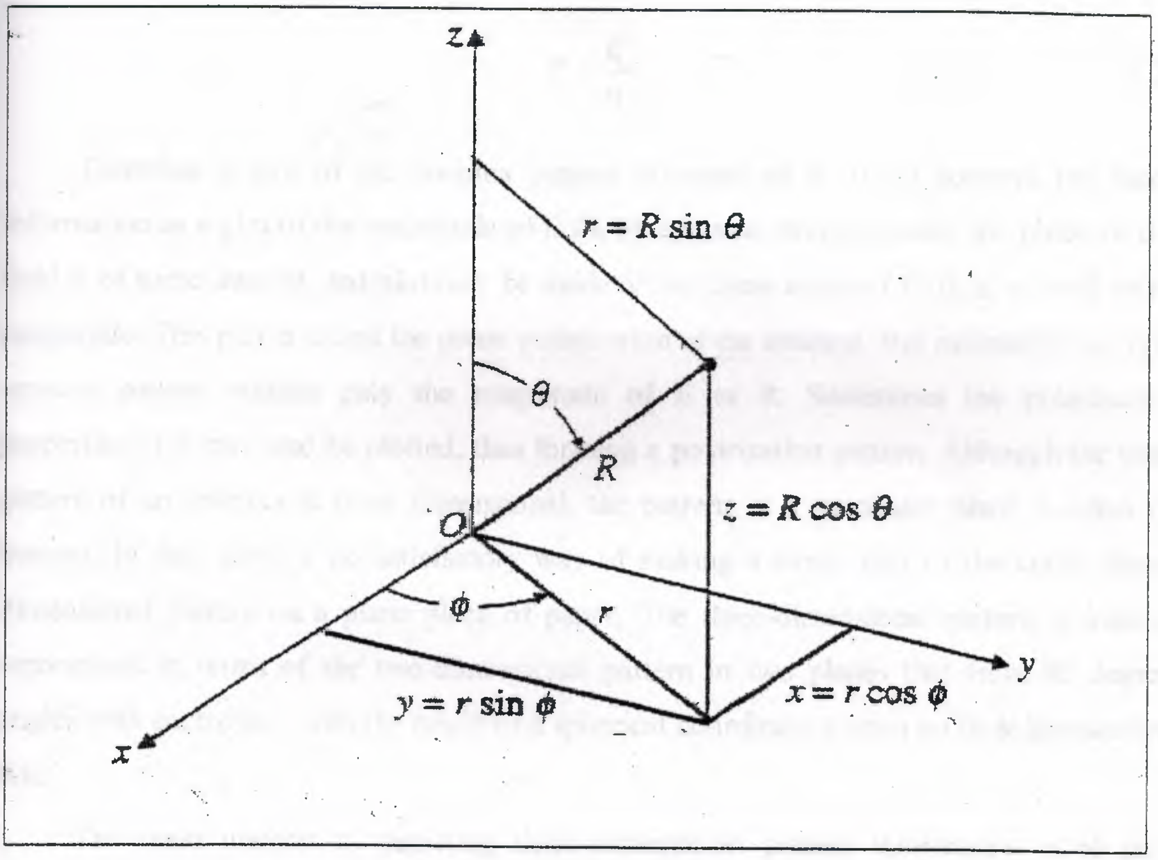

Figure 1.1 Showing interrelationship of space variables (x, y, z) and (R, θ, $\o$).

An antenna is supposed to be located at the center of a spherical coordinate system, its radiation pattern is determined by measuring the electric field intensity over the surface of a sphere at some fixed distance, R. Since the field E is then a function of the two variables θ and ø, so it is written E (θ, ø) in functional notation.

A measurement of the electric field intensity E (θ, ø) of an electromagnetic field in free space is equivalent to a measurement of the magnetic field intensity H (θ, ø), since the magnitudes of the two quantities are directly related by

$$E = \eta_0 H \tag{1.1}$$

(Of course, they are at right angles to each other and their phase angles are equal) where

$\eta_o = 377\Omega$ for air. Therefore the pattern could equally be given in terms of E or H.

The power density of the field, P (θ, ø), can also be computed when E (θ, ø) is known, the relation being

$$P = \frac{E^2}{\eta} \qquad (1.2)$$

Therefore a plot of the antenna pattern in terms of P (θ, ø) conveys the same information as a plot of the magnitude of E (θ, ø). In some circumstances, the phase of the field is of some interest, and plot may be made of the phase angle of E (θ, ø) as well as its magnitude. This plot is called the phase polarization of the antenna. But ordinarily the term antenna pattern implies only the magnitude of E or P. Sometimes the polarization properties of E may also be plotted, thus forming a polarization pattern. Although the total pattern of an antenna is three dimensional, the pattern in a particular plane is often of interest. In fact, there is no satisfactory way of making a single plot of the entire three-dimensional pattern on a plane piece of paper. The three-dimensional pattern is usually represented in terms of the two-dimensional pattern in two planes that from 90 degree angles with each other, with the origin of a spherical coordinate system on their intersection line.

The main method of depicting three-dimensional pattern information is to plot contours of constant signal strength on the surface of a sphere containing the antenna at its center. But ordinarily only the principal plane patterns are given, as they convey an adequate picture of the three-dimensional pattern for most purposes.

Pattern in a plane involves only one angle, so that, it is represented by polar coordinates, it would be possible to use Cartesian coordinates. If this were done, the shape of the pattern would be unchanged; but because interpretation of the meaning of the pattern in terms of the Cartesian coordinates would be relatively difficult, this is never done. It is fairly common to plot the pattern on rectangular-coordinate graph paper but in terms of the direction angle as the abscissa and field strength or power density as the ordinate. This type of plot distorts the appearance of the pattern geometrically but preserves the interpretability of an angle representation and makes the plotting and the reading of the low amplitude portions of the pattern easier. Figures 1 -2a and 1 -2b compare these two representations.

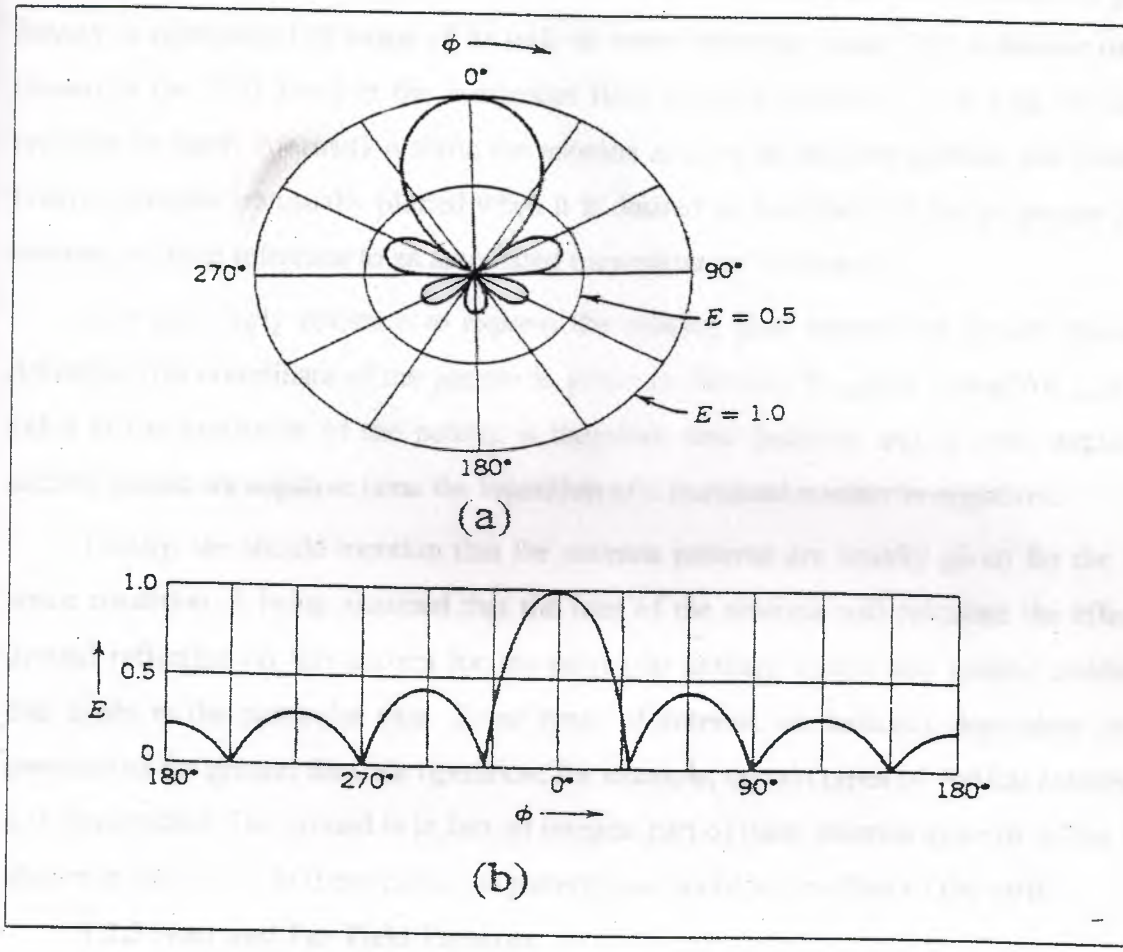

Figure 1.2 Comparison of plane pattern plotted in polar and rectangular form. The same pattern is represented in both cases and the coordinates are the same. Only the plot is different (a) polar (b) rectangular plot.

Note that it is easier to locate the angular positions of nulls (zeros) of the pattern on the rectangular plot.

If the radiation pattern is plotted in terms of the field strength in electrical units, such as volts per meter or the power density in watts per square meter, it is called an absolute pattern. An absolute pattern actually describes not only the characteristics of an antenna but also those of the associated transmitter, since the absolute field strength at a given point in space depends on the total amount of power radiated as well as on the directional properties of the antenna.

Often when the pattern is plotted in relative terms, that is, the field strength or power density is represented in terms of its ratio to some reference value. The reference usually chosen is the field level in the maximum field strength direction. This type of pattern provides as much information about the antenna as does an absolute pattern, and therefore relative patterns are usually plotted when it is desired to describe only the properties of the antenna, without reference to an associated transmitter (or receiver).

It is also fairly common to express the relative field strength or power density in decibels. This coordinate of the pattern is given as $20\log(E/E_{max})$ or $10\log(P/P_{max})$. The value at the maximum of the pattern is therefore zero decibels, and at other angles the decibel values are negative (sine the logarithm of a fractional number is negative).

Finally, we should mention that the antenna patterns are usually given for the free-space condition, it being assumed that the user of the antenna will calculate the effect of ground reflection on this pattern for the particular antenna height and ground conditions that apply in the particular case. Some types of antenna are basically dependent on the presence of the ground for their operation, for example, certain types of vertical antennas at low frequencies. The ground is in fact an integral part of these antenna systems as has been shown in Sec. 1.1.3. In these cases, the pattern must include the effect of the earth.

1.2.3 Near and Far Field Patterns

In principle it is possible to calculate the values of the electric and magnetic field components set up in space by any antenna.

The mathematical difficulties may be formidable if the antenna is complicated, but the calculation is always possible in principle when we use Maxwell's equations. For some simple types of antennas such calculations may be carried out in considerable detail, and the results illustrate certain features that apply to all antennas and are confirmed by experimental investigations of antenna fields. One such feature is that the radiation pattern in the region close to the antenna is not exactly the same as the pattern at great distances. The term near field refers to the field pattern that exists close to the antenna; the term far field refers to the field pattern at great distances. The significance of these terms is conveniently illustrated by considering the fields set up by a simple dipole antenna. The mathematical analysis reveals that in a given direction the total electric field can be expressed as the sum of three terms, each of which decreases in magnitude as the distance from the antenna, R, increases; but they decrease at different rates. The electric field intensity is inversely proportional to the first power of the distance. The dipole field is found to have components that decrease inversely as the square of the distance and inversely as the cube of the distance, in addition to the inverse-first-power term. Mathematically this means that one term contains factors $1 / R$, $1 / R^2$, and $1 / R^3$. The behavior of such terms, as R increases, is illustrated in Fig. 1-3. These terms are equal in magnitude at $R = 1$. Or smaller values of R, the factor $1 / R^3$ is largest, and the $1 / R$ term is smallest. But for large values of R, the $1 / R$ factor is larger than the other two, becoming increasingly so as R increases.

Practically in the far zone the field consists of only the term containing the $1 / R$ factor. The field at great distance from the dipole behaves like the field of point source, with inverse-first-power dependence of the electric field intensity on the distance from the dipole.

At very close distance, on the other hand, $1 / R^3$ and $1/R^2$ terms becomes much larger than the $1 / R$ term dominates the far-field region, as seen in Figure 1-3

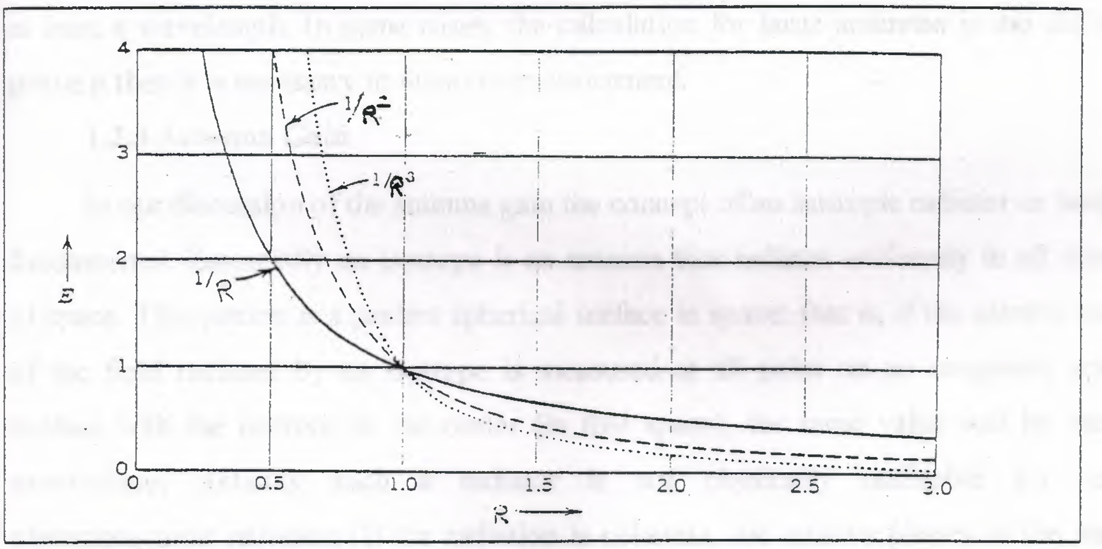

Figure 1.3 Relative variation with distance of short-dipole static ($1 / R^3$), induction ($1 / R^2$), and radiation ($1/R$) field components (electric intensity).

For more complicated antennas, the near field has more complicated dependence on R. The near-and far-field pattern is in general different; that is, plots of relative field strength at a constant distance do not have the same form. In fact, the pattern taken at different distances in the near field will differ from one another, but all patterns taken in the far field are alike, ordinarily it is the radiated power that is of interest, and so antenna patterns are usually measured in the far field region. For pattern measurement it is therefore important to choose a distance sufficiently large to be definitely in the far field, well out of the near field. The minimum permissible distance depends on the dimension of the antenna in relation to the wavelength. An accepted formula for this distance is

$$R_{\min} = \frac{2d^2}{\lambda} \tag{1.3}$$

where R_{Min} is the distance from the antenna, d is the largest dimension of the antenna, and λ is the wavelength.

The factor 2 in this expression is somewhat arbitrary, but it is the factor usually observed in antenna measurement practice. The formula also assumed that d is at least equal to about a wavelength, when d is smaller than λ the distance R_{min} should be equal to at least a wavelength. In some cases, the calculation for large antennas is too difficult to prove it then it is necessary to resort to measurement.

1.2.4 Antenna Gain

In our discussion of the antenna gain the concept of an isotropic radiator or isotrope is fundamental. Essentially an isotrope is an antenna that radiates uniformly in all directions of space. This pattern is a perfect spherical surface in space; that is, if the electric intensity of the field radiated by an isotrope is measured at all point on an imaginary spherical surface with the isotrope at the center (in free space), the same value will be measured everywhere. Actually such a radiator is not physically realizable for coherent electromagnetic radiation (If the radiation is coherent, the relative phases of the waves in different directions from the source maintain a constant difference. For a noncoherent radiator, these phase difference vary in a random manner, or fluctuate. The sun is an example of a noncoherent radiator) all actual antennas have some degree of non-uniformity in their three-dimensional radiation pattern. It is possible for an antenna to radiate uniformly in all directions in a plane, and to design an antenna that has approximate omnidirectionality in three dimensions, but perfect omnidirectionality in three dimensional space can never be achieved. Nevertheless, the concept of such an ideal omnidirectional radiation, an isotrope, is most useful for theoretical purposes. A nonisotropic antenna will radiate more power in some directions than in others and therefore has a directional pattern. Any directional antenna will radiate more power in its direction (or directions) of maximum radiation than an isotrope would, with both radiating the same total power. It is intuitively apparent that this should be so, since the directional antenna sends less power in some directions than an isotrope does, it follows that it must sent more power in other directions, if the total powers radiated are to be the same. This conclusion will now be demonstrated more rigorously. If an isotrope radiates a total power P_t and is located at the center of a transparent (or imaginary) sphere of radius R meters, the power density over the spherical surface is shown bellow

$$P_{isotrope} = \frac{P_t}{4\pi R^2} \quad (\mathrm{W/m^2}) \tag{1.4}$$

Since the total P_t is distributed uniformly over the surface area of the sphere, which is $(4\pi R^2)$ (m^2).

Imagine that in some way it is possible to design an antenna that radiates the same total power uniformly through one half of the same spherical surface, with no power radiated to the other half. Such a fictitious radiator may be called a semi-isotrope. Since the half sphere has a surface area $(2\pi R^2)$ square meters, the power density is

$$P_{semi-isotrope} = \frac{P_t}{2\pi R^2} \quad (\mathrm{W/m^2}) \tag{1.5}$$

Therefore, we get

$$\frac{P_{semi-isotrope}}{P_{isotrope}} = \frac{(P_t / 2\pi R^2)}{(P_t / 4\pi R^2)} = 2 \tag{1.6}$$

The last result shows that at any distance, *R,* the power density radiated by the semi-isotrope is twice as great as that radiated by the isotrope, in the half-sphere within which the semi-isotrope radiates. In this region, therefore, the semi-isotrope is said to have a directive gain of 2. It is fairly apparent that if the radiation were confined to smaller portions of the total imaginary spherical surface, the resulting directive gain would be greater. For example, if the power P_t uniformly into only on fourth of the spherical surface, the directive gain would be 4, and so on.

1.2.4.1 Directive Gain

The directive gain D, of an antenna is defined, in a particular direction, as the ratio of the power density radiated in that direction, at a given distance, to the power density that would be radiated at the same distance by an isotrope radiating the same total power. The directive gain of a semi-isotrope in the hemisphere into which it radiates is 2; its directive gain in the other hemisphere (where no power is radiated) is zero,

Thus D of an antenna is defined as a quantity that may be different in different directions. In fact, the relative power density pattern of an antenna becomes a directive gain pattern if the power density reference value is taken as the power density of an isotrope radiating the same total power (instead of using as a reference the power density of the antenna in its maximum radiation direction). In this case, we define the direction gain of the antenna as

$$D = \frac{P_{antenna}}{P_{isotrope}} \tag{1.7}$$

were P antenna is the antenna power density, from Eqs.1-2 and 1-4, we find that

$$D = \frac{4\pi R^2 E^2}{377 P_t} = \frac{4\pi R^2 P_{antenna}}{P_t} \tag{1.8}$$

where P_t is the total radiation power. If P_t represents the input power to the actual antenna rather than the power radiated, G should be substituted for D on the left hand side of this equation, that is, give the power gain rather than the directive gain. The efficiency factor ξ is the ratio of the power radiated by the antenna to the total input power, it is a number between zero to unity, and it connects the direction gain D with the power gain G in

$$G = \xi D \tag{1.9}$$

The maximum directive gain (directivity) is quite important value, as we will see in gain measurement later. This value can be calculated from

$$D_{\max} = \frac{4\pi}{\int_0^{2\pi}\int_0^{\pi}[E(\theta,\phi)/E_{\max}]^2 \sin\theta \, d\theta \, d\phi} \tag{1.10}$$

Once the directivity D_{max} has been calculated from the relative pattern, the directive gain in any other direction θ, ø can also be simply determined from the following relationship

$$D_{(\theta_1,\phi_1)} = D_{\max}\left[\frac{E(\theta_1,\phi_1)}{E_{\max}}\right]^2 \tag{1.11}$$

1.2.4.2 Gain in Decibels

Antenna gain is a power ratio. The gain of practical antennas may be range from zero to as much as 10,000 or more. As with any power ratio, antenna gain may be expressed in decibels. To illustrate in terms of the antenna power gain G, the value in decibels will be donated by G (dB) and is given by G (dB)=10 $\log_{10}$ G. The directive gain in decibels is calculated from the same formula, with D substituted for G.

1.2.4.3 Practical Significance of Power Gain

It is apparent for a given amount of input power in antenna; the power density at a given point in space is proportional to the power gain of the antenna in that direction. Therefore increasing the power gain of the transmitting antenna, without increasing the transmitting power can increase the signal available to a receiving antenna at that location. A transmitter with a power output of 1000 watts and antenna with a power gain of 10 (10dB) will provide the same power density at a receiving point as will a transmitter of 500 watts power and an antenna power gain of 20 (13dB). Obviously this relationship has great economic significance. Sometimes it may be much less expensive to double the gain of the antenna (add 3dB) than it would be to double the transmitter power (though in other cases the converse may be true). But generally speaking it is desirable to use as much antenna gain as may feasibly be obtained, when it is desired to provide the maximum possible field strength in a particular direction.

1.2.5 Beamwidth

When the radiated power of an antenna is concentrated into a single major lobe as seen in the pattern of Fig. 1-2, the angular width of this lobe is the beamwidth. The term is applicable only to antennas whose patterns are of this general type. Some antennas have a pattern consisting of many lobes, all of them more or less comparable in their maximum power density, or gain, and not necessarily all of the same angular width. But large classes of antennas do have patterns to which the beamwidth parameter may be appropriately applied.

1.2.5.1 Definition of Beamwidth

It is logical to define the width of a beam in such a way that it indicates the angular range within which radiation of useful strength is obtained, or over which good reception may be expected. From this point of view the convention has been adopted of measuring bean-width between the points on the beam pattern at which the power density is half the value at the maximum. In a plot of the electric intensity pattern, the corresponding points are those at which the intensity is equal to 0.707 of the maximum value. The angular width of the beam between these points is called the half-power beamwidth. When a beam pattern is plotted with the ordinate scale in the minus 3dB points. For this reason the half power beamwidth is often referred to as the -3dB beamwidth. Figure 2-4 illustrates the procedure of determining the -3dB beamwidth on a rectangular pattern plot.

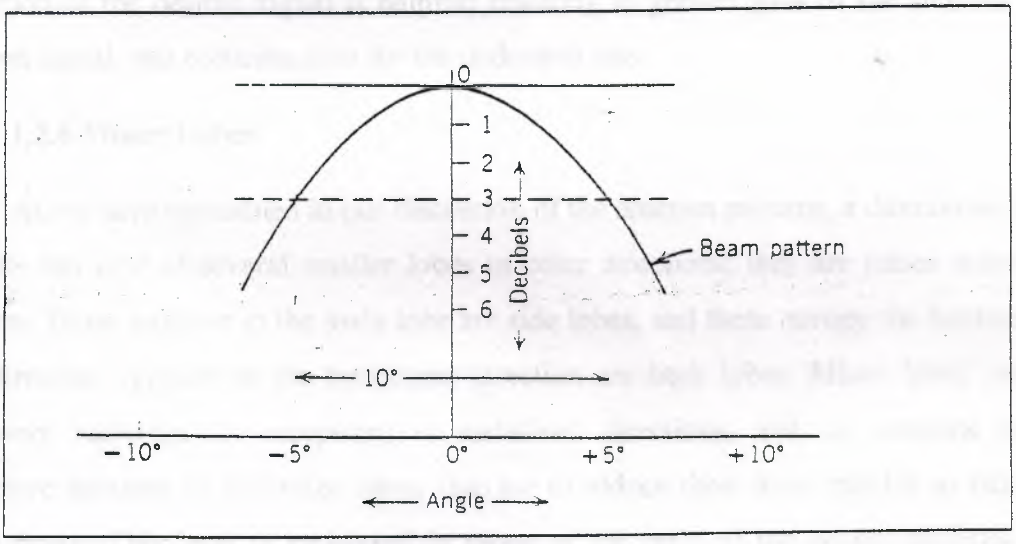

Figure 1.4 Determination of half-power (3dB-down) beamwidth.

This criterion of beamwidth, although adequate and convenient in many situations, it does not always provide a sufficient description of the beam characteristics. When beams have different shapes. An additional description may be given by measuring the width of the beam at several points, As an example, at -3dB, -10dB, and at the nulls (if they are present).

Some beams may have an asymmetric shape. Special methods of describing such beams can be employed. In the final analysis the best description of a beam is a plot of its pattern.

1.2.5.2 Practical Significance of Beamwidth

If an antenna has a narrow beam and is used for reception, it can be used to determine the direction from which the received signal is arriving, and consequently it provides information on the direction of the transmitter. To be useful for this purpose, the antenna beam must be steerable; that is, capable of being pointed in various directions. It is intuitively apparent that for this direction-finding application, a narrow beam is desirable and the accuracy of direction determination will be inversely proportional to the beamwidth. In some applications receiving may be unable to discriminate completely against an unwanted signal that is either at the same frequency as the desired signal or on nearly the same frequency. In such a case, pointing a narrow receiving antenna beam in the direction of the desired signal is helpful; resulting in greater gain of the antenna for the desired signal, and reducing gain for the undesired one.

1.2.6 Minor Lobes

As we have mentioned in our discussion of the antenna patterns, a directional antenna usually has lobe of several smaller lobes in other directions; they are minor lobes of the pattern. Those adjacent to the main lobe are side lobes, and these occupy the hemisphere in the direction opposite to the mainbeam direction are back lobes. Minor lobes ordinarily represent radiation (or reception) in undesired directions, and the antenna designer therefore attempts to minimize them, that are to reduce their level relative to that of the main beam. This level is expressed in terms of the ratio of the power densities in the mainbeam maximum and in the strongest minor lobe, and often expressed in decibels.

Since the side lobes are usually the largest of the minor lobes, this ratio is often called the side-lobe ratio or side-lobe level. A typical side-lobe level, for an antenna in which some attempt has been made to reduce the side-lobe level, is 20dB, which means that the power density in the strongest side lobe is 1 % of the power density in the main beam. Side-Lobe levels of practical well-designed directional antennas typically range from about 13dB (power-density ratio 20) to about 40dB (power density ratio 10,000). Attainment of a side-lobe level better than 30dB requires very careful design and construction. Figure 1-5 shows a typical antenna pattern with a main beam and minor lobes, plotted on a decibel scale to facilitate determination of the side-lobe level, which is here seen to be 25dB.

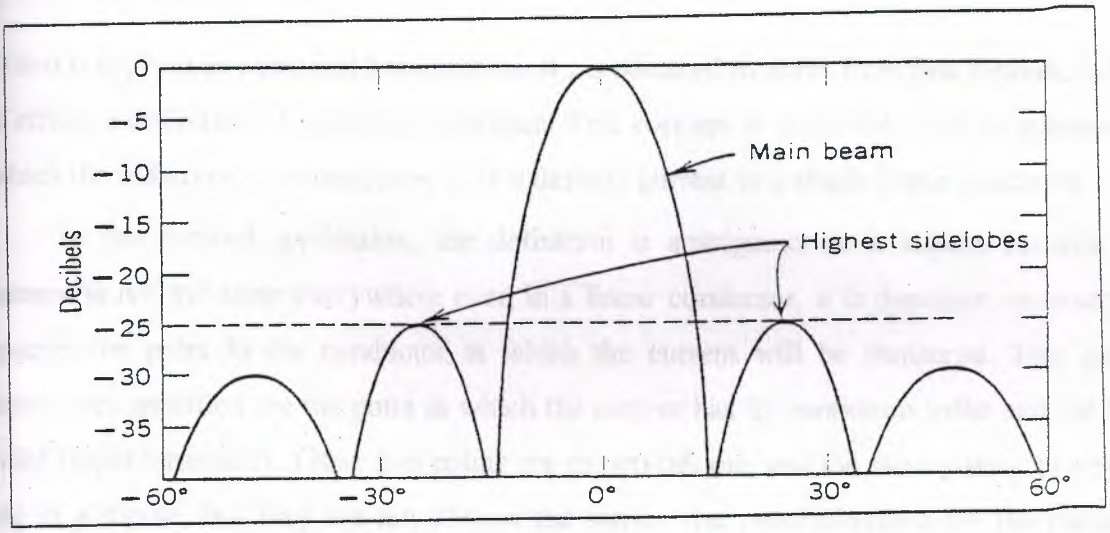

Figure 1.5. Decibel pattern plot indicated side lobe level.

In some applications side lobes are not especially harmful unless their level becomes comparable to the main-beam level. In other applications it may be important to hold the side-level to an absolute minimum. In most radar systems, a low side-lobe level is important. If the radar is very sensitive, a large target located in the direction of one of the antenna side lobes (or even a back lobe) may appear on indicator oscilloscope as though it were a target in the main beam.

1.2.7 Radiation Resistance and Efficiency

In a large class of antennas the radiation is associated with a flow of RF current in a conductor or conductors. As is well known in elementary electric circuit theory, when a current I flows in a resistance R, an amount of power $P = RI^2$ will be dissipated, that is, electrical energy will be converted into heat at this rate. In an antenna, even if there is no resistance in the conductors, the electrical energy supplied by the transmitter is lost just as though it had been converted in to heat a resistance, although in fact it is radiated. It is customary to associate this loss of power, through radiation, with a fictitious radiation resistance that bears the same relationship to the current and the radiation power as an actual resistance bears to the current and dissipated power. If the power radiated by the antenna is P and the antenna current is I, the radiation resistance is defined as

$$R_r = \frac{P}{I^2} \tag{1.12}$$

When P is given in watts and I in amperes, R_r is obtained in ohms from this formula, which is effect, a definition of radiation resistance. This concept is applicable only to antennas in which the radiation is an associated with a definite current in a single linear conductor.

In this limited application, the definition is ambiguous as it stands, because the current is not the same everywhere even in a linear conductor, it is therefore necessary to specify the point in the conductor at which the current will be measured. Two points sometimes specified are the point at which the current has its maximum value and the feed point (input terminals). These two points are sometimes one and the same points, as center-fed in a dipole, but they are not always the same. The value obtained for the radiation resistance of the antenna depends on which point is specified; this value of the radiation resistance referred to that point. The current maximum of a standing-wave pattern is known as a current loop, so the radiation resistance referred to the current maximum is sometimes called the loop radiation resistance.

The word maximum here refers to the effect current rms in that part of the antenna where it has its greatest value. It does not mean the peak value of the current at this point during the RF cycle, when Eq.1-12 is used as the definition. In some texts, however, formulas for radiation resistance are written in terms of this peak value, which is the amplitude of the current sine wave. Equation 1-12 will yield a value of radiation resistance only half as great as the true value -if the current amplitude is used for I, the correct formula in terms of the current amplitude I_0, is $R_r=2P/I^2$, note that $I_0 = \sqrt{2I_{rms}}$.

The radiation resistance of some types of antennas can be calculated, when there is clearly defined current value to which it can be referred, but for other types the calculation cannot be made practically, and the value must be obtained by measurement. Methods of making such a measurement will be described later.

The typical values of the loop radiation resistance of actual antennas range from a fraction of an ohm to several hundred ohms. The very low values are undesirable because they imply large antenna current, and therefore the possibility of considerable ohmic loss of power, that is, dissipation of power as heat rather than as radiation. An excessively high value of radiation resistance would also be undesirable because it would require a very high voltage to be applied to the antenna. Very high voltage values do not occur in practical antennas, because there is always some ohmics resistance whereas very low values sometimes do occur unavoidably.

Antennas always do have some ohmic resistance, although sometimes it may be so small as to be negligible. The ohmic resistance is usually distributed over the antenna, and since the antenna current varies, the resulting loss may be quite complicated to calculate. In general, however, the actual loss can be considered to be equivalent to the loss in a fictitious lumped resistance placed in series with the radiation resistance. If Ro denotes this equivalent ohmic loss resistance, the full power (dissipated plus radiated) is $I^2=(R_0+R_r)$, whereas the radiation power is $I^2 R_r$. Hence the antenna radiation efficiency ξ_r is given by

$$\xi_r = \frac{R_r}{R_0 + R_r} \tag{1.13}$$

It must be acknowledged that this definition of efficiency is not really very useful even though it may occasionally be convenient. The fact is both Ro and R_o is fictitious quantities, derived from measurements of current and power; R_r is given in these terms by Eq.1-12, and R_o is correspondingly equal to P_o / I^2. Making these substitutions into Eq.1-13, then it gives the more basic definition of the efficiency:

$$\xi_r = \frac{P_r}{P_0 + P_r} \tag{1.14}$$

1.2.8 Input Impedance

An antenna whose radiation results directly from the flow of RF current in a wire or other linear conductor must somehow have this current introduced into it from a source of RF power transmitters. The current is usually carried to the antenna through a transmission line. To connect the line to the antenna, a small gap is made in the antenna conductor, and the two wires of the transmission line are connected to the terminals of the gap at antenna input terminals. At this point of connection the antenna presents load impedance to the transmission line. This impedance is also the input impedance of the antenna and it is equal to the characteristic of the line Z_o, the input impedance of the antenna is one of it is important parameters. Measurement of the antenna input impedance would be discussed later. The input impedance determines how large a voltage must be applied at the antenna input terminals to obtain the desired current flow and hence the desired amount of radiated power. Thus, the impedance is equal to the ratio of the input voltage E_i to the input current I_i and it can be written as

$$Z = \frac{E_i}{I_i} \tag{1.15}$$

Which is in general complex. If the gap in the antenna conductor (feed point) is at a current maximum, and if there is no reactive component to the input impedance, it will be equal to the sum of the radiation resistance and the loss resistance; that is

$$Z_i = R_i = R_r + R_0 \tag{1.16}$$

If this reactance has a large value, the antenna-input voltage must be very large to produce an appreciable input current. If in addition the radiation resistance is very small, the input current must be very large to produce appreciable radiated power. Obviously this combination of circumstances, which occurs with the short dipole antenna that must be used at very low frequencies, results in a very difficult feed problem or impedance-matching problem, they are usually fed by waveguides rather than by transmission line. The equivalent of input impedance can be defined at the point of connection of the waveguide to the antenna, just as waveguides have characteristic wave impedance analogous to the characteristic impedance of a transmission line. For some types of antennas consisting of current-carrying conductors this is difficult, and it may even be difficult to define input impedance. This is true, as an example, for an array of dipoles, when each dipole is fed separately; sometimes each dipole, or groups of dipole, will be connected to separate transmitting amplifiers and receiving amplifiers. The input impedance of each dipole or group may then be defined, but the concept becomes meaningless for the antenna as a whole, as does also for simple linear-current radiation elements; but they comprise a very large class of antennas.

1.2.9 Bandwidth

All antennas are limited in the range of frequency over which they will operate satisfactorily. This range is called the bandwidth of the antenna. Bandwidth is a concept that is probably familiar in other applications, sometimes by another name. For example, a television I-f amplifier must have a bandwidth of approximately 4MHz in order to pass all the frequency components of a television signal. A television-transmitting antenna must have sufficient bandwidth to receive all the channels to which the receiving set can be tuned. If an antenna were capable of operating satisfactory from a minimum frequency of 155MHz to a maximum frequency of 205MHz, its bandwidth would be 10MHz. It would also be said to have a 5% bandwidth (the actual bandwidth divided by the center frequency of band, times 100). Some antennas are required to operate only at a fixed frequency with a signal that is narrow in its bandwidth; consequently there is no bandwidth problem in designing such an antenna. In other applications much greater bandwidths may be required; in such cases special techniques are needed. Some recent developments in broadband antennas permit bandwidths so great as they are described by giving the numerical ratio of the highest to the lowest operating frequency, rather than as a percentage of the center frequency. In these terms, bandwidths of 20 to 1 are readily achieved with these antennas, and ratios as great as 100 to 1 are possible.

1.2.10 Beam Area or Beam Solid Angle

An arc of a circle seen from the center of this circle subtends an angle θ. Thus, referring to Fig.2-6a, the arc length θR subtends the angle θ. The total angle in the circle is 2π rad so the total arc length is $2\pi R$. By using the same concept, an area A of a sphere surface seen from the center of the sphere subtends a solid angle Ω as shown in Fig. 1-6b

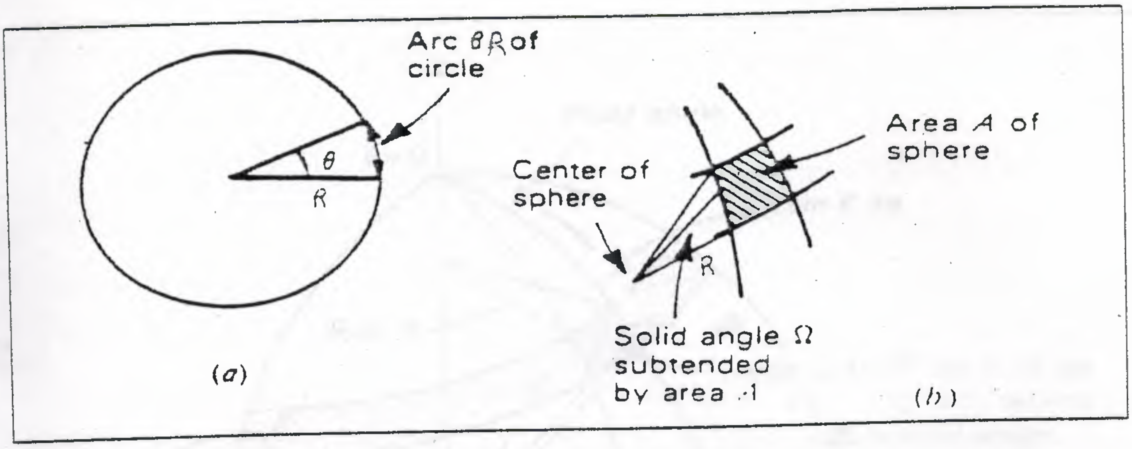

Figure 1.6 (a) Arc length Rθ of circle has radius R subtends the angle θ. The area A of a sphere of radius R subtends a solid angle Ω.

The total solid angle subtended by the sphere is 4 π steradians (or square radians), abbreviated sr.

By using Fig. 1-7 we can discuss the solid angle in more details. From Fig. 1-7, it is shown that the solid angel dΩ subtended by dA is

$$d\Omega = \sin\theta \; d\theta \; d\phi \tag{1.17}$$

To more declaration, the incremental area dA of the surface of a sphere is given by $dA = (R\sin\theta \; d\phi)\,(R\; d\theta) = R^2 \sin\theta \; d\phi \; d\theta = R^2 \; d\Omega$ the area of the strip of width R dθ extending around the sphere at a constant angle θ is given by dAs= (2 π R sinθ)(R dθ). Integrating this for θ values from 0 to π yields the area of the sphere. Thus,

$$Area\ of\ sphere = 2\pi R^2 \int_0^{\pi} \sin\theta \; d\theta = 4\pi R^2 \tag{1.18}$$

By comparing this result with dA = (R Sin θ dθ(R dθ) = R^2 sin θ dθ = R^2dΩ we fined that dΩ for the whole sphere is 4 π

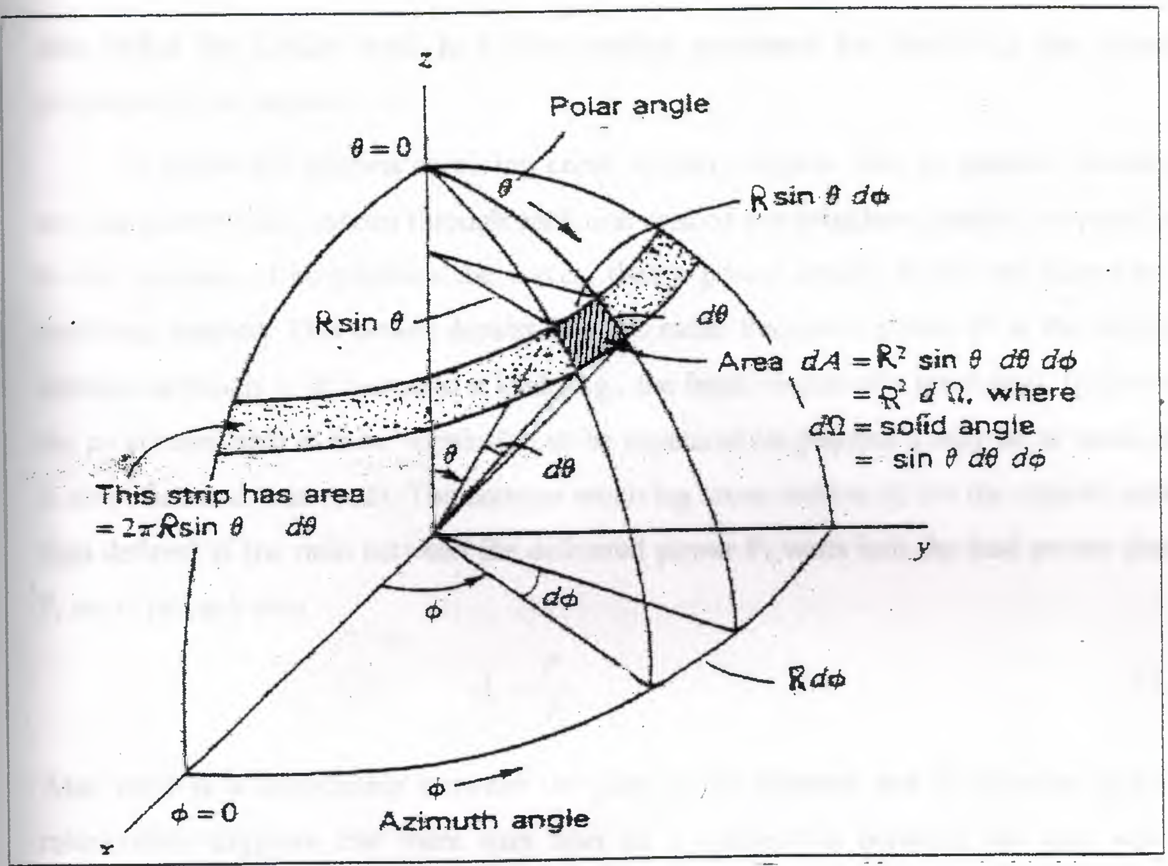

Figure 1-7 Spherical coordinates in relation to the area dA solid angle $d\Omega = \sin\theta \; d\theta \; d\phi$

Now the beam area (or beam solid angle) Ω_A for an antenna is given by the integral of the normalized power pattern over a sphere (4 π, sr)

$$\Omega_A = \int_0^{2\pi}\int_0^{\pi} P_n(\theta,\phi) \, d\Omega \tag{1.19}$$

1.2.11 Capture Area or Receiving Cross Section

Although there is a reciprocal relationship between the transmitting and the receiving properties of antennas, it is sometimes more convenient to describe the receiving properties in a somewhat different way. Whereas the power gain is the natural parameter to use for describing the increases power density of the transmitted signal due to the directional properties of the antenna, a related quantity called the receiving cross section, sometimes also called the capture area, is a more natural parameter for describing the reception properties of the antenna.

To define the antenna receiving cross section, suppose that an antenna radiates an amount power which passes through each unit area of any imaginary surface perpendicular to the direction of propagation the waves, then a power density P_i will be passed to the receiving antenna. This power density induces radio frequency power P_r at the receiving antenna terminals is delivered to a load (e.g., the input circuit of a receiving). In principle the power available at these terminals can be measured (in practice it may be so small, so it is amplified and then read). The antenna receiving cross section A_r (or the capture area) is then defined as the ratio between the delivered power P_r watts into the load power density P_t watts per unit area

$$A_r = \frac{P_r}{P_i} \tag{1.20}$$

Also there is a relationship between the gain of the antenna and its physical size, this relationship suggests that there may also be a connection between the gain and the receiving cross section area and this indeed turns out to be true. The receiving cross section area in isotropic A_{ro} is given as

$$A_{r0} = \frac{\lambda^2}{4\pi} \Rightarrow A_r = \frac{G\lambda^2}{4\pi} \tag{1.21}$$

where $G = \xi D$, λ is the wavelength, note that λ has relationship with the size, then A_r , G and the size. Equation 1-20 may be proved theoretically and verified experimentally. From this relationship it follows that

$$D = \left(\frac{4\pi\, A_r}{\xi\, \lambda^2}\right) \tag{1.22}$$

where D is the directive gain. It is clear from this relationship that the gain increases when A_r increases, and λ and ξ decrease, and vice versa. Thus, the power is

$$P_r = \xi\left(\frac{P_i D \lambda^2}{4\pi}\right) \tag{1.23}$$

Therefore the concept of the receiving cross section of an antenna is not a necessary one. It is possible to calculate the received-signal power without using equation 1-23.

CHAPTER TWO

CIRCULARLY POLARIZED ANTENNAS

2.1 Panel Types Antennas

In many cases, the supporting structure is a triangular or square tower. Panel antennas are primarily used to control or minimize the reflections from the supporting structure. Some panel antennas are made of a single horizontal dipole or two crossed dipoles (circularly polarized panel) in front of a reflector.

The reflector can be a flat panel, a comer reflector, or a pillbox (commonly referred to as *cavity-backed)*. The reflector is usually a wire grid for VHF or a solid sheet for UHF.

In order to obtain an omni directional radiation pattern, three- or four-panel antennas are placed around a triangular or square tower, respectively. In general, panel antennas with 6-dB beam width of 90 and 120 ° are used for arrays around square and triangular towers, respectively. When several panels are arranged around a cylindrical structure in a single layer, as shown in Fig. 2.1,

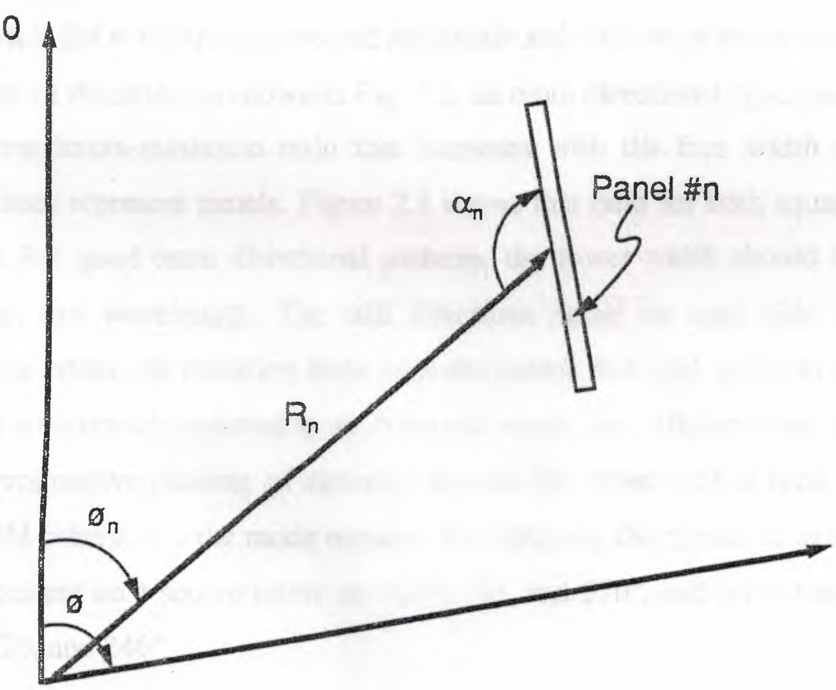

Figure 2.1 Geometry for equation (2.1)

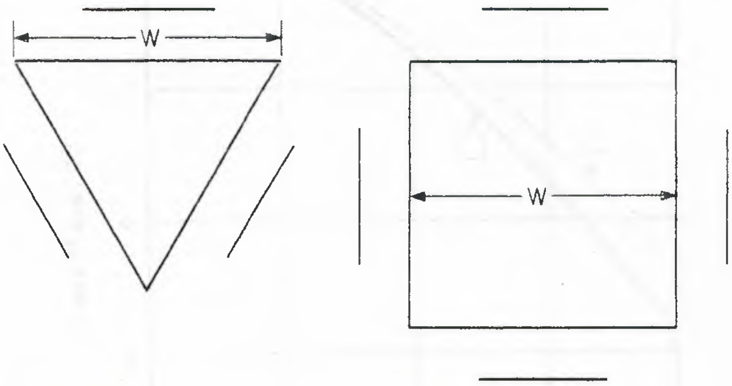

Figure 2.2 Panel-type antennas for triangular and square

The combined pattern $E(\Phi)$ may be calculated by using the following expression:

$$E(\Phi) = \sum_{n=1}^{N} I_n M_n(\psi) \exp \mathrm{i}\left[\xi_n(\psi) + \sigma_n + kR_n \cos(\Phi - \Phi_n)\right] \qquad \textbf{(2-1)}$$

Where $I_n e^{i\sigma_n}$ = excitation current of the *nth* panel

$M_n(\psi) e^{i\xi_n(\psi)}$ = pattern of the *nth* panel

Φ_n = polar angle of the *nth* panel

R_n = length of the radial to the *nth* panel

$\Psi = \pi - \alpha_n \Phi - \Phi_n \alpha_n$ = tilt angle of the *nth* panel

With the panels fed with equal phase and amplitude and with the antenna elements placed in the center of the sides, as shown in Fig. 2.2, an omni directional type pattern is obtained with a maximum-minimum ratio that increases with the face width of the tower. The short lines represent panels. Figure 2.3 shows this ratio for both square and triangular towers. For good omni directional patterns, the tower width should not be much greater than one wavelength. The null directions occur on each side of the crossover directions where the radiation from adjacent panels does not arrive in phase. This arrangement is commonly referred to as *azimuthal mode zero.* Higher-order modes are obtained by progressive phasing of elements around the tower with a total phase progression *of360M,* where *M* is the mode number. For instance, the phases of panels in the mode 1 arrangement on a square tower are 0,90, 180, and *270°,* and on a triangular tower they are 0,120, and 240°.

NEAR EAST UNIVERSITY

Faculty of Engineering

Department of Electrical and Electronic Engineering

TV AND FM TRANSMITTING ANTENNAS

Graduation Project

EE - 400

Student: ERGÜN ÖZDEMİR (961187)

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ABSTRACT

We had thought to do our work on the antenna, and then we search for the important parts on this subject since the antenna is one of the most common and important parts in the communication system.

The term antenna is defined by the dictionary as a usually metallic device (as a rod or wire) for radiating or receiving radio waves. The official definition of the Institute of Electrical and Electronics Engineers (IEEE) is simply as a means for radiating or receiving radio waves. The ideal antenna is, in most application, one that will radiate all the power delivered to it by a transmitter in the desired direction or directions and with the desired polarization.

The objective in this chapter is to provide the reader with specifications and descriptions of television and FM radiobroadcast antennas. The various antennas to be described also may be used for other applications in the frequency range of 10 MHz to 10 GHz. Broadcast antennas have frequency, pattern, power capacity, impedance, and environmental requirements which are imposed by regulatory agencies such as the Federal Communications Commission (FCC) or by system specifications. For instance, frequency and pattern are regulated by the FCC, while impedance, power capacity, and environmental requirements are system-related.

TABLE OF CONTENTS

INTRODUCTION

The first antennas were employed by Hertz in 1887 in his classic demonstrations of the electromagnetic waves that had been predicted earlier by Maxwell. Hertz's receiving antenna was a circular loop of wire broken by a microscopic gap. The radius of the loop was 35 cm, a dimension that had been found by experiment to put the loop into resonance with the transmitter. Hertz later placed his rod antenna in the focal plane of a cylindrical mirror. About 8 years after the early work of Hertz, a Professor Popoff of Kronstadt was engaged in a study atmospheric electricity, and in connection with this study, he placed a receiving antenna, consisting of a metallic rod, above this housetop. The receiver ("coherer") was connected between this rod and the earth.

In 1897, Marconi described a complete system for wireless telegraphy. In this system, one terminal of the spark transmitter was connected to an elevated wire and the other terminal was connected to the earth. Apparently, Marconi was the first to realize the importance of elevating the transmitting antenna. The transmitting antenna for the first wireless transatlantic communication (from Poldhu in Cornwall to Newfoundland, in 1901) was a vertical fanlike structure consisting of 50 vertical copper wires supported by a horizontal wire. The horizontal wire was stretched between two masts about 150 ft high and 200 ft apart. The receiving antenna in Newfoundland was supported by kites. By 1907, commercial telegraph services had been established, and the advantage of top-loaded antennas was widely recognized. The frequency of operation of these early communications systems was usually in the range from 50 to 100 kHz, and consequently the antennas were small compared to the wavelength. Unlike the transmitters of today, the antennas that were employed in the early systems usually strongly influenced the operating frequency.

One of the earliest papers on the subject was that of Abraham, who, in extending some earlier work on spheres by J.J. Thomson, studied the natural oscillations of a conducting prolate spheroid. He determined the natural frequencies and calculated the fields of a half-wave dipole. Hertz himself had studied the fields of point dipoles. His work was carried on further by Sommerfeld and others, and by 1914, Hertz potentials and vector potentials had been employed extensively in calculations of the radiation patterns of known current. Further, Poynting's theorem had been employed to calculate the total power radiated from antennas together with their radiation resistances. Interest in resonant length antennas (half-wavelength dipoles or quarter-wavelength monopoles above ground) began to grow about 1920, after the discovery that the De Forest triode tube could be made to produce continuous wave oscillations at the higher frequencies (hundreds or even thousands of kilohertz). A these higher frequencies, it become practical to construct resonant length antennas or even arrays of these. By about 1930, the theory and practice of simple linear arrays had been developed and applied to broadcast transmitters for interference control.

As antennas that were of the order of a wavelength came into use, the need for a better understanding of the interaction of the antennas with transmission lines and transmitters grew more pressing. Outstanding contributions on this subject were made by J. R. Carson, who presented a generalization of the reciprocity theorem, and by P. S. Carter, who published antenna terminal-impedance definitions and calculations. Later in the thirties the treatment of the antenna as an electromagnetic boundary-value problem was revived. King and Hallen formulated the linear-antenna problem as an integral equation. Stratton and Chu employed a prolate spheroidal model for the linear antenna and deduced some of its properties by means of spheroidal wave functions. Perhaps the most illuminating model was introduced by Schelkunoff. According to his initial model, the straight-wire antenna was regarded as a limiting case of a biconical horn antenna. With such a model, the solutions may be expressed in spherical coordinates.

In the mid-thirties, a new branch of antenna technology began to develop. The developments went hand in hand with the development of generators in the microwave frequency range and the use of metallic pipes as waveguides. These waveguides were flared out into horns, a rather natural step by analogy with the corresponding acoustic problem. Later, radiating slots were introduced into the walls of the waveguides. Still later, the World War II requirements for special high-gain antennas led to the development of large parabolic reflectors and lenses. Schelkunoff also published a beautiful generalization of linear array theory.

The commonly used "whip" antennas on cars "rabbit ears" on TV receivers single turn loop antennas for UHF TV reception roof mounted log-periodic TV antennas and satellite paraboloidal reflector receiving antennas are so prevalent that most people are clearly aware of need for antennas in the support of our daily communication needs. These commonly occurring antennas represent only a small segment of the antenna systems that have been developed. For specialized and high performance communication links, radar systems, navigational systems, and scientific studies highly complex antenna systems are needed.

The objective in this project is to provide the reader with specifications and descriptions of TV and FM radiobroadcast antennas. The various antennas to be described also may be used for other applications in the frequency range of 10 MHz to 10 GHz. Broadcast antennas have frequency, pattern, power capacity, impedance, and environmental requirements which are imposed by regulatory agencies such as the Federal Communications Commission (FCC) or by system specifications. For instance, frequency and pattern are regulated by the FCC, while impedance, power capacity, and environmental requirements are system-related.

Broadcast frequencies in the United States are allocated and regulated by the FCC. The following frequency bands are assigned to television broadcasting:

Table 1. Frequency bands in TV broadcasting.

Low VHF	Channels 2-4	54-72 MHz
	Channels 5-6	76-88 MHz
High VHF	Channels 7-13	174-216 MHz
UHF	Channels 14-83	470-890 MHz

Each channel is assigned 6 MHz of bandwidth, with visual carrier and color subcarrier at 1.25 and 4.83 MHz above the lower edge of the channel, respectively, and with aural carrier at 0.25 MHz below the upper edge of the channel. The power levels of the visual subcarrier and aural carrier usually are within 20 percent of the visual carrier.

FM radio frequencies are limited to the band between 88 and 108 MHz. There are 100 channels, each with a 200-kHz bandwidth. Pattern requirements are functions of coverage goals, site location, local terrain, and the available options on mounting structures. Coverage goals are regulated and limited by FCC specifications as spelled out in the *Code of Federal Regulations (CFR* 47). For instance, a TV station's coverage is specified by the distance to the "city grade," "grade A," and "grade B" contours. Table 28-1 shows the minimum levels of the field strength (present at 50 percent of locations 50 percent of the time) assigned to these contours. The transmitter location and power and the antenna height and gain are chosen such that the city-grade contour covers the entire principal community to be served.

For an FM station, the coverage limits are defined by two contours. These are the 70-dBu contour (3.16 mV/m), for city grade, and the 60-dBu contour (1.0 mV/m). These limits, along with the path-loss curves (also known as the 50-50 curves), which are docu mented in *CFR* 47, are used to determine the ERP* and/or the gain of the antenna for a given antenna height and location. The FCC manual specifies the maximum ERP for TV and FM stations. The maximum power[2] varies with regions or zones of the United States and with the antenna height above the average terrain (HAAT). Towers that support broadcast antennas are either self-supported or guyed, and their heights range from 100 to 2000 ft.
Since broadcast antenna structures usually consist of vertical arrays of radiating elements, antenna 'directivity is approximately equal to the product of azimuth-pattern directivity and the elevation directivity (the number of bays in some antennas). The antenna gain is always referenced to the gain of a half-wave dipole (2.15 dB above the isotropic element) and is equal to the directivity less the losses, such as impedance loss and/or polarization-mismatch loss.The majority of applications call for omnidirectional azimuth patterns. The circularity of the pattern depends on the type of antenna when top-mounted and also on the

Table 2. Grade contours of TV broadcast stations.

Channel	City grade (dbu)	Grade A (dbu)	Grade B (dbu)
2-6	74	68	47
7-13	77	71	56
14-69	80	74	64

configuration of the support structure when side-mounted. Other requirements call for various types of azimuth patterns, such as cardioid, skull-shaped, peanut-shaped, etc., to protect other stations or reduce radiation into low-population areas. An azimuth pattern with a circularity of ±2 dB is considered omnidirectional. The maximum to minimum ratio of directional patterns should not exceed 15 dB for FM and UHF TV (with the ERP greater than 1 kW). For Channels 2 to 13 this ratio is 10 dB.

TV and FM broadcasting was originally limited to horizontal polarization. In the 1960s, the FCC allowed circular polarization (CP) for FM broadcasting. This provided improved reception, especially for vehicles with whip antennas, which are predominantly vertically polarized. In 1977, the FCC permitted TV broadcasting in right-hand CP as well. In going from horizontal polarization to CP, the stations were allowed to maintain their maximum ERP in horizontal polarization so as to maintain the field strength existing before the conversion. By allowing the same ERP for vertical polarization, the FCC actually allowed doubling the radiated power. This has provided improved reception for receivers with indoor antennas such as monopoles and rabbit ears.

In some instances, use of CP has reduced ghosting because reflections from buildings and other objects tend to have the opposite sense of CP. The acceptable axial ratio for CP antennas is 3 dB or less. The receiving antennas are almost all linearly polarized, and because of this and the unfriendly propagation path of most broadcast environments, the importance of axial ratio is somewhat superfluous. For instance, the majority of FM antennas are omni directional CP antennas side mounted on towers without regard to the effect of the tower on the axial ratio of the radiated CP wave. FCC regulations are limited to the shape and directivity of the vertical and horizontal polarization patterns and do not specify the axial ratio of the radiated CP wave

Chapter one is primary concerned with definitions and releated terminology. There is an explanation of antenna parameters and structure with some equations and the figures. Chapter two gives an information about the types of broadcasting antennas such as circularly polarized antennas. Chapter three presents an explanation of horizontally polarized antennas and FM antennas. Finally, in chapter four we give a conclusion of the project.

CHAPTER ONE

ANTENNA FUNDAMENTALS AND STRUCTURES

1.1Antenna Structure

The structure of the antennas depends upon the type and the destination, but in general, all antennas have the following structure.

1.1.1 Size

The size of antenna range from microminiature to gigantic, and it depends on the wavelength, which has proportionality with the operations frequency, and this relationship is simple and fast.

The large antennas are used for low frequencies (high wavelength), and vice versa, small antennas are used for high frequencies (low wavelength), but sometimes-large antennas are used at short wavelength (high frequencies) to obtain a highly directional radiation pattern and high gain in a preferred direction.

In practice field, the increasing of the size is limited, because at determining size, there is no point in increasing this size because it produce a little or no additional gain and the required precision of construction or maintenance of phase relationship is not attainable. Moreover, very small antennas can be used at long wavelength, when efficiency is not important. In general, the largest antennas are used at the VLF, especially for transmitting, where radiation efficiency is important. As an example of the extremely large VLF antenna is Navy's installation that has tower 1000 feet high, extends over an area of 2 square miles. In contrast, a half wave dipole at the microwave frequencies may be considerably less than an inch long.

1.1.2 Supports

There must often be some supporting structure to place the radiating element or elements in a clear location (with often is synonymous with a high location). Such devices as towers, masts, and pedestals support antennas.

Towers are used when great height is required. Masts may be quite high, but they are often as short as a few feet. Pedestals are the base structures of antennas such as reflectors and lenses, for which height is not important as strength. Sometimes an antenna may be mounted directly on a vehicle, such as an automobile, ship, aircraft, or spacecraft, where no intermediate support is required. Moreover, towers and masts are sometimes themselves used as antennas rather than as supports. In the standard broadcast band (550-1600KHz). As an example, vertical towers of heights up to several hundred feet are used as transmitting antennas.

1.1.3 Feed Line

We can simply define the feed lines as the transmission lines. These lines are used to connect the transmitter or receiver to the antenna. The design of the feed lines and any necessary impedance matching or power-dividing devices associated with it is one of the most important problems in the calculation of antenna design. At the very lowest frequencies the earth (ground) is a part of the antenna electrical system. Therefore, one terminal of the antenna input is a rod driven into the ground or a wire leading to a system of buried conductors, especially if the earth is dry in the vicinity of the antenna. The other terminal is then usually the base of a tower or other vertically rising conductor. Towers used in this way are usually supported at the base by a heavy insulator or insulators (series feed), but occasionally they are directly grounded and fed by connecting the feed wire a short distance up from the ground (shunt feed).

At somewhat higher frequencies, up to (up to 30MHz), the antenna may be a horizontal wire strung between towers, or other supports (from which it is insulated). The feed line is then often a two-wire balanced line connected at the center of the antenna, either to the two terminals provided by a gap in the antenna wire (series feed), or to two points somewhat separated on the unbroken antenna *wire* (shunt feed). Sometimes the feed line is connected at the end of the horizontal span, or elsewhere of center, but center feed is preferred because it results in better balance of the currents in the feed wires. The spacing between the two-wire-line is range from less than an inch to 12 inches or more. The last method is used for high frequencies. But coaxial feed lines are commonly used for upper high frequencies UHF (up to 1 GHz), because the two-wire-line spacing becomes too great a fraction of the wavelength to prevent appreciable radiation and because waveguides below 1000MHz are quite large and expensive. Coaxial line diameters range from a fraction of an inch up to 9 inches or more. Above 1000MHz, waveguides are commonly use, with some use of mall-diameter coaxial lines in low-power noncritical applications.

We should mention that, when the antenna rotates on a pedestal, or has other motion with respect to its support, the feed line must contain flexing sections or rotating joints, this require is quite important on the antenna measurement operations, as we will see later.

1.1.4 Conductors

Metals are the usual conducting materials of antennas. Metals of high conductivity, such as copper and aluminum (and its alloys), are naturally preferred. Brass may be used for machined parts. Magnesium is sometimes used where ultralight weight is important, usually in an alloy and with a protective coating or treatment. The steel may be used, when the strength is of primary importance, either with or without a coating or plating of copper. The conductivity of unplated steel is adequate when it is used in the form of sheets or other large-surface-area forms (as for the surface of a paraboloidal reflector). Antenna ware is sometimes made with a steel core for strength and to minimize stretching and with a copper coating to increase the conductivity. Such wire is virtually as good a conductor as solid copper. Since the radio frequency RF currents are concentrated near the surfaces of conductors (skin effect). For this reason brass and other metals are sometimes silver plated when exceptionally high conductivity is required. For the same reason large-diameter conductors may be hollow tubes without loss of conductivity. At low radio frequencies the conductivity of large-diameter conductors may be increased, compared to a solid conductor, by interweaving strands of small-diameter insulated wires; the resulting conductor is called Litz wire. This technique is most effective below about 500KHz. At higher frequencies it is not effective because the currents tend to flow only in the outer strands.

Conductor size in antenna design is determined by many factors, principally the permissible ohmic losses and resultant heating effects in some cases, mechanical strength requirements, permissible weight, electrical inductance and capacitance effects, and corona considerations in high-voltage portions of transmitting antennas. Large-diameter conductors minimize the Corona, by avoidance of sharp or highly curved edges, and by using insulators with metal end caps bonded to the insulating material, so that small air gaps between wires and insulators do not exist. Corona can occur on metal supports of the antenna as well as on the antenna conductor itself, as a result of induced voltages.

1.1.5 Insulators

The conducting portions of an antenna not only carry RF currents but also have RF voltages between their different parts and between the conductors and ground. So that, to avoid the short circuiting these voltages, insulators must sometimes be used between the antenna and its supports, or between different parts of the antenna. The insulators are also used as spacer supports for two-wire and coaxial lines and to break up guy wires with masts and towers to prevent the resonant or near-resonant lengths. The maximum permissible uninterrupted length of guy wire sections is about 1/8 wavelength. Also, the insulators are used to support long heavy spans of wire, so that it must be high strength. Typical insulating materials for such insulators are glass and ceramics, other (low loss) materials such as polystyrene and other plastics are used where less strength is required. Very large and heavy insulators are necessary in high-power transmitting applications to prevent flashover. Coaxial lines and waveguides in high-power applications may be filled with an inert gas, or dry air, at a pressure of several atmospheres, to increase the voltage-breakdown.

1.1.6 Weather Protection

The antennas are ordinarily out doors, so that, it must withstand wind, ice, snow, lightning, and sometimes corrosive gases or salt-laden air. Protection against wind and ice loads is primarily a matter of mechanical strength and bracing. Guy wires are used with tall structures or towers, to prevent their overturning in high winds. In the heavy current networks, the ice is sometimes melted from the heating that is produced from the current. Sometimes an antenna is totally enclosed in a protective housing of low-loss insulating material, which is practically transparent to the electromagnetic radiation. Such housing is called radome. Radomes are commonly used on some types of aircraft antenna for aerodynamic reasons. The protection against lightning-induced currents, and static-charge buildup is necessary for some types of antennas such as broadcasting towers, or any structure that stands high above its surrounding, if the conducting path to ground is not heavy, and direct. Insulators may be protected by horn or ball gaps, and static may be drained by connecting high-ohmage resistors across insulators.

1.2Antenna Parameters

The most fundamental properties of antennas are the following

1.2.1 Polarization

The wave polarization refers to the instantaneous component direction on a surface perpendicular to the direction of energy propagation. In the communication system only sinusoidal varying fields are ordinary used. The radiation of an antenna may be linearly, elliptically, or circularly polarized. Polarization in one part of the total pattern may be different from polarization in anther. As an example, in the case of a directional antenna with a main beam and minor lobes, the polarization may be different in the minor lobes and in the main lobe, or may even vary in different parts of the main lobe.

The simplest antennas radiate (and receive) linearly polarized wave. They are usually oriented so that the polarization (direction of the electric vector) is either horizontal or vertical. But sometimes the choice is dictated by the necessity, at other times by preference based on technical advantages, and sometimes there is no basis for choice one is as good and as easily achieved as the other. For example at the very low frequencies it is practically difficult to radiate a horizontally polarized wave successfully because it will be virtually cancelled by radiation from the image of the antenna in the earth, also vertically polarized waves propagate much more successfully at these frequencies (e.g., below 1000KHz). Therefore vertical polarization is practically required at these frequencies.

At the frequencies of television broadcasting (54 to 890MHz) horizontal polarization has been adopted as standard. The standard frequency is very important to determine the type of polarization. Otherwise, we have to design an antenna such has both polarizations, thus greatly complicating design problem and increasing the received noise level.

At the microwave frequencies (above 1GHz) there is little basis for a choice of horizontal or vertical polarization. Also in specific applications there may be some possible advantages in one or the other. Of course in communication it is essential that the transmitting and receiving antennas have the same polarization.

Circular Polarization has advantages in some VHF, UHF, and microwave applications. As an example, in transmission of VHF and low-UHF signals through the ionosphere, rotation of polarization vector occurs, the amount of rotation being generally unpredictable. Therefore if a linear polarization is transmitted it is advantageous to have a circularly polarized receiving antenna which can receive either polarization, or vice versa. The maximum efficiency is realized if both antennas are circularly polarized.

From the above explanation. It is obvious that in communication circuits it is essential that transmission and receiving antennas have the same polarization. Also it is apparent that the polarization properties of any antenna are an important part of its technical description (parameter of its performance). Sometimes it may be desirable to provide polarization pattern of the antenna, that is, a description of the polarization radiated as a function of the direction angles of a spherical coordinate system, although such a complete picture of the polarization is not ordinarily.

1.2.2 Radiation Pattern

The radiation pattern of an antenna is one of its most fundamental properties, and many of its performance parameters pertain to various aspects of the pattern.

We should mention that antennas have a reciprocal relationship between the processes of radiation and reception, so, it is customary to speak of the antenna pattern as radiation pattern, and a reception pattern as well because it also describes the receiving properties of the antenna. The radiation pattern describes the relative strength of the radiated field in various directions from the antenna, at a fixed or a constant distance.

Because the antenna pattern is three dimensional, a three-dimensional coordinate system is required. So, either Cartesian (rectangular) coordinates (x, y, z) or spherical coordinates (R, θ, ø) is used. The spherical coordinate system is an appropriate coordinate system to describe the antenna pattern because the radiation pattern may be expressed in terms of the electric field intensity, (for example, at some fixed distance R from the antenna), at all points on the spherical surface at that distance. Spherical points on the surface are then defined by the direction angles θ and Φ. The pattern then becomes a function of only two independent variables, since R is a constant, and this fact greatly simplifies the matter.

Figure 1-1 illustrates the relationship between the Cartesian and spherical coordinates. The projection of this distance r onto the xy-plane is designated θ, Φ, this means that changing r causes changing on θ, Φ.

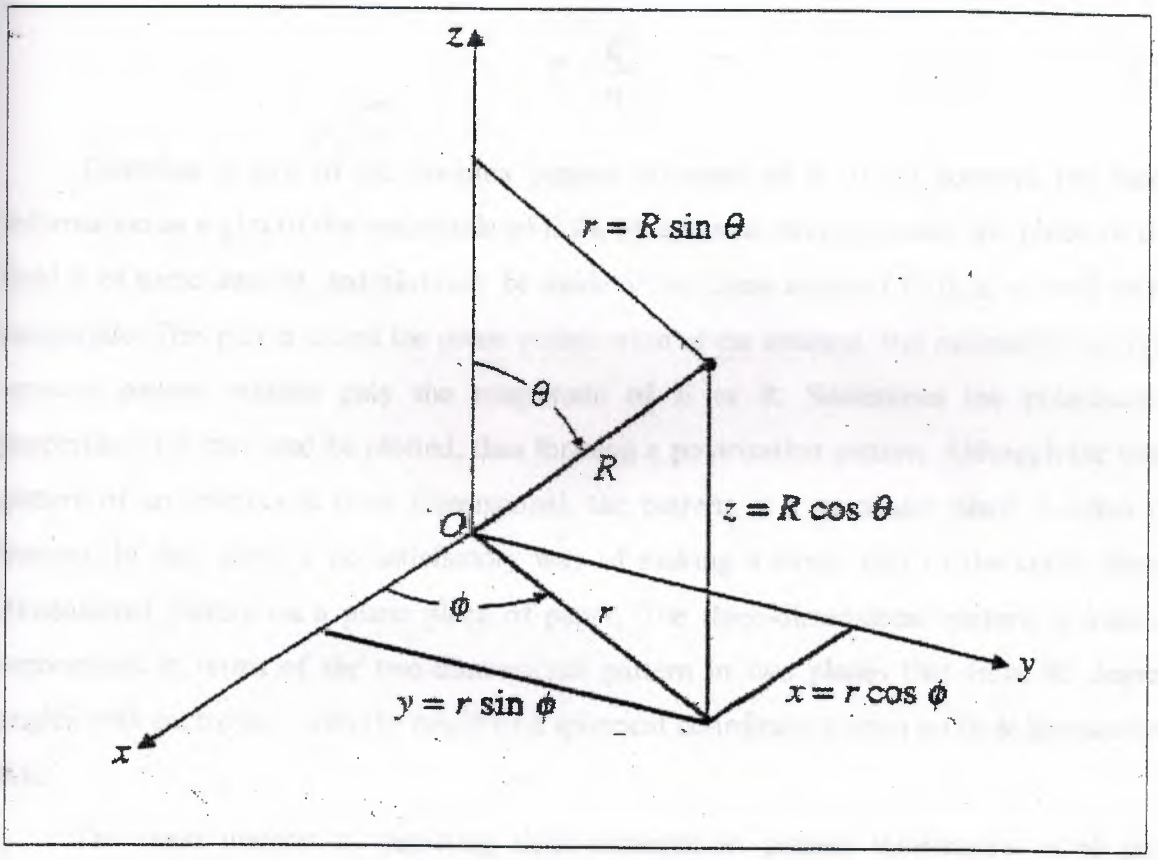


Figure 1.1 Showing interrelationship of space variables (x, y, z) and (R, θ, ϕ).

An antenna is supposed to be located at the center of a spherical coordinate system, its radiation pattern is determined by measuring the electric field intensity over the surface of a sphere at some fixed distance, R. Since the field E is then a function of the two variables θ and ø, so it is written E (θ, ø) in functional notation.

A measurement of the electric field intensity E (θ, ø) of an electromagnetic field in free space is equivalent to a measurement of the magnetic field intensity H (θ, ø), since the magnitudes of the two quantities are directly related by

$$E = \eta_0 H \tag{1.1}$$

(Of course, they are at right angles to each other and their phase angles are equal) where

$\eta_o = 377\Omega$ for air. Therefore the pattern could equally be given in terms of E or H.

The power density of the field, P (θ, ø), can also be computed when E (θ, ø) is known, the relation being

$$P = \frac{E^2}{\eta} \tag{1.2}$$

Therefore a plot of the antenna pattern in terms of P (θ, ø) conveys the same information as a plot of the magnitude of E (θ, ø). In some circumstances, the phase of the field is of some interest, and plot may be made of the phase angle of E (θ, ø) as well as its magnitude. This plot is called the phase polarization of the antenna. But ordinarily the term antenna pattern implies only the magnitude of E or P. Sometimes the polarization properties of E may also be plotted, thus forming a polarization pattern. Although the total pattern of an antenna is three dimensional, the pattern in a particular plane is often of interest. In fact, there is no satisfactory way of making a single plot of the entire three-dimensional pattern on a plane piece of paper. The three-dimensional pattern is usually represented in terms of the two-dimensional pattern in two planes that from 90 degree angles with each other, with the origin of a spherical coordinate system on their intersection line.

The main method of depicting three-dimensional pattern information is to plot contours of constant signal strength on the surface of a sphere containing the antenna at its center. But ordinarily only the principal plane patterns are given, as they convey an adequate picture of the three-dimensional pattern for most purposes.

Pattern in a plane involves only one angle, so that, it is represented by polar coordinates, it would be possible to use Cartesian coordinates. If this were done, the shape of the pattern would be unchanged; but because interpretation of the meaning of the pattern in terms of the Cartesian coordinates would be relatively difficult, this is never done. It is fairly common to plot the pattern on rectangular-coordinate graph paper but in terms of the direction angle as the abscissa and field strength or power density as the ordinate. This type of plot distorts the appearance of the pattern geometrically but preserves the interpretability of an angle representation and makes the plotting and the reading of the low amplitude portions of the pattern easier. Figures 1 -2a and 1 -2b compare these two representations.

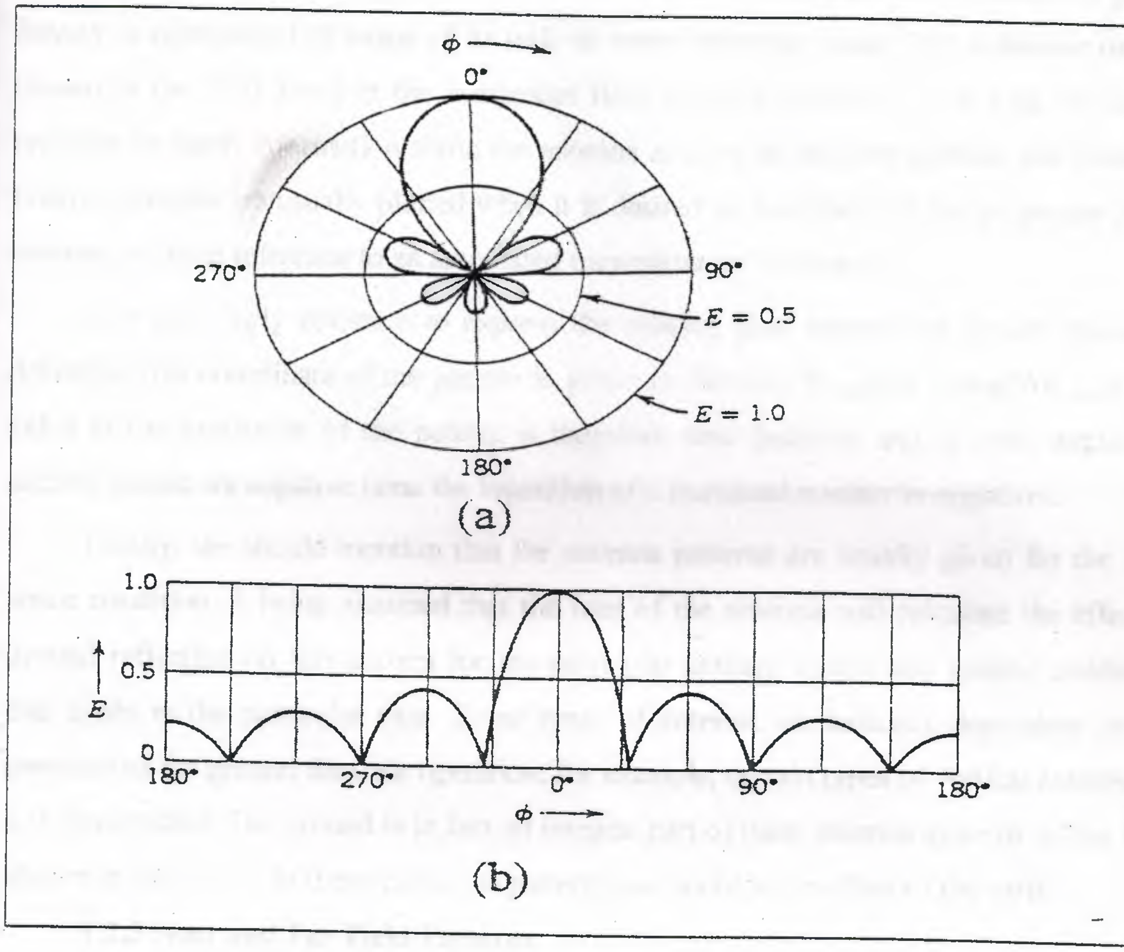


Figure 1.2 Comparison of plane pattern plotted in polar and rectangular form. The same pattern is represented in both cases and the coordinates are the same. Only the plot is different (a) polar (b) rectangular plot.

Note that it is easier to locate the angular positions of nulls (zeros) of the pattern on the rectangular plot.

If the radiation pattern is plotted in terms of the field strength in electrical units, such as volts per meter or the power density in watts per square meter, it is called an absolute pattern. An absolute pattern actually describes not only the characteristics of an antenna but also those of the associated transmitter, since the absolute field strength at a given point in space depends on the total amount of power radiated as well as on the directional properties of the antenna.

Often when the pattern is plotted in relative terms, that is, the field strength or power density is represented in terms of its ratio to some reference value. The reference usually chosen is the field level in the maximum field strength direction. This type of pattern provides as much information about the antenna as does an absolute pattern, and therefore relative patterns are usually plotted when it is desired to describe only the properties of the antenna, without reference to an associated transmitter (or receiver).

It is also fairly common to express the relative field strength or power density in decibels. This coordinate of the pattern is given as $20\log(E/E_{max})$ or $10\log(P/P_{max})$. The value at the maximum of the pattern is therefore zero decibels, and at other angles the decibel values are negative (sine the logarithm of a fractional number is negative).

Finally, we should mention that the antenna patterns are usually given for the free-space condition, it being assumed that the user of the antenna will calculate the effect of ground reflection on this pattern for the particular antenna height and ground conditions that apply in the particular case. Some types of antenna are basically dependent on the presence of the ground for their operation, for example, certain types of vertical antennas at low frequencies. The ground is in fact an integral part of these antenna systems as has been shown in Sec. 1.1.3. In these cases, the pattern must include the effect of the earth.

1.2.3 Near and Far Field Patterns

In principle it is possible to calculate the values of the electric and magnetic field components set up in space by any antenna.

The mathematical difficulties may be formidable if the antenna is complicated, but the calculation is always possible in principle when we use Maxwell's equations. For some simple types of antennas such calculations may be carried out in considerable detail, and the results illustrate certain features that apply to all antennas and are confirmed by experimental investigations of antenna fields. One such feature is that the radiation pattern in the region close to the antenna is not exactly the same as the pattern at great distances. The term near field refers to the field pattern that exists close to the antenna; the term far field refers to the field pattern at great distances. The significance of these terms is conveniently illustrated by considering the fields set up by a simple dipole antenna. The mathematical analysis reveals that in a given direction the total electric field can be expressed as the sum of three terms, each of which decreases in magnitude as the distance from the antenna, R, increases; but they decrease at different rates. The electric field intensity is inversely proportional to the first power of the distance. The dipole field is found to have components that decrease inversely as the square of the distance and inversely as the cube of the distance, in addition to the inverse-first-power term. Mathematically this means that one term contains factors $1 / R$, $1 / R^2$, and $1 / R^3$. The behavior of such terms, as R increases, is illustrated in Fig. 1-3. These terms are equal in magnitude at $R = 1$. Or smaller values of R, the factor $1 / R^3$ is largest, and the $1 / R$ term is smallest. But for large values of R, the $1 / R$ factor is larger than the other two, becoming increasingly so as R increases.

Practically in the far zone the field consists of only the term containing the $1 / R$ factor. The field at great distance from the dipole behaves like the field of point source, with inverse-first-power dependence of the electric field intensity on the distance from the dipole.

At very close distance, on the other hand, $1 / R^3$ and $1/R^2$ terms becomes much larger than the $1 / R$ term dominates the far-field region, as seen in Figure 1-3

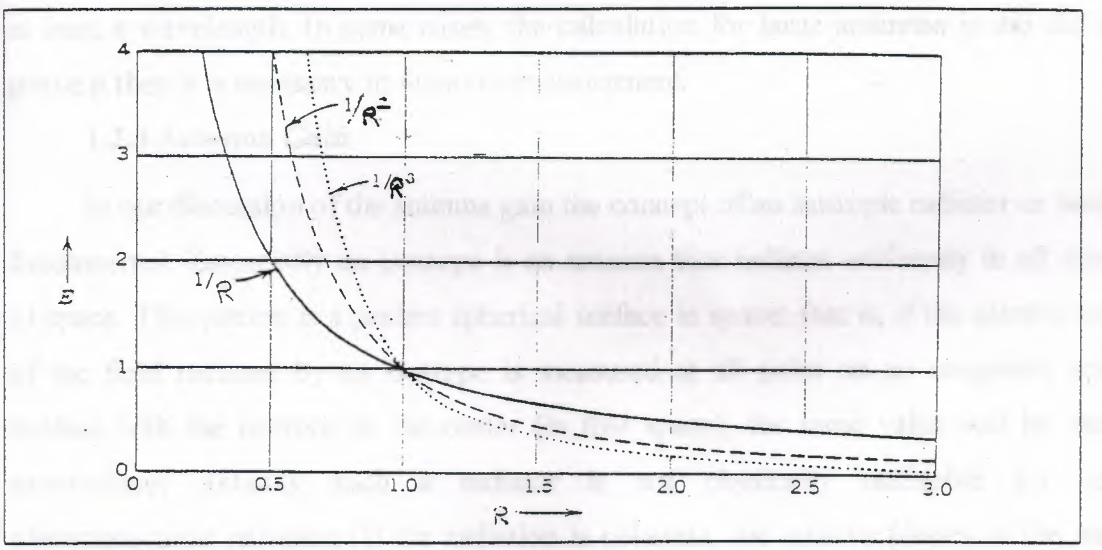


Figure 1.3 Relative variation with distance of short-dipole static ($1 / R^3$), induction ($1 / R^2$), and radiation ($1/R$) field components (electric intensity).

For more complicated antennas, the near field has more complicated dependence on R. The near-and far-field pattern is in general different; that is, plots of relative field strength at a constant distance do not have the same form. In fact, the pattern taken at different distances in the near field will differ from one another, but all patterns taken in the far field are alike, ordinarily it is the radiated power that is of interest, and so antenna patterns are usually measured in the far field region. For pattern measurement it is therefore important to choose a distance sufficiently large to be definitely in the far field, well out of the near field. The minimum permissible distance depends on the dimension of the antenna in relation to the wavelength. An accepted formula for this distance is

$$R_{\min} = \frac{2d^2}{\lambda} \tag{1.3}$$

where R_{Min} is the distance from the antenna, d is the largest dimension of the antenna, and λ is the wavelength.

The factor 2 in this expression is somewhat arbitrary, but it is the factor usually observed in antenna measurement practice. The formula also assumed that d is at least equal to about a wavelength, when d is smaller than λ the distance R_{min} should be equal to at least a wavelength. In some cases, the calculation for large antennas is too difficult to prove it then it is necessary to resort to measurement.

1.2.4 Antenna Gain

In our discussion of the antenna gain the concept of an isotropic radiator or isotrope is fundamental. Essentially an isotrope is an antenna that radiates uniformly in all directions of space. This pattern is a perfect spherical surface in space; that is, if the electric intensity of the field radiated by an isotrope is measured at all point on an imaginary spherical surface with the isotrope at the center (in free space), the same value will be measured everywhere. Actually such a radiator is not physically realizable for coherent electromagnetic radiation (If the radiation is coherent, the relative phases of the waves in different directions from the source maintain a constant difference. For a noncoherent radiator, these phase difference vary in a random manner, or fluctuate. The sun is an example of a noncoherent radiator) all actual antennas have some degree of non-uniformity in their three-dimensional radiation pattern. It is possible for an antenna to radiate uniformly in all directions in a plane, and to design an antenna that has approximate omnidirectionality in three dimensions, but perfect omnidirectionality in three dimensional space can never be achieved. Nevertheless, the concept of such an ideal omnidirectional radiation, an isotrope, is most useful for theoretical purposes. A nonisotropic antenna will radiate more power in some directions than in others and therefore has a directional pattern. Any directional antenna will radiate more power in its direction (or directions) of maximum radiation than an isotrope would, with both radiating the same total power. It is intuitively apparent that this should be so, since the directional antenna sends less power in some directions than an isotrope does, it follows that it must sent more power in other directions, if the total powers radiated are to be the same. This conclusion will now be demonstrated more rigorously. If an isotrope radiates a total power P_t and is located at the center of a transparent (or imaginary) sphere of radius R meters, the power density over the spherical surface is shown bellow

$$P_{isotrope} = \frac{P_t}{4\pi R^2} \quad (\mathrm{W/m^2}) \tag{1.4}$$

Since the total P_t is distributed uniformly over the surface area of the sphere, which is $(4\pi R^2)$ (m^2).

Imagine that in some way it is possible to design an antenna that radiates the same total power uniformly through one half of the same spherical surface, with no power radiated to the other half. Such a fictitious radiator may be called a semi-isotrope. Since the half sphere has a surface area $(2\pi R^2)$ square meters, the power density is

$$P_{semi-isotrope} = \frac{P_t}{2\pi R^2} \quad (\mathrm{W/m^2}) \tag{1.5}$$

Therefore, we get

$$\frac{P_{semi-isotrope}}{P_{isotrope}} = \frac{(P_t / 2\pi R^2)}{(P_t / 4\pi R^2)} = 2 \tag{1.6}$$

The last result shows that at any distance, *R*, the power density radiated by the semi-isotrope is twice as great as that radiated by the isotrope, in the half-sphere within which the semi-isotrope radiates. In this region, therefore, the semi-isotrope is said to have a directive gain of 2. It is fairly apparent that if the radiation were confined to smaller portions of the total imaginary spherical surface, the resulting directive gain would be greater. For example, if the power P_t uniformly into only on fourth of the spherical surface, the directive gain would be 4, and so on.

1.2.4.1 Directive Gain

The directive gain D, of an antenna is defined, in a particular direction, as the ratio of the power density radiated in that direction, at a given distance, to the power density that would be radiated at the same distance by an isotrope radiating the same total power. The directive gain of a semi-isotrope in the hemisphere into which it radiates is 2; its directive gain in the other hemisphere (where no power is radiated) is zero,

Thus D of an antenna is defined as a quantity that may be different in different directions. In fact, the relative power density pattern of an antenna becomes a directive gain pattern if the power density reference value is taken as the power density of an isotrope radiating the same total power (instead of using as a reference the power density of the antenna in its maximum radiation direction). In this case, we define the direction gain of the antenna as

$$D = \frac{P_{antenna}}{P_{isotrope}} \tag{1.7}$$

were P antenna is the antenna power density, from Eqs.1-2 and 1-4, we find that

$$D = \frac{4\pi R^2 E^2}{377 P_t} = \frac{4\pi R^2 P_{antenna}}{P_t} \tag{1.8}$$

where P_t is the total radiation power. If P_t represents the input power to the actual antenna rather than the power radiated, G should be substituted for D on the left hand side of this equation, that is, give the power gain rather than the directive gain. The efficiency factor ξ is the ratio of the power radiated by the antenna to the total input power, it is a number between zero to unity, and it connects the direction gain D with the power gain G in

$$G = \xi D \tag{1.9}$$

The maximum directive gain (directivity) is quite important value, as we will see in gain measurement later. This value can be calculated from

$$D_{\max} = \frac{4\pi}{\int_0^{2\pi}\int_0^{\pi}[E(\theta,\phi)/E_{\max}]^2 \sin\theta \; d\theta \; d\phi} \tag{1.10}$$

Once the directivity D_{max} has been calculated from the relative pattern, the directive gain in any other direction θ, ø can also be simply determined from the following relationship

$$D_{(\theta_1,\phi_1)} = D_{\max}\left[\frac{E(\theta_1,\phi_1)}{E_{\max}}\right]^2 \tag{1.11}$$

1.2.4.2 Gain in Decibels

Antenna gain is a power ratio. The gain of practical antennas may be range from zero to as much as 10,000 or more. As with any power ratio, antenna gain may be expressed in decibels. To illustrate in terms of the antenna power gain G, the value in decibels will be donated by G (dB) and is given by G (dB)=10 $\log_{10}$ G. The directive gain in decibels is calculated from the same formula, with D substituted for G.

1.2.4.3 Practical Significance of Power Gain

It is apparent for a given amount of input power in antenna; the power density at a given point in space is proportional to the power gain of the antenna in that direction. Therefore increasing the power gain of the transmitting antenna, without increasing the transmitting power can increase the signal available to a receiving antenna at that location. A transmitter with a power output of 1000 watts and antenna with a power gain of 10 (10dB) will provide the same power density at a receiving point as will a transmitter of 500 watts power and an antenna power gain of 20 (13dB). Obviously this relationship has great economic significance. Sometimes it may be much less expensive to double the gain of the antenna (add 3dB) than it would be to double the transmitter power (though in other cases the converse may be true). But generally speaking it is desirable to use as much antenna gain as may feasibly be obtained, when it is desired to provide the maximum possible field strength in a particular direction.

1.2.5 Beamwidth

When the radiated power of an antenna is concentrated into a single major lobe as seen in the pattern of Fig. 1-2, the angular width of this lobe is the beamwidth. The term is applicable only to antennas whose patterns are of this general type. Some antennas have a pattern consisting of many lobes, all of them more or less comparable in their maximum power density, or gain, and not necessarily all of the same angular width. But large classes of antennas do have patterns to which the beamwidth parameter may be appropriately applied.

1.2.5.1 Definition of Beamwidth

It is logical to define the width of a beam in such a way that it indicates the angular range within which radiation of useful strength is obtained, or over which good reception may be expected. From this point of view the convention has been adopted of measuring bean-width between the points on the beam pattern at which the power density is half the value at the maximum. In a plot of the electric intensity pattern, the corresponding points are those at which the intensity is equal to 0.707 of the maximum value. The angular width of the beam between these points is called the half-power beamwidth. When a beam pattern is plotted with the ordinate scale in the minus 3dB points. For this reason the half power beamwidth is often referred to as the -3dB beamwidth. Figure 2-4 illustrates the procedure of determining the -3dB beamwidth on a rectangular pattern plot.

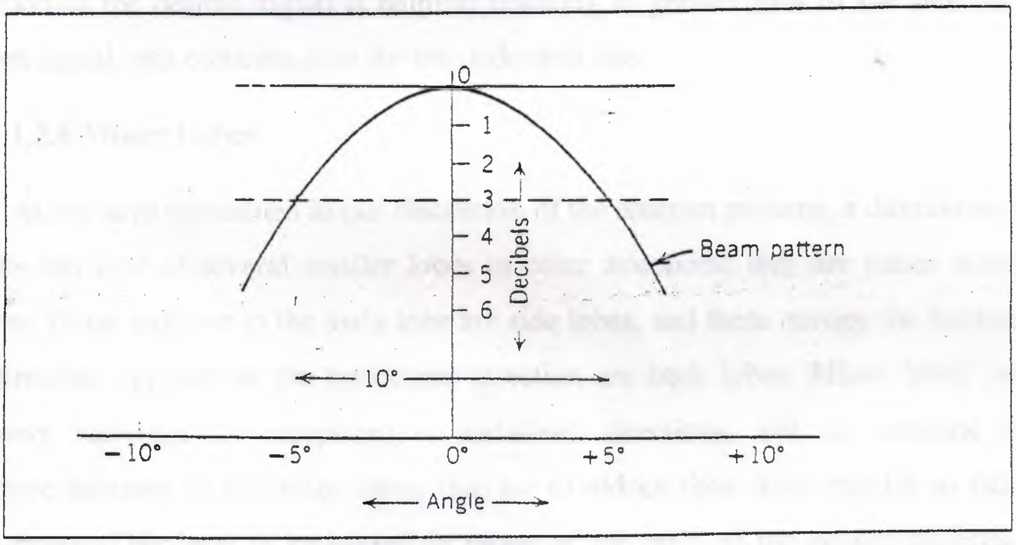


Figure 1.4 Determination of half-power (3dB-down) beamwidth.

This criterion of beamwidth, although adequate and convenient in many situations, it does not always provide a sufficient description of the beam characteristics. When beams have different shapes. An additional description may be given by measuring the width of the beam at several points, As an example, at -3dB, -10dB, and at the nulls (if they are present).

Some beams may have an asymmetric shape. Special methods of describing such beams can be employed. In the final analysis the best description of a beam is a plot of its pattern.

1.2.5.2 Practical Significance of Beamwidth

If an antenna has a narrow beam and is used for reception, it can be used to determine the direction from which the received signal is arriving, and consequently it provides information on the direction of the transmitter. To be useful for this purpose, the antenna beam must be steerable; that is, capable of being pointed in various directions. It is intuitively apparent that for this direction-finding application, a narrow beam is desirable and the accuracy of direction determination will be inversely proportional to the beamwidth. In some applications receiving may be unable to discriminate completely against an unwanted signal that is either at the same frequency as the desired signal or on nearly the same frequency. In such a case, pointing a narrow receiving antenna beam in the direction of the desired signal is helpful; resulting in greater gain of the antenna for the desired signal, and reducing gain for the undesired one.

1.2.6 Minor Lobes

As we have mentioned in our discussion of the antenna patterns, a directional antenna usually has lobe of several smaller lobes in other directions; they are minor lobes of the pattern. Those adjacent to the main lobe are side lobes, and these occupy the hemisphere in the direction opposite to the mainbeam direction are back lobes. Minor lobes ordinarily represent radiation (or reception) in undesired directions, and the antenna designer therefore attempts to minimize them, that are to reduce their level relative to that of the main beam. This level is expressed in terms of the ratio of the power densities in the mainbeam maximum and in the strongest minor lobe, and often expressed in decibels.

Since the side lobes are usually the largest of the minor lobes, this ratio is often called the side-lobe ratio or side-lobe level. A typical side-lobe level, for an antenna in which some attempt has been made to reduce the side-lobe level, is 20dB, which means that the power density in the strongest side lobe is 1 % of the power density in the main beam. Side-Lobe levels of practical well-designed directional antennas typically range from about 13dB (power-density ratio 20) to about 40dB (power density ratio 10,000). Attainment of a side-lobe level better than 30dB requires very careful design and construction. Figure 1-5 shows a typical antenna pattern with a main beam and minor lobes, plotted on a decibel scale to facilitate determination of the side-lobe level, which is here seen to be 25dB.

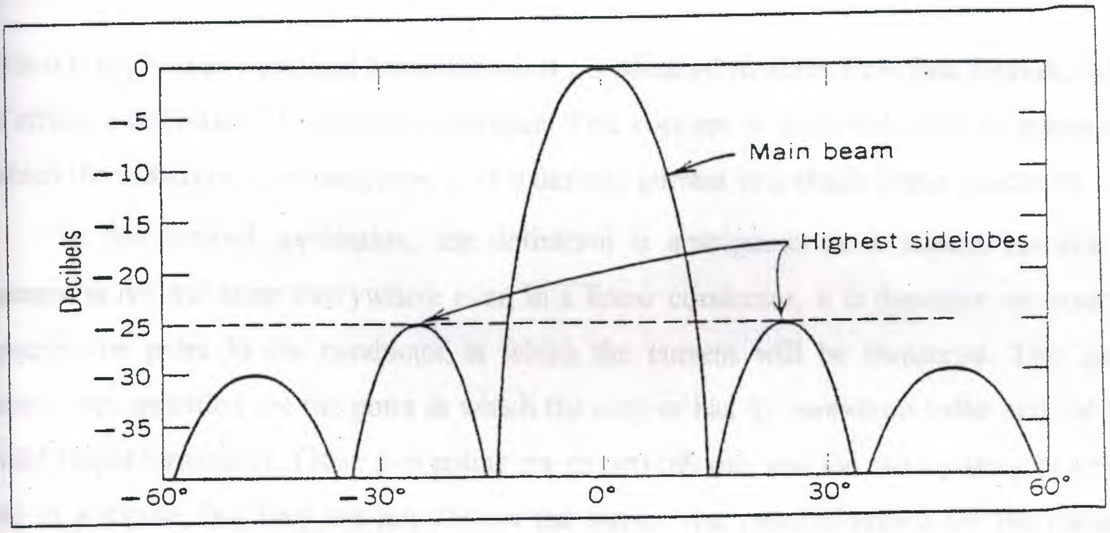


Figure 1.5. Decibel pattern plot indicated side lobe level.

In some applications side lobes are not especially harmful unless their level becomes comparable to the main-beam level. In other applications it may be important to hold the side-level to an absolute minimum. In most radar systems, a low side-lobe level is important. If the radar is very sensitive, a large target located in the direction of one of the antenna side lobes (or even a back lobe) may appear on indicator oscilloscope as though it were a target in the main beam.

1.2.7 Radiation Resistance and Efficiency

In a large class of antennas the radiation is associated with a flow of RF current in a conductor or conductors. As is well known in elementary electric circuit theory, when a current I flows in a resistance R, an amount of power $P = RI^2$ will be dissipated, that is, electrical energy will be converted into heat at this rate. In an antenna, even if there is no resistance in the conductors, the electrical energy supplied by the transmitter is lost just as though it had been converted in to heat a resistance, although in fact it is radiated. It is customary to associate this loss of power, through radiation, with a fictitious radiation resistance that bears the same relationship to the current and the radiation power as an actual resistance bears to the current and dissipated power. If the power radiated by the antenna is P and the antenna current is I, the radiation resistance is defined as

$$R_r = \frac{P}{I^2} \tag{1.12}$$

When P is given in watts and I in amperes, R_r is obtained in ohms from this formula, which is effect, a definition of radiation resistance. This concept is applicable only to antennas in which the radiation is an associated with a definite current in a single linear conductor.

In this limited application, the definition is ambiguous as it stands, because the current is not the same everywhere even in a linear conductor, it is therefore necessary to specify the point in the conductor at which the current will be measured. Two points sometimes specified are the point at which the current has its maximum value and the feed point (input terminals). These two points are sometimes one and the same points, as center-fed in a dipole, but they are not always the same. The value obtained for the radiation resistance of the antenna depends on which point is specified; this value of the radiation resistance referred to that point. The current maximum of a standing-wave pattern is known as a current loop, so the radiation resistance referred to the current maximum is sometimes called the loop radiation resistance.

The word maximum here refers to the effect current rms in that part of the antenna where it has its greatest value. It does not mean the peak value of the current at this point during the RF cycle, when Eq.1-12 is used as the definition. In some texts, however, formulas for radiation resistance are written in terms of this peak value, which is the amplitude of the current sine wave. Equation 1-12 will yield a value of radiation resistance only half as great as the true value -if the current amplitude is used for I, the correct formula in terms of the current amplitude I_0, is R_r=2P/I^2, note that $I_0 = \sqrt{2 I_{rms}}$.

The radiation resistance of some types of antennas can be calculated, when there is clearly defined current value to which it can be referred, but for other types the calculation cannot be made practically, and the value must be obtained by measurement. Methods of making such a measurement will be described later.

The typical values of the loop radiation resistance of actual antennas range from a fraction of an ohm to several hundred ohms. The very low values are undesirable because they imply large antenna current, and therefore the possibility of considerable ohmic loss of power, that is, dissipation of power as heat rather than as radiation. An excessively high value of radiation resistance would also be undesirable because it would require a very high voltage to be applied to the antenna. Very high voltage values do not occur in practical antennas, because there is always some ohmics resistance whereas very low values sometimes do occur unavoidably.

Antennas always do have some ohmic resistance, although sometimes it may be so small as to be negligible. The ohmic resistance is usually distributed over the antenna, and since the antenna current varies, the resulting loss may be quite complicated to calculate. In general, however, the actual loss can be considered to be equivalent to the loss in a fictitious lumped resistance placed in series with the radiation resistance. If Ro denotes this equivalent ohmic loss resistance, the full power (dissipated plus radiated) is $I^2=(R_0+R_r)$, whereas the radiation power is $I^2 R_r$. Hence the antenna radiation efficiency ξ_r is given by

$$\xi_r = \frac{R_r}{R_0 + R_r} \tag{1.13}$$

It must be acknowledged that this definition of efficiency is not really very useful even though it may occasionally be convenient. The fact is both Ro and R_o is fictitious quantities, derived from measurements of current and power; R_r is given in these terms by Eq.1-12, and R_o is correspondingly equal to P_o / I^2. Making these substitutions into Eq.1-13, then it gives the more basic definition of the efficiency:

$$\xi_r = \frac{P_r}{P_0 + P_r} \tag{1.14}$$

1.2.8 Input Impedance

An antenna whose radiation results directly from the flow of RF current in a wire or other linear conductor must somehow have this current introduced into it from a source of RF power transmitters. The current is usually carried to the antenna through a transmission line. To connect the line to the antenna, a small gap is made in the antenna conductor, and the two wires of the transmission line are connected to the terminals of the gap at antenna input terminals. At this point of connection the antenna presents load impedance to the transmission line. This impedance is also the input impedance of the antenna and it is equal to the characteristic of the line Z_o, the input impedance of the antenna is one of it is important parameters. Measurement of the antenna input impedance would be discussed later. The input impedance determines how large a voltage must be applied at the antenna input terminals to obtain the desired current flow and hence the desired amount of radiated power. Thus, the impedance is equal to the ratio of the input voltage E_i to the input current I_i and it can be written as

$$Z = \frac{E_i}{I_i} \tag{1.15}$$

Which is in general complex. If the gap in the antenna conductor (feed point) is at a current maximum, and if there is no reactive component to the input impedance, it will be equal to the sum of the radiation resistance and the loss resistance; that is

$$Z_i = R_i = R_r + R_0 \tag{1.16}$$

If this reactance has a large value, the antenna-input voltage must be very large to produce an appreciable input current. If in addition the radiation resistance is very small, the input current must be very large to produce appreciable radiated power. Obviously this combination of circumstances, which occurs with the short dipole antenna that must be used at very low frequencies, results in a very difficult feed problem or impedance-matching problem, they are usually fed by waveguides rather than by transmission line. The equivalent of input impedance can be defined at the point of connection of the waveguide to the antenna, just as waveguides have characteristic wave impedance analogous to the characteristic impedance of a transmission line. For some types of antennas consisting of current-carrying conductors this is difficult, and it may even be difficult to define input impedance. This is true, as an example, for an array of dipoles, when each dipole is fed separately; sometimes each dipole, or groups of dipole, will be connected to separate transmitting amplifiers and receiving amplifiers. The input impedance of each dipole or group may then be defined, but the concept becomes meaningless for the antenna as a whole, as does also for simple linear-current radiation elements; but they comprise a very large class of antennas.

1.2.9 Bandwidth

All antennas are limited in the range of frequency over which they will operate satisfactorily. This range is called the bandwidth of the antenna. Bandwidth is a concept that is probably familiar in other applications, sometimes by another name. For example, a television I-f amplifier must have a bandwidth of approximately 4MHz in order to pass all the frequency components of a television signal. A television-transmitting antenna must have sufficient bandwidth to receive all the channels to which the receiving set can be tuned. If an antenna were capable of operating satisfactory from a minimum frequency of 155MHz to a maximum frequency of 205MHz, its bandwidth would be 10MHz. It would also be said to have a 5% bandwidth (the actual bandwidth divided by the center frequency of band, times 100). Some antennas are required to operate only at a fixed frequency with a signal that is narrow in its bandwidth; consequently there is no bandwidth problem in designing such an antenna. In other applications much greater bandwidths may be required; in such cases special techniques are needed. Some recent developments in broadband antennas permit bandwidths so great as they are described by giving the numerical ratio of the highest to the lowest operating frequency, rather than as a percentage of the center frequency. In these terms, bandwidths of 20 to 1 are readily achieved with these antennas, and ratios as great as 100 to 1 are possible.

1.2.10 Beam Area or Beam Solid Angle

An arc of a circle seen from the center of this circle subtends an angle θ. Thus, referring to Fig.2-6a, the arc length θR subtends the angle θ. The total angle in the circle is 2π rad so the total arc length is $2\pi R$. By using the same concept, an area A of a sphere surface seen from the center of the sphere subtends a solid angle Ω as shown in Fig. 1-6b

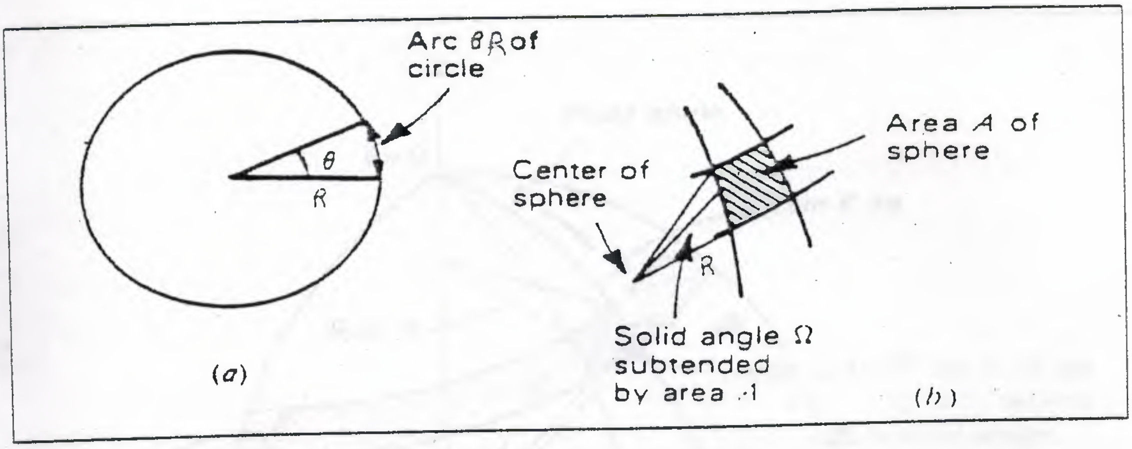


Figure 1.6 (a) Arc length Rθ of circle has radius R subtends the angle θ. The area A of a sphere of radius R subtends a solid angle Ω.

The total solid angle subtended by the sphere is 4 π steradians (or square radians), abbreviated sr.

By using Fig. 1-7 we can discuss the solid angle in more details. From Fig. 1-7, it is shown that the solid angel dΩ subtended by dA is

$$d\Omega = \sin\theta \; d\theta \; d\phi \tag{1.17}$$

To more declaration, the incremental area dA of the surface of a sphere is given by $dA = (R\sin\theta \; d\phi)\,(R \; d\theta) = R^2 \sin\theta \; d\phi \; d\theta = R^2 \; d\Omega$ the area of the strip of width R dθ extending around the sphere at a constant angle θ is given by dAs= (2 π R sinθ)(R dθ). Integrating this for θ values from 0 to π yields the area of the sphere. Thus,

$$Area\ of\ sphere = 2\pi R^2 \int_0^{\pi} \sin\theta \; d\theta = 4\pi R^2 \tag{1.18}$$

By comparing this result with dA = (R Sin θ dθ(R dθ) = R^2 sin θ dθ = R^2dΩ we fined that dΩ for the whole sphere is 4 π

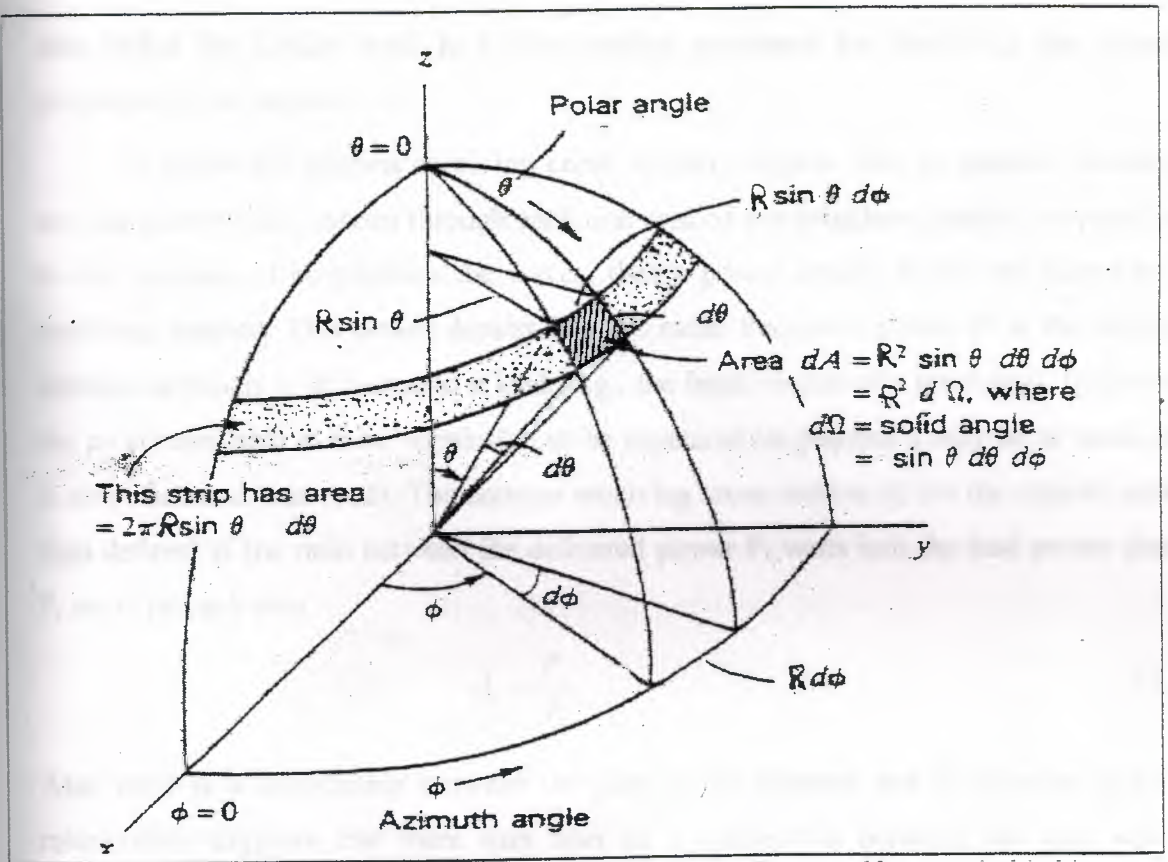


Figure 1-7 Spherical coordinates in relation to the area dA solid angle $d\Omega = \sin\theta \, d\theta \, d\phi$

Now the beam area (or beam solid angle) Ω_A for an antenna is given by the integral of the normalized power pattern over a sphere (4 π, sr)

$$\Omega_A = \int_0^{2\pi}\int_0^{\pi} P_n(\theta,\phi)\, d\Omega \tag{1.19}$$

1.2.11 Capture Area or Receiving Cross Section

Although there is a reciprocal relationship between the transmitting and the receiving properties of antennas, it is sometimes more convenient to describe the receiving properties in a somewhat different way. Whereas the power gain is the natural parameter to use for describing the increases power density of the transmitted signal due to the directional properties of the antenna, a related quantity called the receiving cross section, sometimes also called the capture area, is a more natural parameter for describing the reception properties of the antenna.

To define the antenna receiving cross section, suppose that an antenna radiates an amount power which passes through each unit area of any imaginary surface perpendicular to the direction of propagation the waves, then a power density P_i will be passed to the receiving antenna. This power density induces radio frequency power P_r at the receiving antenna terminals is delivered to a load (e.g., the input circuit of a receiving). In principle the power available at these terminals can be measured (in practice it may be so small, so it is amplified and then read). The antenna receiving cross section A_r (or the capture area) is then defined as the ratio between the delivered power P_r watts into the load power density P_t watts per unit area

$$A_r = \frac{P_r}{P_i} \tag{1.20}$$

Also there is a relationship between the gain of the antenna and its physical size, this relationship suggests that there may also be a connection between the gain and the receiving cross section area and this indeed turns out to be true. The receiving cross section area in isotropic A_{ro} is given as

$$A_{r0} = \frac{\lambda^2}{4\pi} \Rightarrow A_r = \frac{G\lambda^2}{4\pi} \tag{1.21}$$

where $G = \xi D$, λ is the wavelength, note that λ has relationship with the size, then A_r , G and the size. Equation 1-20 may be proved theoretically and verified experimentally. From this relationship it follows that

$$D = \left(\frac{4\pi\, A_r}{\xi\, \lambda^2}\right) \tag{1.22}$$

where D is the directive gain. It is clear from this relationship that the gain increases when A_r increases, and λ and ξ decrease, and vice versa. Thus, the power is

$$P_r = \xi \left(\frac{P_i D \lambda^2}{4\pi}\right) \tag{1.23}$$

Therefore the concept of the receiving cross section of an antenna is not a necessary one. It is possible to calculate the received-signal power without using equation 1-23.

CHAPTER TWO

CIRCULARLY POLARIZED ANTENNAS

2.1 Panel Types Antennas

In many cases, the supporting structure is a triangular or square tower. Panel antennas are primarily used to control or minimize the reflections from the supporting structure. Some panel antennas are made of a single horizontal dipole or two crossed dipoles (circularly polarized panel) in front of a reflector.

The reflector can be a flat panel, a comer reflector, or a pillbox (commonly referred to as *cavity-backed)*. The reflector is usually a wire grid for VHF or a solid sheet for UHF.

In order to obtain an omni directional radiation pattern, three- or four-panel antennas are placed around a triangular or square tower, respectively. In general, panel antennas with 6-dB beam width of 90 and 120 ° are used for arrays around square and triangular towers, respectively. When several panels are arranged around a cylindrical structure in a single layer, as shown in Fig. 2.1,

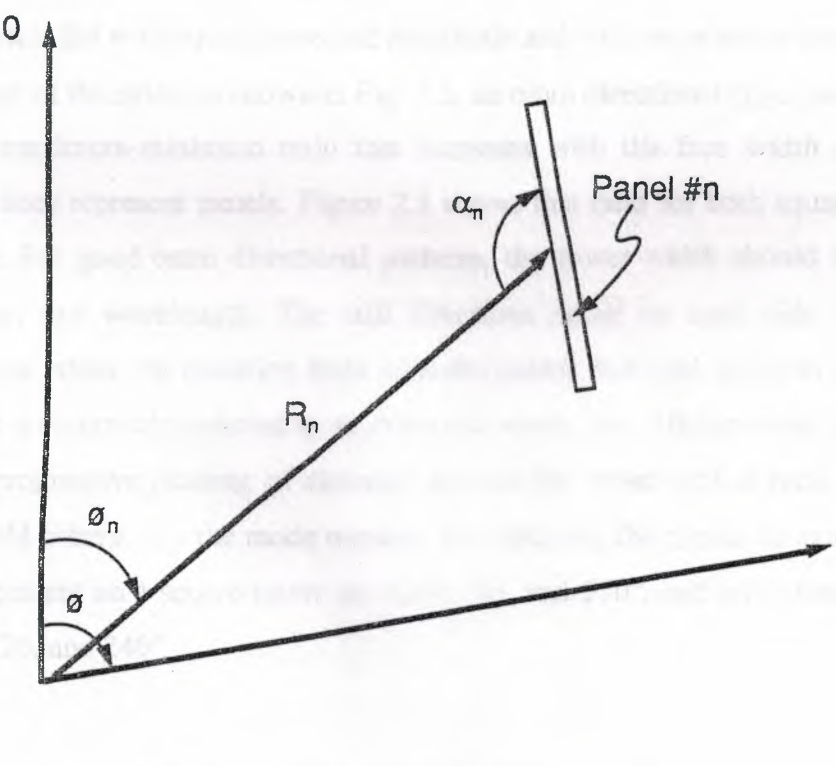


Figure 2.1 Geometry for equation (2.1)

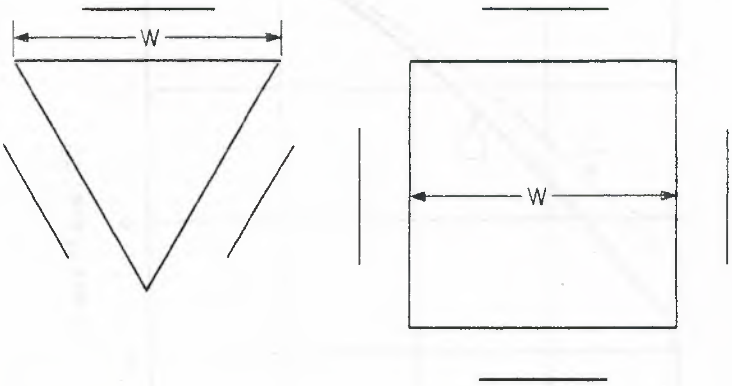


Figure 2.2 Panel-type antennas for triangular and square

The combined pattern $E(\Phi)$ may be calculated by using the following expression:

$$E(\Phi) = \sum_{n=1}^{N} I_n M_n(\psi) \exp \mathrm{i}\left[\xi_n(\psi) + \sigma_n + kR_n \cos(\Phi - \Phi_n)\right] \qquad \textbf{(2-1)}$$

Where $I_n e^{i\sigma_n}$ = excitation current of the *nth* panel

$M_n(\psi)e^{i\xi_n(\psi)}$ = pattern of the *nth* panel

Φ_n = polar angle of the *nth* panel

R_n = length of the radial to the *nth* panel

$\Psi = \pi - \alpha_n\Phi - \Phi_n\alpha_n$ = tilt angle of the *nth* panel

With the panels fed with equal phase and amplitude and with the antenna elements placed in the center of the sides, as shown in Fig. 2.2, an omni directional type pattern is obtained with a maximum-minimum ratio that increases with the face width of the tower. The short lines represent panels. Figure 2.3 shows this ratio for both square and triangular towers. For good omni directional patterns, the tower width should not be much greater than one wavelength. The null directions occur on each side of the crossover directions where the radiation from adjacent panels does not arrive in phase. This arrangement is commonly referred to as *azimuthal mode zero.* Higher-order modes are obtained by progressive phasing of elements around the tower with a total phase progression *of360M,* where *M* is the mode number. For instance, the phases of panels in the mode 1 arrangement on a square tower are 0,90, 180, and *270°,* and on a triangular tower they are 0,120, and 240°.

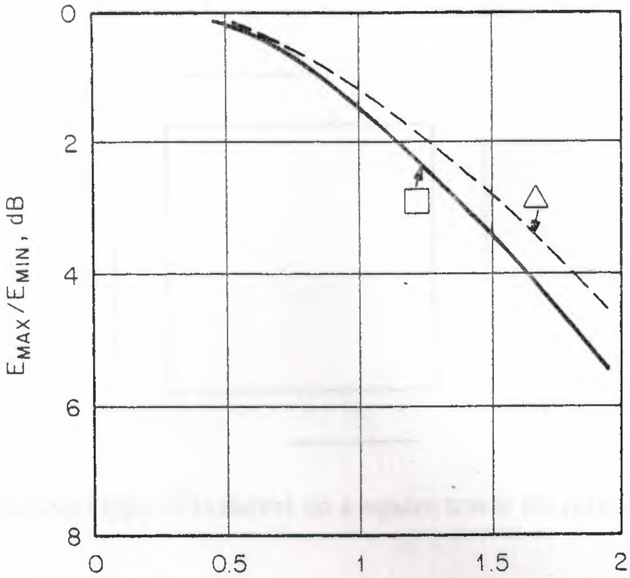


Fig 2.3 Maximum-minimum ratio versus tower width in wavelengths for triangular and square towers.

Higher modes can be used to achieve a broader bandwidth for the antenna-input impedance. Use of higher modes requires different lengths of feed lines from the power divider. This provides reflection cancellation at the power divider, which in turn allows lower VSWR at the power divider than at the panels. To eliminate the difference in the arrival phase of adjacent panels in the crossover direction, the elements are displaced laterally as shown in Fig. 2.4 Equations (2.2) and (2.3) give the amount of offset S for triangular and square towers for mode 1, respectively:

$$S = \lambda /(4\sqrt{2}) \tag{2.2}$$

$$S = \lambda / 3 \tag{2.3}$$

An adverse effect of using higher-order modes with panels having high VSWR is that multiple reflections produce power-division errors to the panels, which result in pattern distortions.

Skewed-panel antennas may be placed on the comers of large-face towers (such as those measuring five wavelengths), as shown in Fig. 2.5; the antenna elements are placed on narrow panels. The panels are skewed so that the crossover direction coincides with the tower face. Thus the relative phase of the radiation from adjacent antennas varies more slowly near the crossover direction than in a nonskewed arrangement. Theoretically, this should greatly improve the pattern circularity.

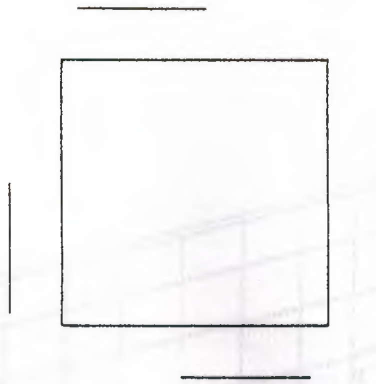


Figure 2.4 Offset-panel type of radiators on a square tower for reflection cancellation

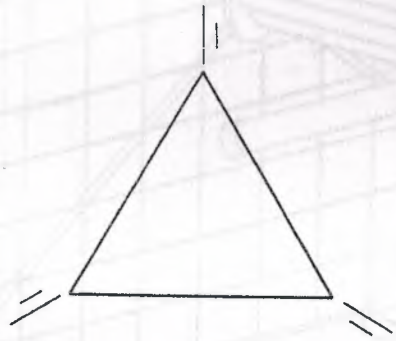


Figure 2.5 Skewed-panel type of radiators for triangular towers

However, reflections from the tower members and the backside of the panel degrade the pattern. In general, it is difficult to achieve pattern circularities better than ± 2 dB.

2.2. Crossed-Dipole Panel Antennas

A common technique for producing circular polarization has been to place two linear dipoles at right angles in front of a reflecting screen and to feed them with equal voltage magnitudes and with a 90°-phase difference. However, the azimuth beam width for horizontal and vertical polarization is about 60 and 120°, respectively. Thus the axial ratio is low only for directions near the normal to the screen. This deficiency may be corrected in several ways.

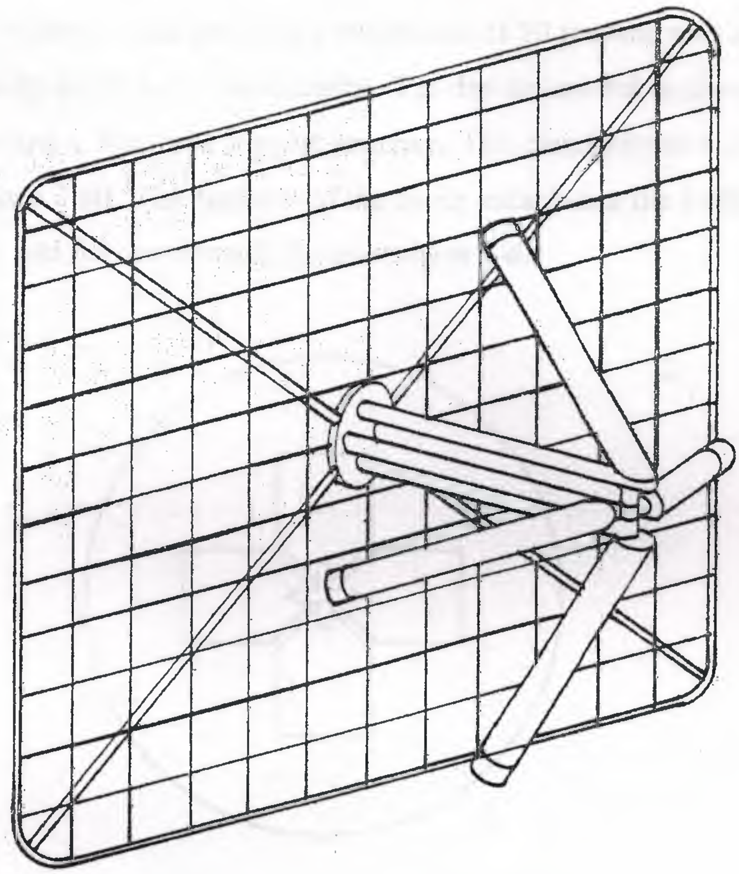


Figure 2.6 Crossed V screen dipoles

V dipoles, as illustrated in Fig. 2.6, may be used to increase the azimuth beam width for horizontal polarization. The crossed dipoles may be identical and fed in phase quadrature or be unequal in length and fed in phase with the lengths adjusted to produce quadrature currents in the dipoles. Three crossed V dipoles may be placed around a triangular tower to obtain a circularity of ±2 dB and a maximum axial ratio of 4 dB. In this arrangement, the circularity of the vertically polarized component of the azimuth pattern is degraded relative to the circularity of the horizontally polarized component. Furthermore, the peaks of one pattern coincide with the valleys of the other. Another version of this type of antenna has three reflecting panels placed in an Y configuration and supported by a central mast. Three crossed V dipoles are placed in the 120 ° sectors formed by the panels. This provides a more compact structure than the triangular tower.

A better technique for equalizing the azimuth beamwidth for vertical and horizontal polarization is to enclose flat crossed dipoles in a cylindrical cavity,[6] as shown in Fig. 2.7 The use of cylindrical cavitybacked reflectors results in a good match of the vertically and horizontally polarized azimuth patterns. The length-to-width ratio of the flat dipoles is about 3 and provides a bandwidth of 10 percent with a VSWR less than 1.1:1. The cavity depth is 0.2 wavelengths. The dipoles are fed in phase quadrature by two baluns forming a four-tube support structure. The circularity is ± 2 dB, and the axial ratio is less than 2 dB. The diameter of the cavity establishes the 6-dB beamwidth. Cavities with 0.65- and 0.8-wavelength diameters have 6-dB

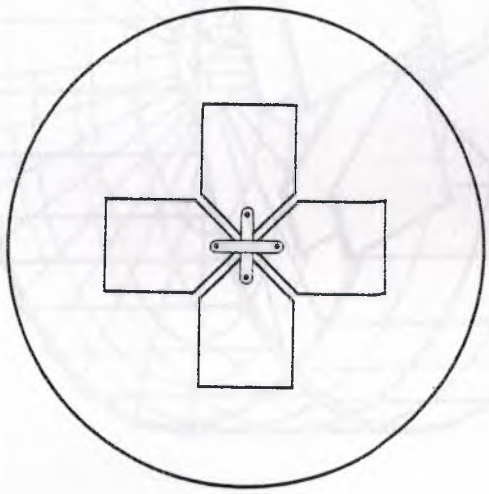


Figure 2.7 Crossed dipoles in a cylindrical cavity.

beamwidths of 120 and 90°, respectively. The quadrature feeding of crossed dipoles results in a broadband small VSWR. To maintain an axial ratio of less than 3 dB, it is essential that each one of the dipoles have a low VSWR over the band. This is achieved by placing a fan-shaped conducting screen (or flat sheet) over the crossed dipoles. This

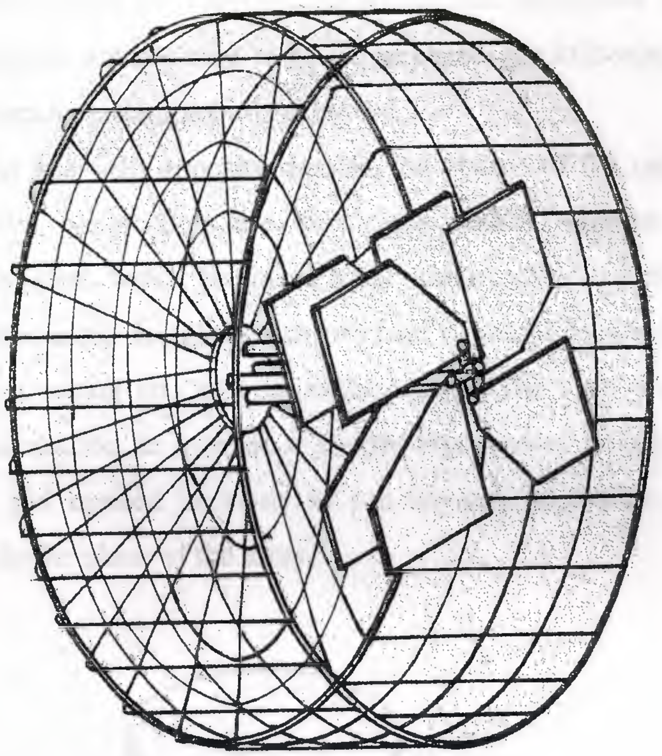


Figure 2.8 Broadband cavity-backed antenna.

conductor acts as a common sleeve for both dipoles. By adjusting the dimensions of this sleeve and its height above the crossed dipoles, the axial ratio over the 20 percentbandwidth is improved considerably. Figure 2.8 shows a version of this concept.

Another approach is to place four half-wavelength dipoles in a square arrangement, withthe side of the square being somewhat larger than a half-wavelength. The vertical andhorizontal dipoles are fed in phase quadrature. Since the 6-dB azimuth beamwidth isabout 90°, four panels around a square tower may be used for omnidirectionalapplications.

2.3. Slanted-Dipole Antennas

Many circularly polarized FM and TV broadcast antennas are based on the concept of a circular array of slanted dipoles. The dipoles may be linear, V-shaped, curved, or of similar configuration. Each dipole radiates linear polarization, but the slant angles and diameter of the circular array are adjusted so that an omnidirectional, circularly polarized radiation pattern is obtained. The term *circular array* is used here to include two or more dipoles placed on a circle with rotational symmetry. Fig 2.9 shows a circular array of several slanted dipoles uniformly spaced around a cylindrical conductor with a slant angle ψ. Figure 2.10 shows the coordinate system for this arrangement. If the dipoles are assumed to be fed in phase, the following is a simplified explanation of how circular polarization is achieved.

In a direction in line with opposite dipoles, the phases of the radiation from the vertical and horizontal components are vectorially added, whereas the horizontal components are subtracted, which produces a 90°-phase difference between the two polarizations. The slant angle is adjusted to produce equal magnitudes of vertical and horizontal polarization, taking into account radiation from the other two dipoles, which results in circular polarization. In general, *N* slanted dipoles may be placed in a circular array, of radius p_2, and excited in mode M (an integer) to radiate omnidirectional circular polarization in the plane of the array.

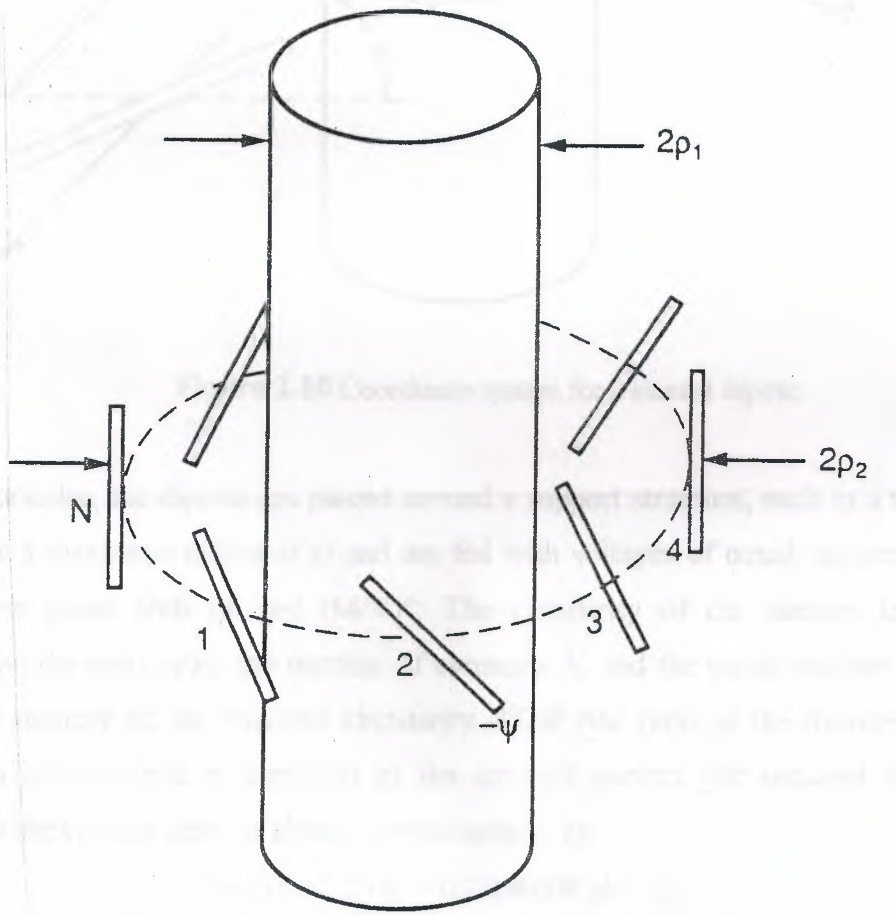


Figure 2.9 *N* skewed (slanted) dipoles placed around a conducting cylinder.

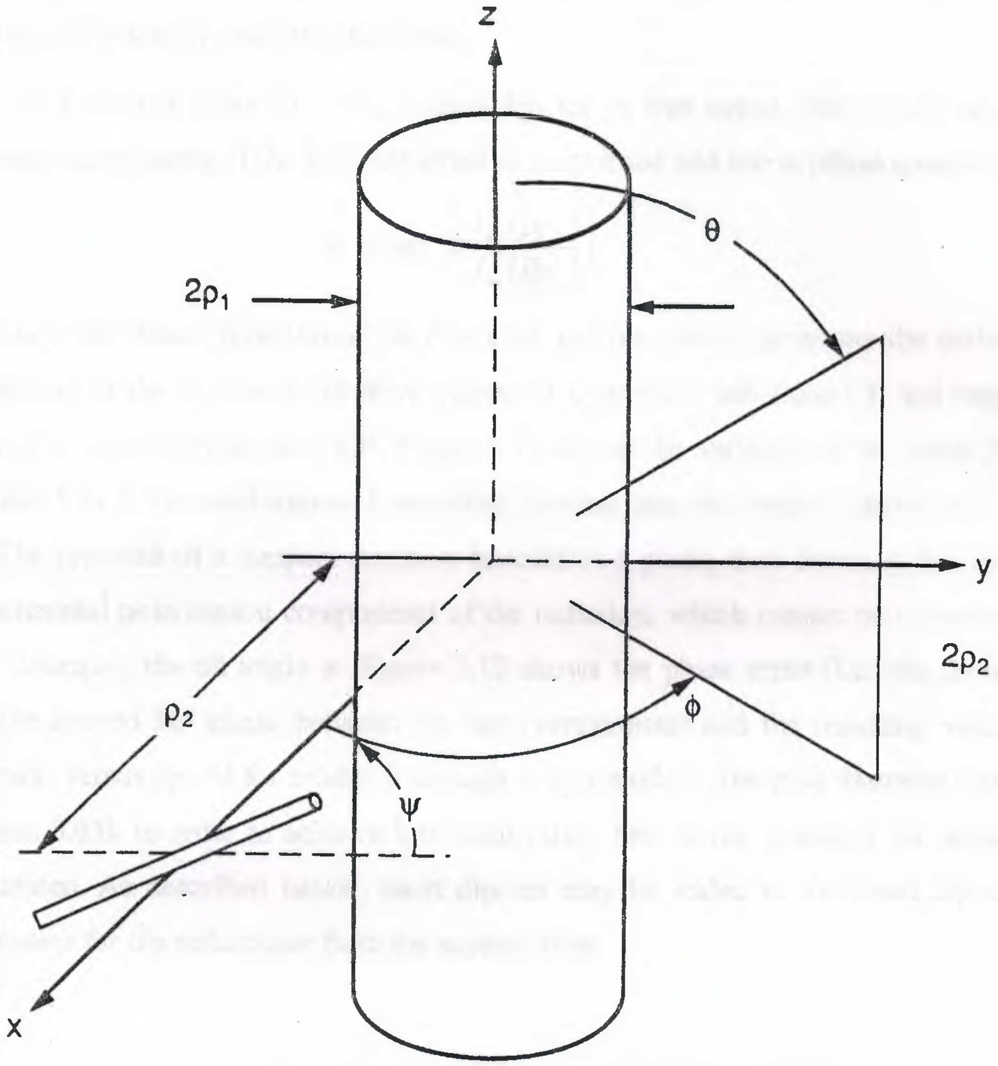


Figure 2.10 Coordinate system for a slanted dipole.

In practice, the dipoles are placed around a support structure, such as a tower or a pole, with a maximum radius *of pi* and are fed with voltages of equal magnitude and a progressive phase shift of 360 (M/V)°. The circularity of the pattern, in general, depends on the ratio p_2/p_1, the number of elements *N,* and the mode number *M.* Given the mode number *M,* the required circularity *WOW (the* ratio of the maximum to the minimum electric field in decibels) of the azimuth pattern, the required number of dipoles in the circular array is given, approximately, by

$$N \approx 2.1 + 2.25M - 0.25(WOW)M / \beta p_2 \qquad (2.4)$$

where $\beta = 2\pi / \lambda$. This formula is obtained by linear regression methods from a series of calculations. The end result is that the spacing between the dipoles must be about one-half wavelength for mode 0 and less than that for higher-order modes in order to obtain good circularity and low axial ratio.

In a circular array of short, slanted dipoles in free space, the vertical and the horizontal components of the field are equal in magnitude and are in phase quadrature if

$$\psi = \tan^{-1}\left[\frac{J_M(\beta p_2)}{J_M'(\beta p_2)}\right] \tag{2.5}$$

where J_M is the Bessel function of the first kind and the prime represents the derivative with respect to the argument. Positive values of ψ produce left-hand CP and negative values of ψ produce right-hand CP. Figure 2.11 shows the variation of ψ versus $\beta p_2/M$ for modes 0 to 4. The well-known Lindenblad antenna uses the mode 0 excitation.

The presence of a support structure introduces a phase error between the vertical and horizontal polarization components of the radiation, which cannot be compensated for by changing the tilt angle ψ. Figure 2.12 shows the phase error (i.e., the deviation from the desired 90° phase between the two components) and the resulting minimum axial ratio versus $\beta p_2/M$ for modes 0 through 4. For mode 0, the pole diameter must be less than 0.03λ in order to achieve low axial ratio. This is not practical for broadcast applications. As described below, short dipoles may be added to the tilted dipoles to compensate for the reflections from the support pole.

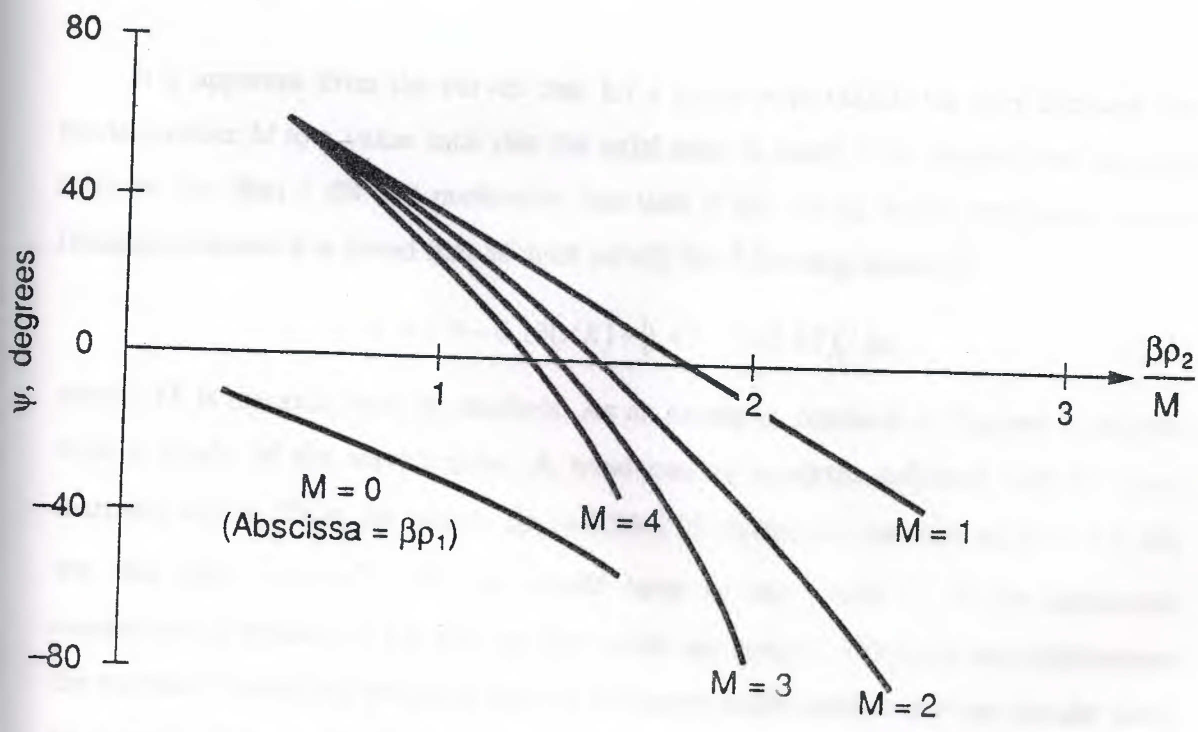


Figure 2.11 Skew angles for a circular array of dipoles in free space for modes 0 to 4.

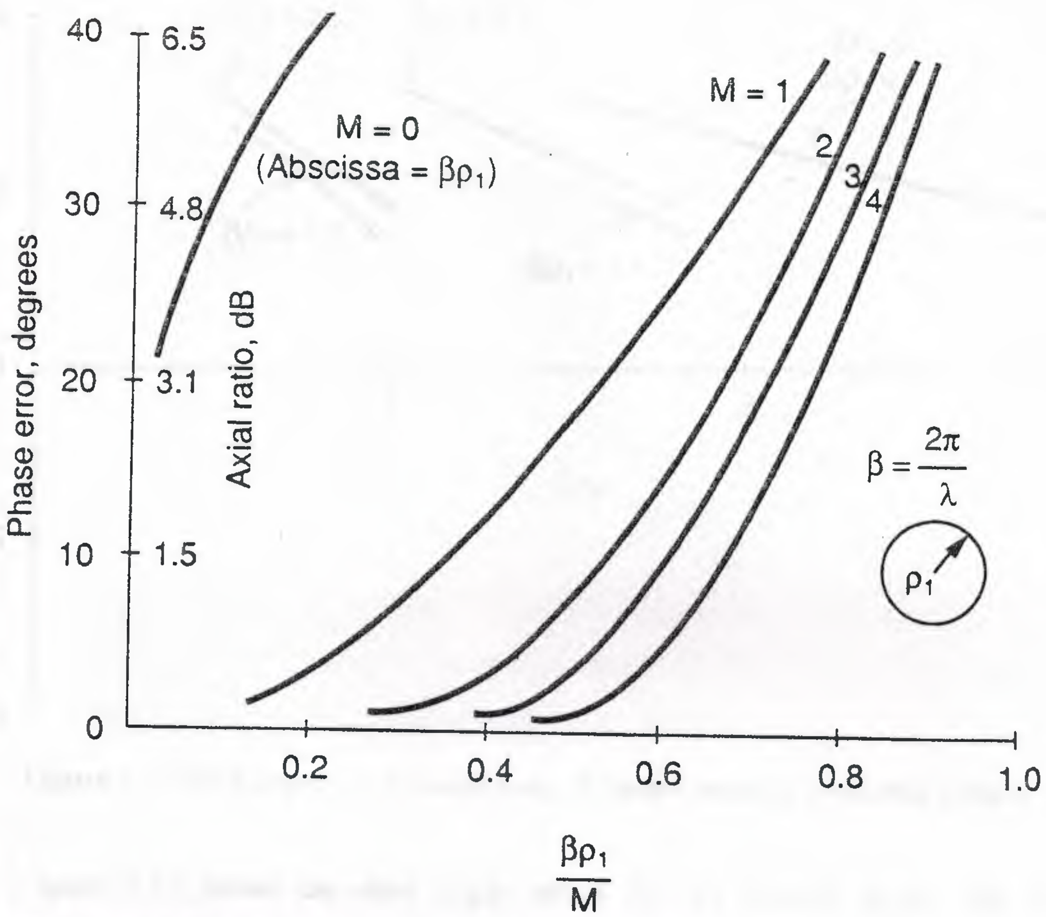


Figure 2.12 Phase error and axial ratio for a circular array of skewed dipoles around a conducting cylinder

It is apparent from the curves that for a given pole radius, we may increase the mode number M to a value such that the axial ratio is small. It is desired that the axial ratio be less than 3 dB and preferably less than 2 dB. Using linear regression curve-fitting techniques it is found that M must satisfy the following equation:

$$M \geq 0.8 - 0.13(AR) + [1.47 - 0.9(AR)] / \beta p_1 \tag{2.6}$$

where AR is the axial ratio in decibels. As an example, consider a Channel 2 antenna with a height of six wavelengths. A wind-loading analysis indicates that the mast diameter will be 20 in, for which $\beta p_1 = 0.304$. If we specify an axial ratio of 0.5 dB, we find that $M \geq 1.17$; i.e., we would have to use mode 2. If the axial-ratio requirement is relaxed to 1.6 dB, we then could use mode 1. We have now determined the number of radiating elements and the minimum mode number for the circular array of slanted dipoles. The dipole length is, in general, chosen to be about one-half wavelength because of low VSWR and low windloading requirements.

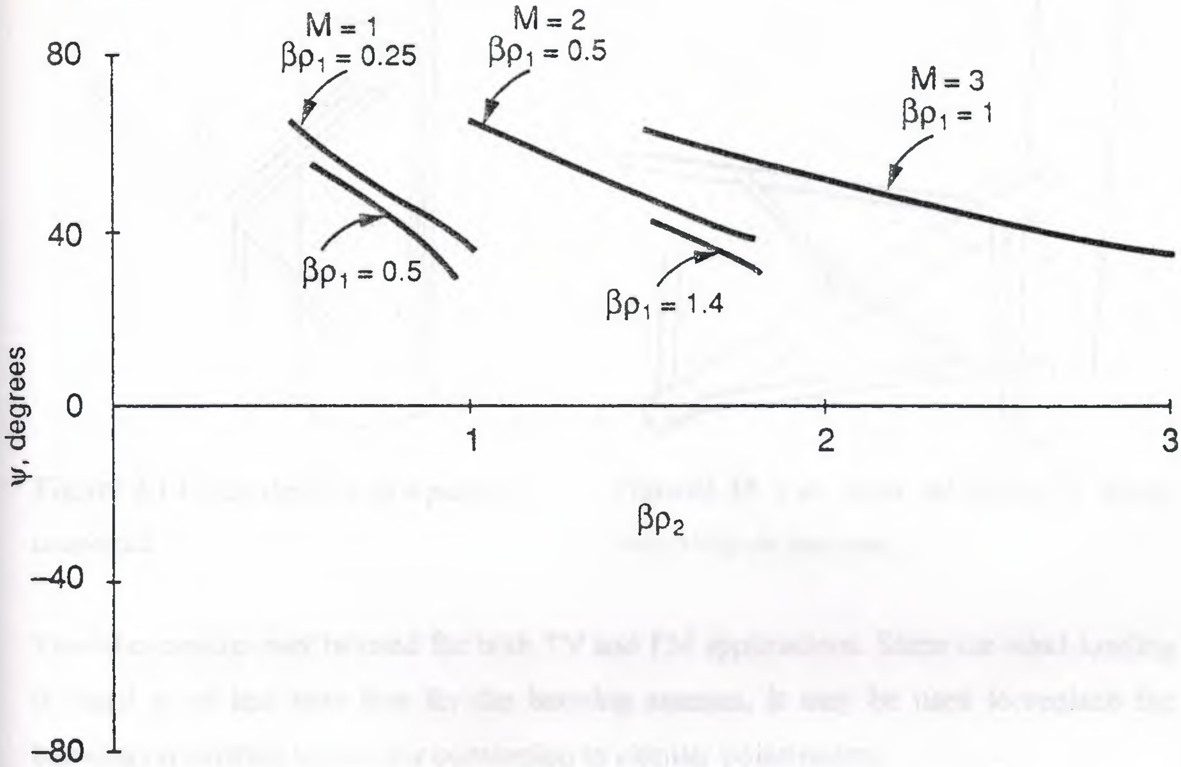


Figure 2.13 Skew angles for a circular array of dipoles around a conducting cylinder

Figure 2.13 shows the slant angle versus βp_2 for several modes and pole diameters. It is concluded from several calculations that the optimal slant angle is insensitive to the pole size as long as the pole diameter is small enough to produce small phase errors. The reason for this is as follows. Modes different than 0 produce null tangential E fields on the z-axis. The extent of the null region increases with the mode number. In other words, the field of the circular array, in the interior region of the array, is like a waveguide mode below cutoff. However, for larger pole sizes, the presence of the pole introduces a phase error that may require a change of tilt angle by 5 to 10°. Without this compensation, the axial ratio will increase by 1.6 to 3.1 dB.

Figure 2.14 shows one element of a circular array which consists of a slant half-wave dipole and a short vertical dipole that are fed and supported by a balun structure. For mode 0 and an array diameter of about one-half wavelength, only three elements are needed in the circular array to produce a circularity of ± 1.5 dB and an axial ratio less than 3 dB. The elements are fed in phase with equal power. The slant angle is approximately that given by Eq. (28-13) (without mast reflections), and the length of the vertical shunt dipole is adjusted to achieve a low axial ratio. A bandwidth greater than 10 percent may be achieved with a bay spacing of 0.8 wavelength.

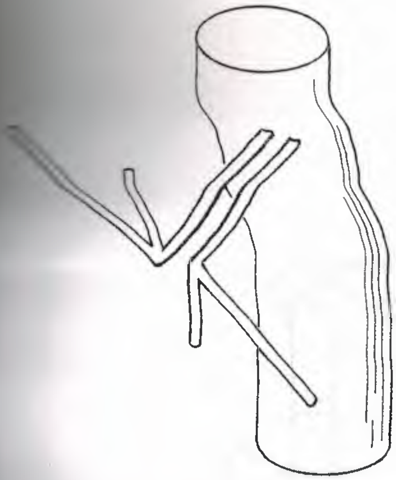


Figure 2.14 Slant dipole with a parallel connected

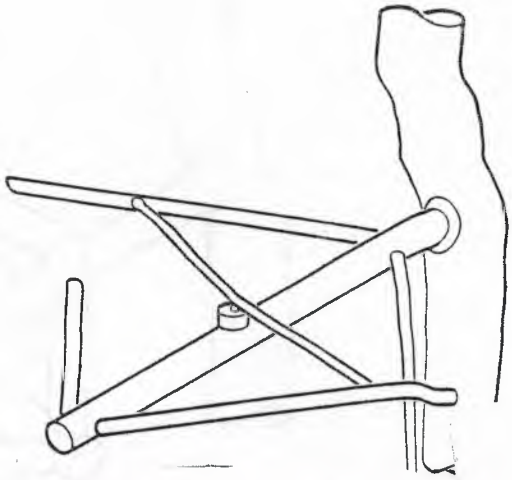


Figure2.15 Two shunt fed slanted V dipole vertical dipole antennas.

Thus the antenna may be used for both TV and FM applications. Since the wind loading is equal to or less than that for the batwing antenna, it may be used to replace the batwing on existing towers for conversion to circular polarization.

Figure 2.15 illustrates another version of this concept in which two V dipoles are supported by a horizontal mast. With an included angle of about 90°, the V dipoles perform approximately as a four-dipole circular array. One half of each dipole is shunt excited from the center of the support mast. If the dipole length is about one-half wave-length, then the current on the parasitic arms will be about the same as the current on the shunt-driven arms of the dipoles. The antenna is matched by adjusting the positions of the shunt feeds and the dipole lengths.

Alternatively, one-half of each dipole may be series-fed, as illustrated in Fig. 2.16, in which two dipoles are supported in a T arrangement. The internal coaxial feeds are connected to gaps in the monopoles. Since the impedance bandwidth is on the order of 1 percent or less, these antennas are most useful for FM applications. A multiplicity of bays with wavelength spacing may be fed and supported by a vertical transmission line. The array is usually supported on the side of a mast or tower. Reflections from the support distort the azimuth patterns, especially for vertical polarization, and degrade the axial ratio. Parasitic dipoles may be added to reduce these effects. A circular array of four-curved dipoles is shown in Fig. 2.17. The dipoles form a short section of a four-arm helix antenna and are approximately onehalf wavelength long.

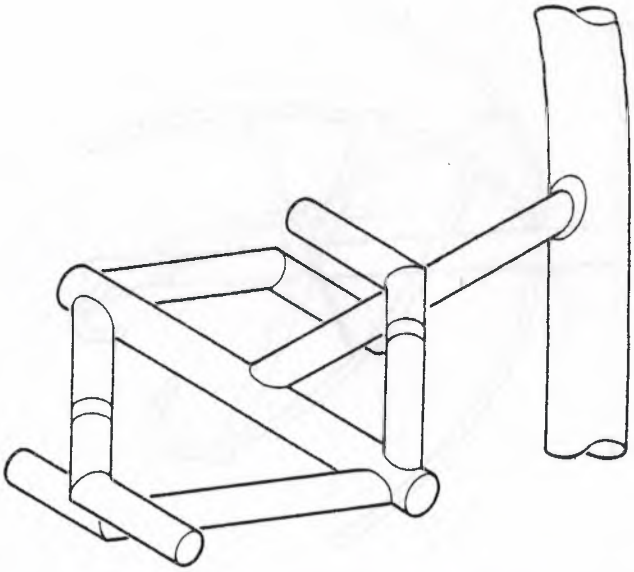


Figure 2.16 Series-fed slanted dipoles.

The array circumference is approximately one wavelength. Thus the overlap of the dipoles provides approximately the equivalent of a constant circular current distribution for both the horizontal and vertical components. The four dipoles are shunt-fed asymmetrically by four rods emanating from the center of the array. The rods are connected to the center conductor of a coaxial feed enclosed in the horizontal support structure. The same approach may be used for circular arrays of two or three curved dipoles when the array circumference is approximately one-half and three-fourths of a wavelength, respectively. The pattern circularity in free space is ± 1 dB, and the axial ratio is about 3 dB. The support mast or tower degrades these values by several decibels. The power rating and bandwidth increase with the number of dipoles in the array. An 11 percent bandwidth has been achieved for a four-element FM array with 2-in-diameter arms.

A single dipole may be bent in the form of a one-turn helical antenna to produce circular polarization. It may be fed by a slotted coaxial-line balun or by inductive loop coupling. A disadvantage is that it radiates up and down so that bay spacings of less than one wavelength must be used. For two or more dipoles in each bay there are nulls up and down for modes other than mode 1.

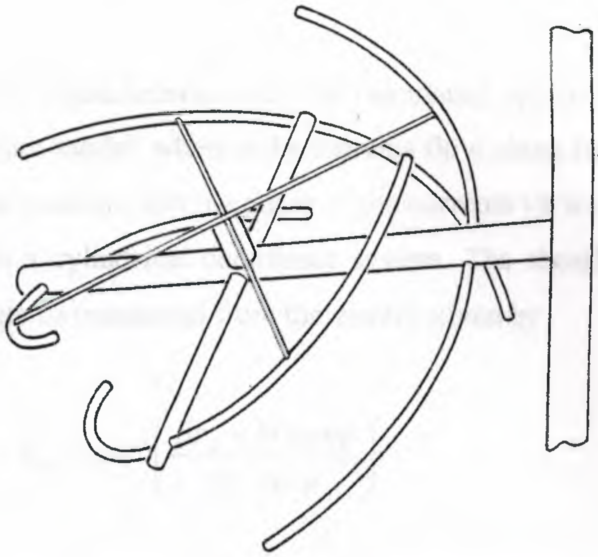


Figure 2.17 Four shunt-fed helical-type dipoles.

2.4. Helical Antennas

The multiarm helix is a versatile antenna for radiating circularly polarized waves. Figure 2.18 shows a three-arm helix with a pitch angle ψ and radius p_2 wrapped around a conducting cylinder of radius p_1, which forms the support for the antenna and allows space for a transmission-line feed network for several bays of helices. For broadside radiation, the turn length of an arm is equal to *M* wavelengths, where *M* is an integer and defined as the mode number. For an *N-arm* helix, the arms are fed with equal powers and a phase progression of $360M/N°$ such that the currents in the arms along a directrix of the helical cylinder are in phase. To obtain a low axial ratio and satisfactory radiation patterns, the number of arms *N* should be larger than the mode number M by a factor in the range of 1.5 to 2.0.

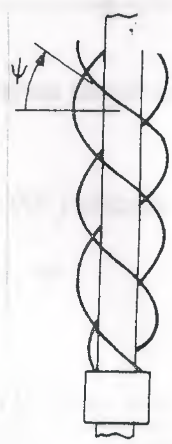


Figure 2.18 Three-arm helical antenna.

The polarization characteristics may be calculated approximately by using a continuous-current sheath model, wherein the currents flow along helical curves, have a free-space propagation constant, and the phase of the currents varies as $M\Phi$, where Φ is the azimuthal angle in a cylindrical coordinate system. The sheath current radiates a conical beam at an angle θ_0 (measured from the zenith) given by

$$\theta_0 = \cos^{-1}\left(\frac{\beta p_2 + M\cos\psi}{\beta p_2 \sin\psi}\right) \tag{2.7}$$

where β is the propagation constant along the directrix and is approximated by the free-space value of $2\pi / \lambda$.

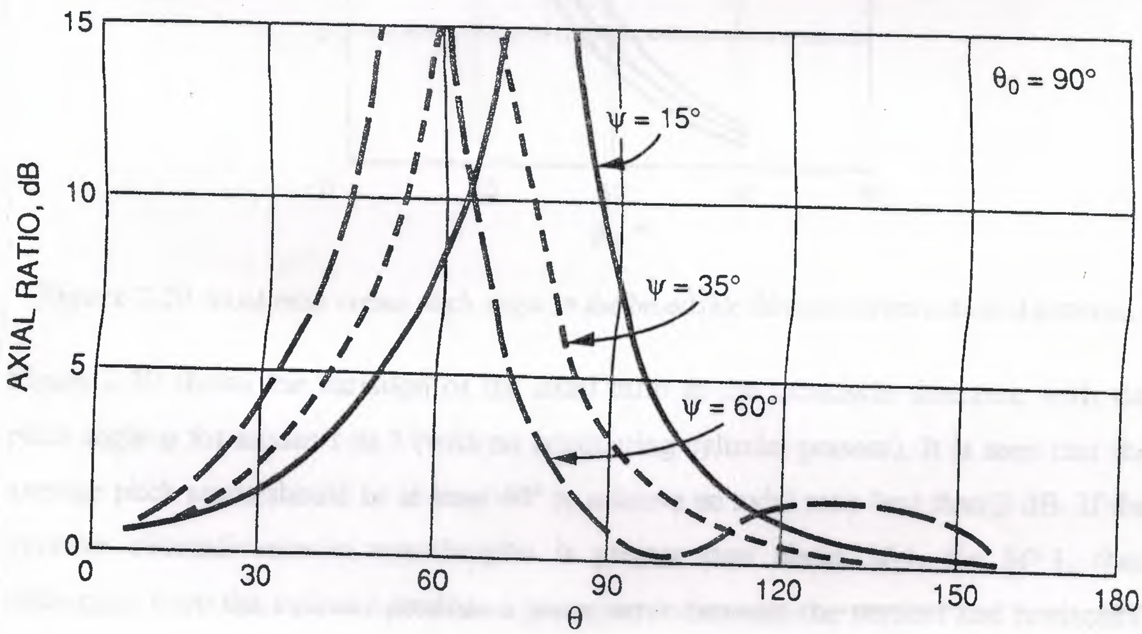


Figure 2.19 Axial ratio versus elevation angle for a helical antenna.

For broadside radiation $(\theta_0 = 90^0)$without the conducting cylinder present, the ratio E_θ / E_ϕ, is given, approximately, by

$$\frac{E_\theta}{E_\phi} \approx J\frac{J_M[\beta p_2 \sin(\theta)]}{J_M[\beta p_2 \sin(\theta)]}\left(\frac{\sin(\psi)\sin^2(\theta)-\cos(\theta)}{\sin(\theta)\cos(\psi)}\right) \tag{2.8}$$

where J_M *is* a Bessel function of the first kind and the prime represents the derivative with respect to the argument. Figure 2.19 shows the variation of axial ratio with elevation angle for three values of the pitch angle ψ and for $M = 2$. The axial ratio is rather insensitive to the mode number. The left-hand helix of Fig. 2.18 radiates left hand CP toward the zenith, right-hand CP toward the nadir, and horizontal polarization at an elevation angle of approximately ψ.

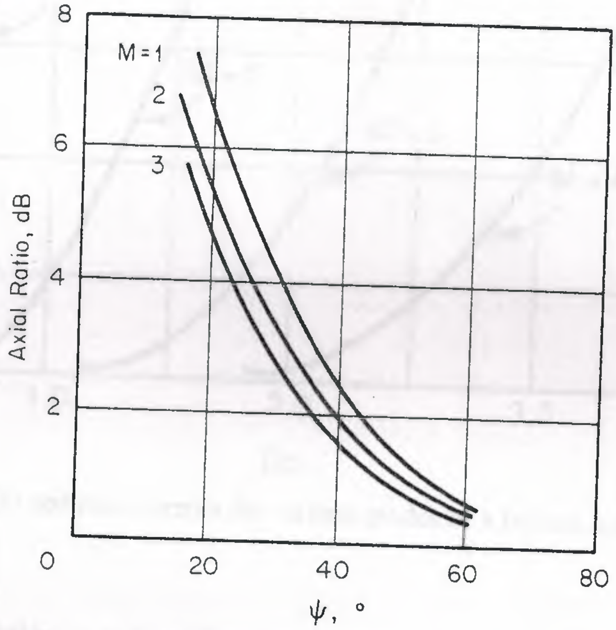


Figure 2.20 Axial ratio versus pitch angle in the broadside direction from a helical antenna.

Figure 2.20 shows the variation of the axial ratio in the broadside direction with the pitch angle ψ for modes 1 to 3 (with no conducting cylinder present). It is seen that the average pitch angle should be at least 40° to achieve an axial ratio less than 3 dB. If the cylinder circumference in wavelengths is greater than about M-1, for M>1, then reflections from the cylinder produce a phase error between the vertical and horizontal polarizations which degrades the axial ratio by more than that shown in these figures.

Figure 2.21 shows the variation of the phase error and axial ratio (assuming 0-dB axial ratio without the cylindrical conductor) versus βp_1 for modes 1 through 6. Assuming a maximum phase error of 5°, the maximum diameter of the cylinder may be calculated from these curves. The spacing of the helical wires from the cylinder is a function of the helix circumference, which is given by

$$\beta p_2 = M\cos(\psi) \tag{2.9}$$

Assuming ψ= 40° and the 5° phase error, it is deduced that the spacing between the helical wires and the core is on the order of 0.1 wavelength. For mode 1, the circumference of the cylinder should be less than one-half wavelength.

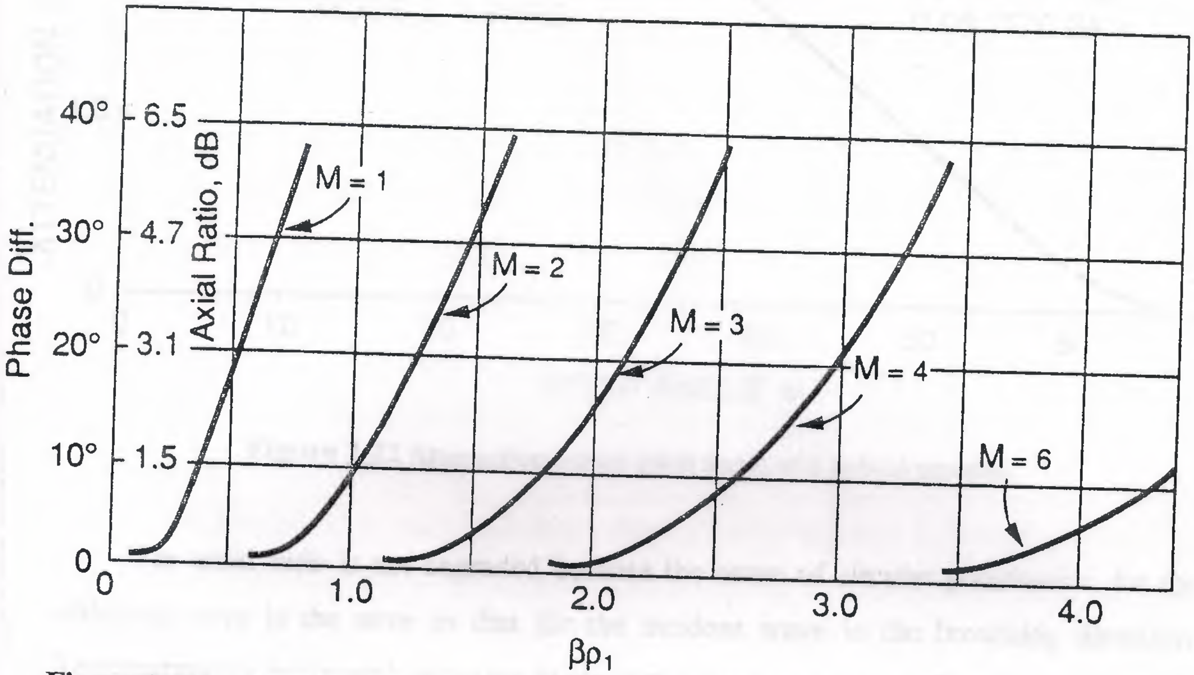


Figure 2.21 Axial ratio and phase errors for various modes of a helical antenna with a conducting cylinder.

The uniform helix is a traveling-wave antenna with an exponential attenuation rate, which is a complex function of *M, N,* ψ, cylinder diameter, and arm diameter. Figure 2.22 shows the variation of attenuation per wavelength along the axis of the helix versus the pitch angle for a mode 3 four-arm helix. These experimental results hold for cylinder diameters of 0.08 to 0.24 wavelength and for a small wire diameter. Similar results are obtained for other modes and numbers of wires. The attenuation also may be computed with a method-of-moments wire-antenna computer program wherein the cylinder is approximated by *N* axial wires for an *N-wire* helix. The attenuation increases with arm diameter and decreases with pitch angle. Attenuation rates of up to 6 dB per axial wavelength may be achieved by using wide strips or larger-diameter helix arms. To approximate a uniformly illuminated aperture, the pitch angle may be varied along the aperture (keeping the turn length constant) which leads to a spiral-type structure. If *2M/N is* not an integer, the reflected wave from the end of the helix will radiate a beam in an elevation direction other than broadside. If this is not the case, the reflected wave will radiate a broadside beam, which produces scallops in the azimuth pattern with 2*M* lobes.

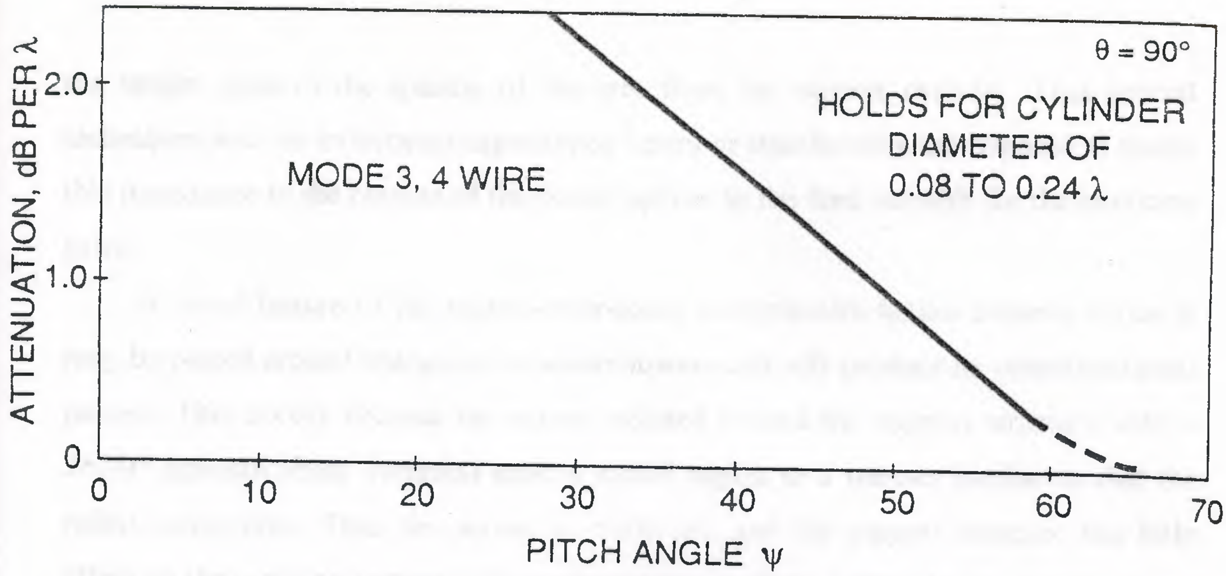


Figure 2.22 Attenuation versus pitch angle of a helical antenna.

The axial ratio is not degraded because the sense of circular polarization for the reflected wave is the same as that for the incident wave in the broadside direction. Terminating the helix with radiating loads or resistors may reduce this effect. The helix is usually designed so that the one-way attenuation is about 15 dB.

Because the helix is a traveling-wave antenna, the impedance bandwidth is large, especially *if 2M/N is* not an integer, since reflection cancellation occurs at the input to the feed network. However, the pattern bandwidth is limited by beam scan with frequency because it is equivalent to an end-fed array. For desirable pitch angles, the beam of a helix bay scans about 1 ° for a 1- percent frequency change. For Channels 2 to 6, this limits the bay length to about two to three wavelengths. Thus two or more bays are generally used. A three-arm mode 1 helix may be used for these channels and has less wind loading than a horizontally polarized batwing antenna. In order to obtain large attenuation, each arm usually consists of a wide strip or two widely spaced rods, which have crossbar connectors, spaced at intervals of less than a quarter of a wavelength.

For Channels 7 to 13, bay lengths of six wavelengths may be used. Three- or four-arm mode 2 helices are used. In the UHF band, bay lengths may be in the range of 16 wavelengths and mode numbers of 5 or more are used, with the number of arms being greater than the mode number.

Because of their symmetry, helical and spiral antennas have an excellent omni directional pattern, with a circularity of less than ± 1 dB. The axial ratio is about 2 dB for the low VHF channels and even less for the other channels. The arms of a helical antenna have characteristic impedance similar to that of a rod over a ground plane, with the height equal to the spacing of the arm from the support cylinder. Thus special techniques such as inductance-capacitance tuners or transformers are required to match this impedance to the outputs of the power splitter in the feed network for the multiarm helix.

A novel feature of the higher-order-mode multiple-arm helical antenna is that it may be placed around triangular or square towers and still produce an omnidirectional pattern. This occurs because the waves radiated toward the support structure with a $360M°$ azimuth phase variation enter a cutoff region in a manner similar to that for radial waveguides. Thus the waves are reflected, and the support structure has little effect on the radiation pattern if the mode number is about 5 times larger than the tower diameter in wavelengths.

The design of a multifilar helix antenna to produce circularly polarized omni directional radiation for TV or FM broadcast application makes use of the preceding information and usually starts by determination of the mode number. The mode number depends on the diameter of the support mast, which is determined from the antenna height (or gain) and environmental considerations such as wind or ice loading. The mode number is chosen such that the helix circumference is about one-half wavelength greater than the support circumference for the assumed pitch angle.

Figure 2.23 A three-arm mode 2 helical antenna

The number of bays should be minimized in order to minimize the cost of the feed network. The length of the single bay is limited by the amount of beam scan with frequency. The amount of beam scan depends on the pitch angle and the design bandwidth. The total beam scan over the channel may be determined for a given pitch angle. The maximum bay length is determined from the specified antenna gain variation. The number of arms *N* usually is set equal to 3 and 4 for modes 2 and 3, respectively. For the higher-order modes, *N* may be less than *M*.

Figure 2.23 is a picture of a three-layer helical antenna, which was designed for multichannel FM broadcast. This antenna is rated at 160 kW for simultaneous broadcast of six FM stations at Healy Heights, Portland, Oregon. The bay length was restricted to about three wavelengths for beam stability and to satisfy the power-capacity requirement. Excellent attenuation along the helix was achieved by using multiple conductors in each arm of the helix. A definite advantage of this antenna over panel arrays is the simplicity of the feed system. In this antenna, the feed system consisted of four three-way power dividers and the coaxial lines that fed the arms, which is simple w hen compared with the more elaborate feed system of a panel antenna with the same gain.

2.5. Ring-Panel Antennas

The ring-panel antenna consists of a multiplicity of ring radiators fed in series by a transmission line. Figure 2.24 shows two circular rings formed by strips over a panel and connected by rods over the panel which provide simple, low-radiation transmission lines. The ring circumference is approximately one wavelength, as is the spacing between rings. The antenna is designed so that the characteristic impedance of the transmission-line rod over ground is the same as the strips over ground. A practical value of the characteristic impedance is 140 Ω. By using a resistive termination on the last ring and/or special tuning techniques, it is possible to achieve a traveling-wave type of antenna.

A traveling wave on a ring of one-wavelength circumference, as shown in Fig. 2.24, radiates right-hand CP. The ring is equivalent to four quarter-wave dipoles placed on a square with -90° progressive phasing. The azimuth beamwidth for horizontal polarization is usually about 10° less than that for vertical polarization. The beamwidth may be equalized and changed by means of parasitic elements such as monopoles on each side of the rings and/or dipoles in front of the rings. A cavity is not required.

Figure 2.25 shows the variation of attenuation through a ring versus its height above the panel for two strip widths *W*. The radiation from, or the attenuation through, the traveling-wave ring increases with the height of the ring above the panel and decreases with the width of the strip. These theoretical results were obtained with a method-of-moments computer program for a single ring with a matched resistive termination at the ground plane. A rod with a diameter equal to $W/2$ was used to simulate the strip width. The computer program does not give accurate results for large strip widths. Experimental results have shown that attenuation's of 6 to 8 dB for $H/\lambda=0.2$ may be obtained with larger strip widths. The axial ratio of the ring radiator increases with the ring height from a small value to 2 to 4 dB for $H/\lambda= 0.2$. Because of this attenuation, it is necessary to increase the height of the rings as one progress from the feed point in order to approximate a uniform array.

Since the distance along the transmission line and ring between similar points on adjacent rings is two wavelengths for broadside radiation, the beam direction will scan 1.15° for a 1 percent change in frequency. This limits the number of end-fed rings to 3 for Channel 2 and about 10 for the UHF channels.

The axial ratio may be reduced to a very low level by introducing reflections on the transmission-line rods, which radiate left-hand CP waves. The size and position of the reflecting devices to cancel undesired left-hand CP from other parts of the antenna may control the magnitude and phase of the reflections.

There is undesired radiation from the rods, which connect the ring to the transmission line over the panel. This may be reduced by using a small spacing between the rods and by adjusting the ring diameter so that the distance between the midpoints of the rods, as measured through the ring, is one wavelength. This ensures that the two currents at the rod midpoints are 180° out of phase, which reduces the radiation from the rods.

Several panels may be stacked vertically to achieve the desired gain and beamwidth. Beam tilt and control of the amplitude and phase of the radiation from each ring and by the height of the ring and the transmission-line rod length may achieve null fill, respectively.

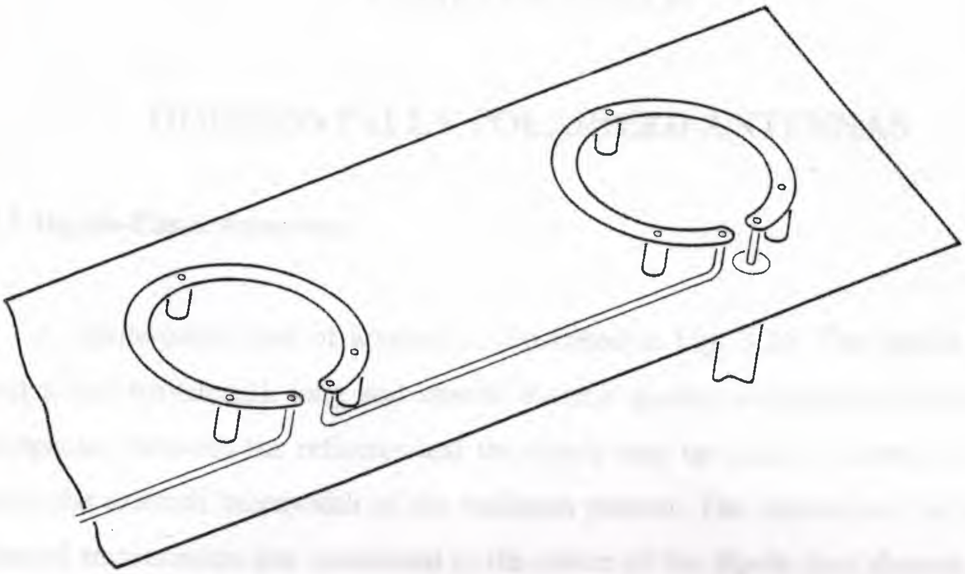


Figure 2.24 Two elements of a series-fed travelling-wave ring-panel antenna.

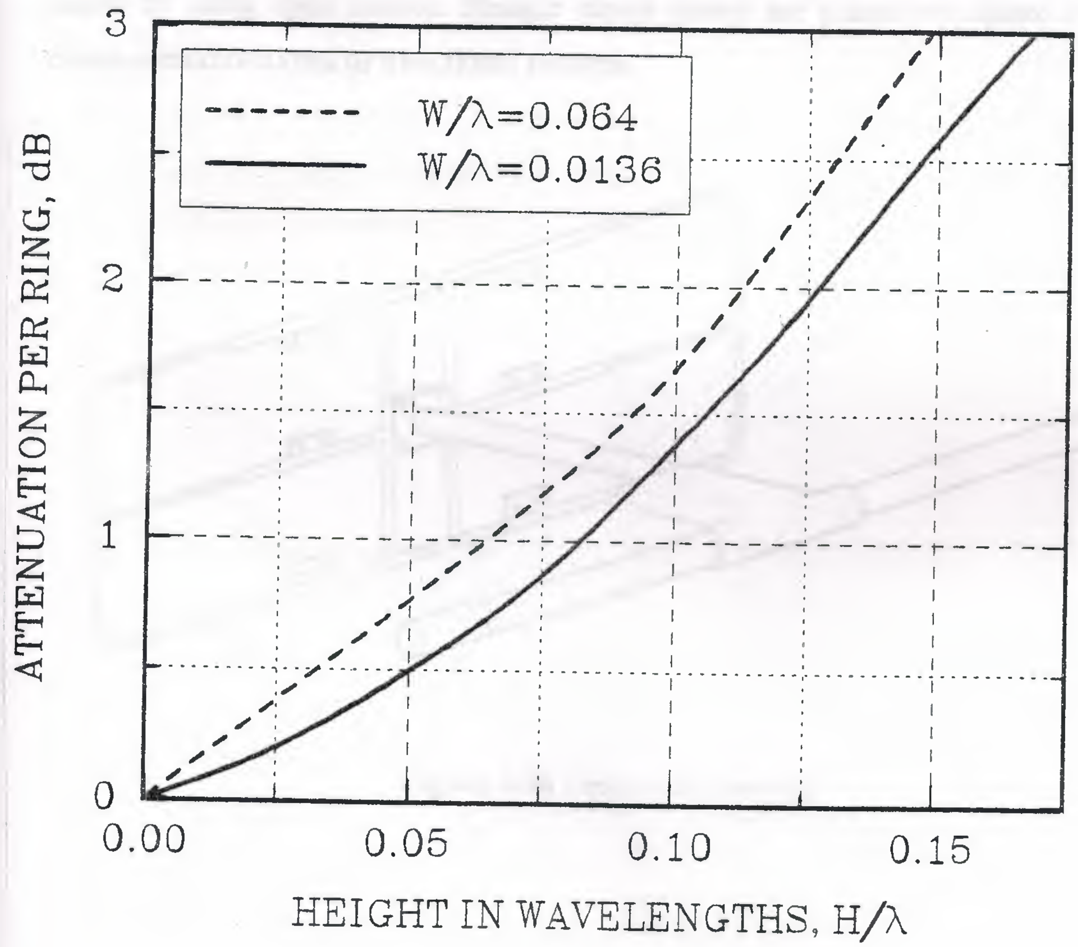


Figure 2.25 Attenuation versus height in wavelength

CHAPTER THREE

HORIZONTALLY POLARIZED ANTENNAS

3.1.1 Dipole-Panel Antennas

A dipole-panel type of antenna is illustrated in Fig. 3.26. The dipole is usually about a half wavelength long and spaced about a quarter wavelength from the panel. The spacing between the reflector and the dipole may be used to control, to a limited extent, the azimuth beamwidth of the radiation pattern. The dipole may be fed from a balanced transmission line connected to the center of the dipole (not shown) or from a coaxial line entering one of the support arms and extending to the central feed gap, forming a balun. A bandwidth of 10 percent may be achieved with a compensated balun and/or by using open sleeves. Straight dipole panels are placed on square towers to obtain omnidirectional or directional patterns.

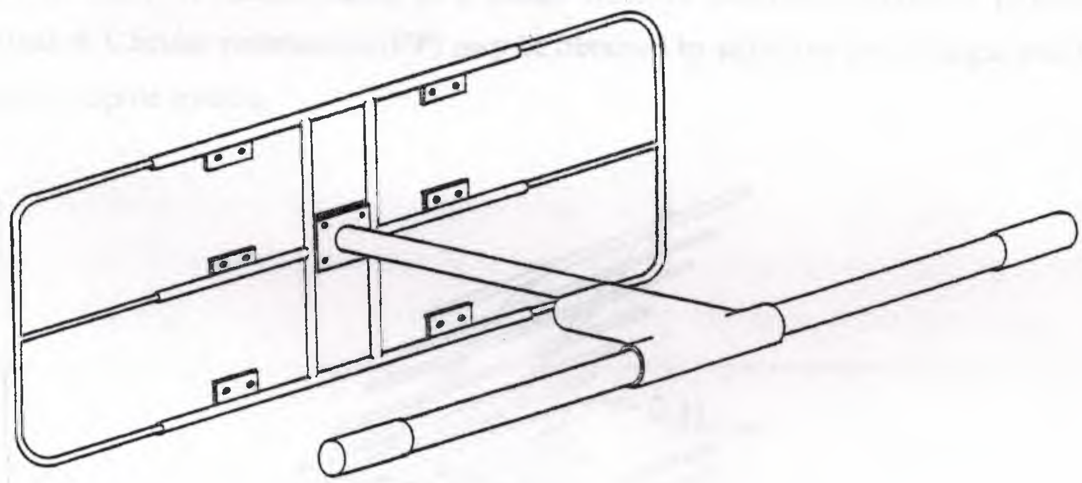


Figure 3.26 Dipole-panel antenna.

3.1.2 Corner-Reflector Antennas

Four corner-reflector antennas with a 90°-apex angle, as shown in Fig. 3.27, may be placed around a square tower to obtain horizontally polarized omnidirectional patterns. For the dimensions shown, the azimuth pattern has a 6-dB beamwidth of 90°. When four such corner reflectors are mounted on a square tower with a face dimension of 0.7 wavelength, the resulting omnidirectional pattern has a circularity of ± 1.5 dB. The gain of the corner-reflector antenna shown in Fig. 3.27 is 9 dBd (gain relative to a dipole), which is about 3.5 dB higher than the gain of a dipole panel. The higher gain is an advantage for the comer-reflector antenna when compared with a dipole-panel antenna. For a specified gain, the required number of comer reflectors is less than the required number of panel antennas; this results in a simpler feed system. The VSWR bandwidth of the comer-reflector antenna is typically 10 percent. By using sleeved dipoles and compensated baluns, the bandwidth can be extended to over 20 percent. The finite sizes of reflectors affect the gain, the front-to-back ratio of the field strength, as well as beam shape. The data reported by Wilson, Cottony, and Ohba are useful in gaining insight into the proper design of the width and length as well as the element-to-apex distance. A slanted dipole in a comer reflector produces elliptically polarized radiation. Circular polarization (CP) may be obtained by adjusting the tilt angle and the apex-to-dipole spacing.

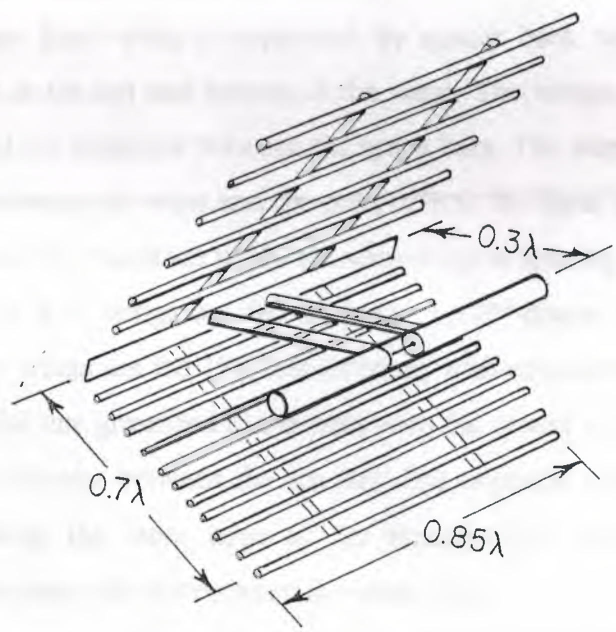


Figure 3.27 Right-angle corner-reflector antenna.

3.1.3 Batwing Antennas

The batwing antenna is the most popular horizontally polarized VHF TV antenna. A single bay of a batwing antenna is shown in Fig. 3.28.

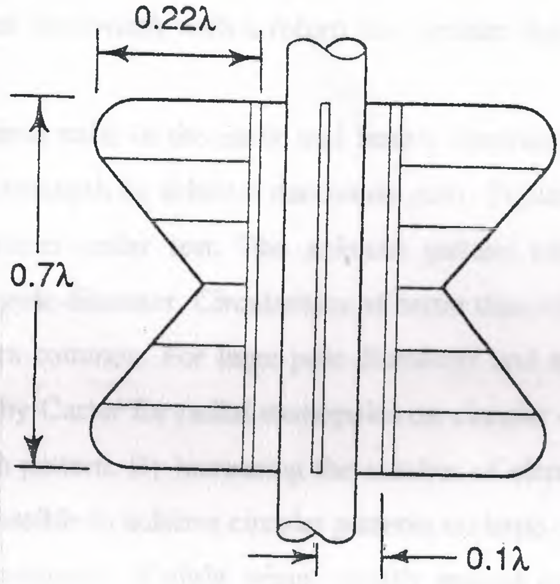


Figure 3.28 Batwing or superturnstile antenna.

Typically, a batwing antenna consists of several bays of turnstile configurations of four broadband planar wings labeled as "East-West" and "North-South." A grid of rods forms the wings. Each wing is supported by spacer bars, which are shorted to the supporting mast at the top and bottom of the wing. The wings are fed via a triangular-shaped jumper at the midpoint between the space bars. The shape of the jumper, as well as the spacing between the wing and the pole, affects the input impedance. The shape of the jumper affects the reactance while the wing-to-pole spacing affects the resistance of the element. The four wings are fed in mode 1 (90°-phase progression) via coaxial cables. Opposite wings are fed 180° out of phase with equal-length cables. One wing is fed from a coaxial line grounded to the mast with the center conductor connected to the element at the midpoint between the spacers. The opposite wing is fed from a coaxial line running along the inner edge of the element and with the center conductor connected to the mast midway between the space bars.

The 90 ° phasing of the quadrature wings may be obtained by different line lengths or from a quadrature hybrid. In the latter case, the visual and aural transmitters may be diplexed into the antenna system by insertion of the visual signal into one port of the hybrid and the aural signal into the other port. The isolation between the two ports is about the same as the return loss from the dipoles, which is greater than 26 dB. The impedance of the wing in the array varies between 72 and 79 Ω and depends on the "effective pole diameter" (diameter of the pole and the feed lines that are attached to the pole). A 20 percent bandwidth with a return loss greater than 26 dB can be achieved with relative ease.

The wings have nulls in the nadir and zenith directions. Thus the bays may be spaced by one wavelength to achieve maximum gain. Figure 3.29 shows a three-bay VHF batwing antenna under test. The azimuth pattern circularity depends on the frequency and the pole diameter. Circularities of better than ±1 dB at low VHF and ± 2 dB at high VHF are common. For large pole diameters and modes other than mode 1, equations derived by Carter for radial monopoles on circular cylinders may be used to predict the azimuth pattern. By increasing the number of elements and using a higher-order mode, it is possible to achieve circular patterns on large cylindrical poles. A mode 2 batwing array composed of eight wings equally spaced around a cylinder 12 ft in diameter on the top of the John Hancock building in Chicago produced a circular pattern within ± 0.5 dB having a bandwidth of about 15 percent at Channel 2.

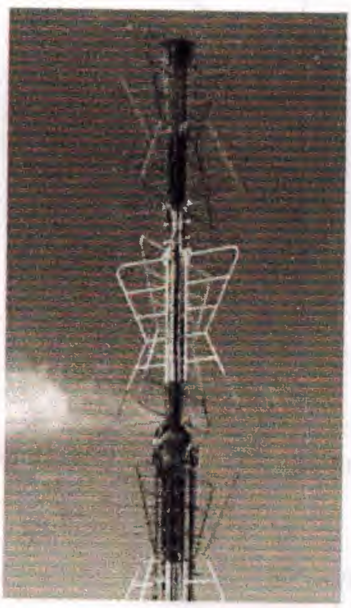


Figure 3.29 A three-bay batwing antenna.

3.1.4Zigzag-Panel Antennas

The panel-type zigzag antenna shown in Fig. 3.30 is a simple type of traveling-wave antenna, which may be placed around triangular or square towers to produce a widevariety of azimuth patterns. The antenna consists of a wire or rod that is bent at half-wavelength intervals to form the zigzag structure. This provides a broadside beam with horizontal polarization, since the radiation from the vertical components of the currents in adjacent half-wavelength segments cancels, whereas that for the horizontal components adds. The azimuth 6-dB beamwidth is about 90°, which is desirable for a square tower. It is preferable to feed the zigzag in a balanced manner at the center of one rod as shown, which eliminates radiation from the feed structure. If an unbalanced feed is used, e.g., at the bend, the feed radiation will distort the azimuth pattern.The zigzag may be designed on the basis of a leaky-wave antenna, in which the attenuation per wavelength due to radiation increases with distance from the feed point.

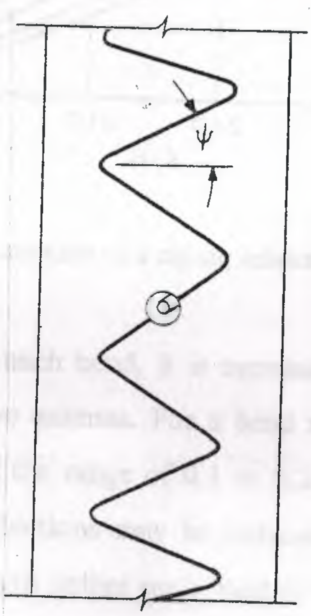


Figure 3.30 Zigzag antenna with balanced feed.

Figure 3.31 shows the attenuation per axial wavelength versus the height of the zigzag above the panel for several pitch angles. The rod diameter and band radius are 0.01 and 0.03 wavelengths, respectively. To simulate a uniform aperture distribution, the pitch angle decreases with distance from the feed point and is adjusted along with the height of the zigzag to give a one-way attenuation of about 15 dB for the current on the zigzag. It is preferable and simpler to use a constant-height zigzag. Since the beam direction of the incident and reflected waves on each half of the zigzag scan with frequency but in the opposite directions, the length of a zigzag panel is limited by the bandwidth of the channel. The average pitch angle is usually about 35°, which produces a beam scan of 1° per 1 percent change in frequency. For UHF, panel lengths of 16 wavelengths may be used.

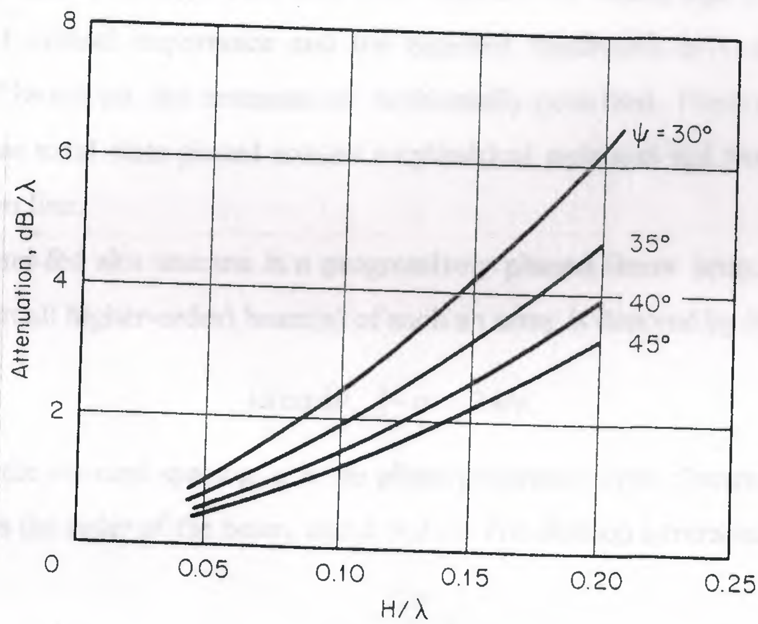


Figure 3.31 Attenuation of a zigzag antenna versus height.

Since reflections occur at each bend, it is necessary to compensate for these in order to achieve a traveling-wave antenna. For a bend radius of 0.05 wavelength, the reflection coefficient varies over the range of 0.1 to 0.2 for practical pitch angles and heights of the zigzag. These reflections may be reduced considerably by placing the support insulators 0.125 wavelength before each bend as viewed from the feed point. It is usually necessary to add shunt-capacitive tuners along the zigzag to achieve a traveling wave condition, which is required for a wide-impedance bandwidth.

A convenient method of measuring the VSWR on the zigzag rod is to use a small balanced loop probe and slide it close to and along the rod, with the rod lying in the plane of the loop. The loop measures the current on the rod.

Beam tilt may be achieved by displacing the feed from the center of the middle zigzag element or changing the lengths of the rods from the feed to the upper and lower zigzags. An alternative technique is to vary the distance between bends so as to produce a progressive phase shift, which in turn gives the desired beam tilt.

3.1.5 Slot Antennas

Because of their structural simplicity and ease of feed design, slot antennas are particularly useful for high VHF and UHF applications where high gain and low wind load are of critical importance and the required bandwidth is 1 to 3 percent. The majority of broadcast slot antennas are horizontally polarized. These are vertical arrays of half-wave axial slots placed around a cylindrical pole and fed from one end of the transmission line.

The end-fed slot antenna is a progressively phased linear array. The direction of the main (an all higher-order) beam(s) of such an array is denoted by θ_M and is given by

$$kd\cos(\theta_M) - \sigma = 2M\pi \tag{3.10}$$

where *a is* the element spacing, *a* is the phase progression per element. At is an integer that denotes the order of the beam, and $k = 2\pi/\lambda$. For slots on a transmission line,

$$\sigma = \frac{2\pi}{\lambda_g} d \tag{3.11}$$

where λ_g *is* the wavelength in the transmission line and is a function of the propagatingmode in the transmission line. Combining the two, we obtain

$$\cos(\theta_M) = M\frac{\lambda}{d} + \frac{\lambda}{\lambda_g} \tag{3.12}$$

All coaxial-type slot antennas are designed to operate in the TEM mode for which $\lambda_g = \lambda$. In order for is necessary that $M = 0$ or $M = -1$. For $M = 0$, the beam is along the axis of the line regardless of the spacing between the slots. This beam is actually suppressed because of the low level of slot radiation in this direction. For M=1 we obtain

$$\theta = \cos^{-1}\left(1 - \frac{\lambda}{d}\right) \tag{3.13}$$

Equation (3.13) shows the dependence of the beam tilt on the slot spacing. It also demonstrates the "beam-scanning" (dependence of the direction of the main beam on the frequency) characteristic of end-fed slot antennas.

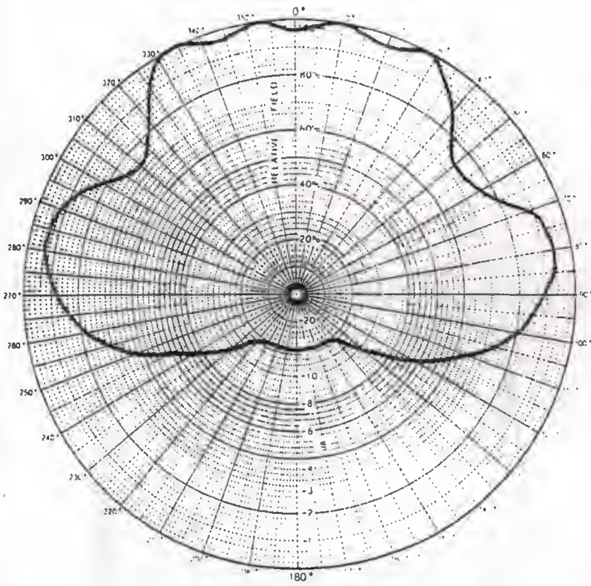
Depending on the spacing between the slots of a uniformly spaced array, the slot antenna may be either "standing-wave" or "traveling-wave." When the slots in the axial array are spaced by exactly one wavelength λ_g in the transmission line (coaxial line or waveguide) and the transmission line is fed from one end, the slots are all in phase and the main beam is broadside. The slot array is then referred to as *resonant or standing wave.* We refer to a slot antenna as *traveling wave* when the slots are spaced less than or greater than λ_g and the transmission line is terminated by a matched load. This array is progressively phased, and the main beam is tilted toward the feed end or the termination end when the spacing is less than or greater than λ_g respectively. For a narrow-beamwidth slot array at a large tower height, it usually is desirable to point the beam below the horizon; this results in a traveling-wave mode for the slotted array.

Most standing-wave slot antennas are of the coaxial type. However, owing to the limited power capacity of the coaxial lines, their application is limited to powers of 100 kW or less. Owing to the absence of an inner conductor, waveguide slot antennas have higher power capacity than coaxial ones. Circular waveguides are generally used, in traveling-wave designs, for high-power applications. Power ratings of up to 250 kW are common for the waveguide slot antennas.

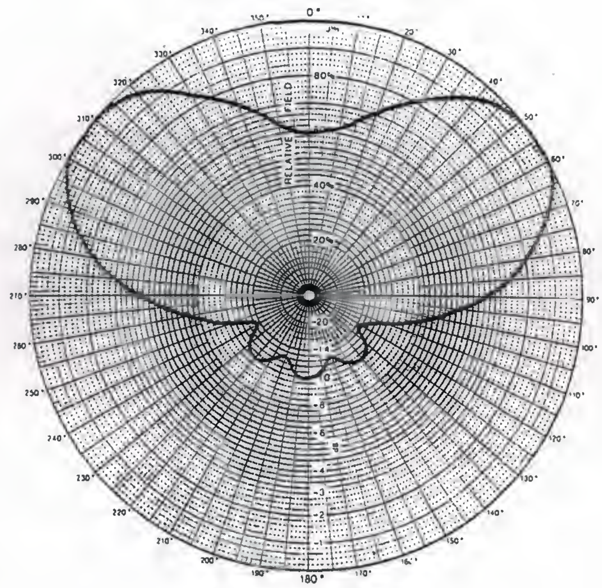
The azimuth pattern as well as the impedance of a slot on a cylindrical pole is the same whether it is an element of a standing wave or traveling-wave antenna. They both, however, depend on the normalized circumference *ka* of the cylinder, where *a* is the radius and $k = 2\pi/\lambda$. The azimuth pattern $E(\Phi)$ of a single slot (at $\Phi = 0$) on a cylinder of radius *a* is given by

$$E(\phi) = \sum_{m=0}^{\infty}\left[\frac{\varepsilon_m \exp(im\pi/2)\cos m\phi}{H_m^{(2)}(ka)}\right] \tag{3.14}$$

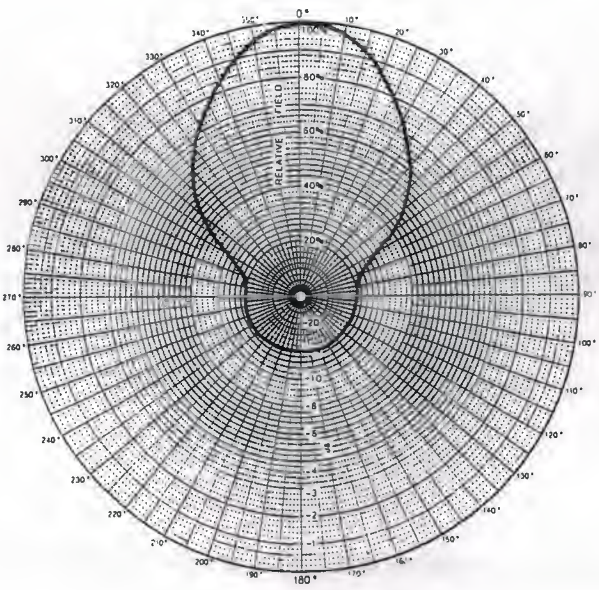
where $\varepsilon_M = 1$ for M = 1 and $\varepsilon_M = 2$ when $M > 1$, $H_m^{(2)}$ is the Hankel function of the second kind, and the prime indicates the derivative with respect to the argument. Normally, the first *3ka* terms of the series are sufficient for accurate results. For coaxial-type slot antennas the value *of ka* ranges between 1 and 1.5, whereas for the waveguide type the range is typically 2.5 to 3.5. The azimuth pattern may be controlled by either using reflectors on the sides of the slots or using combinations of slots around the pole. The pattern of a single slot is skull-shaped, and its circularity ranges from 6 to 10 dB for *ka*=1 to *ka*=3.5, respectively. Figure 3.32 shows some typical patterns that are achievable by using reflectors on the sides of the slot on a cylinder. Equation (3.14) in conjunction with Eq. (2.1) (with $R_n = 0$) may be used to predict the pattern of several slots around a cylindrical pole. Typically, two opposite slots produce a peanut-shaped pattern and three slots produce a trilobed pattern. Between four and six slots are needed to achieve a circular azimuth pattern. The pattern circularity depends on the diameter of the cylinder.



(a)



(b)



(c)

Figure 3.32 Directionalized pattern of slot antennas, (a) Directivity 1.8, *(b)* directivity 2.4, (c)directivity 4.0

Figure 3.33 shows the variation of circularity as a function of *ka*. The radiation conductance of a single resonant slot on a cylindrical pole depends on the radius of the pole. For values of *ka* between 1 and 3.5, the radiation conductance varies between 0.8 and 0.93 mΩ, respectively. The presence of reflectors increases the conductance by a factor $2\pi/\psi$, where ψ is the angle between the reflectors. The active admittance of a slot in an array around a cylinder is quite different from that of an isolated slot.

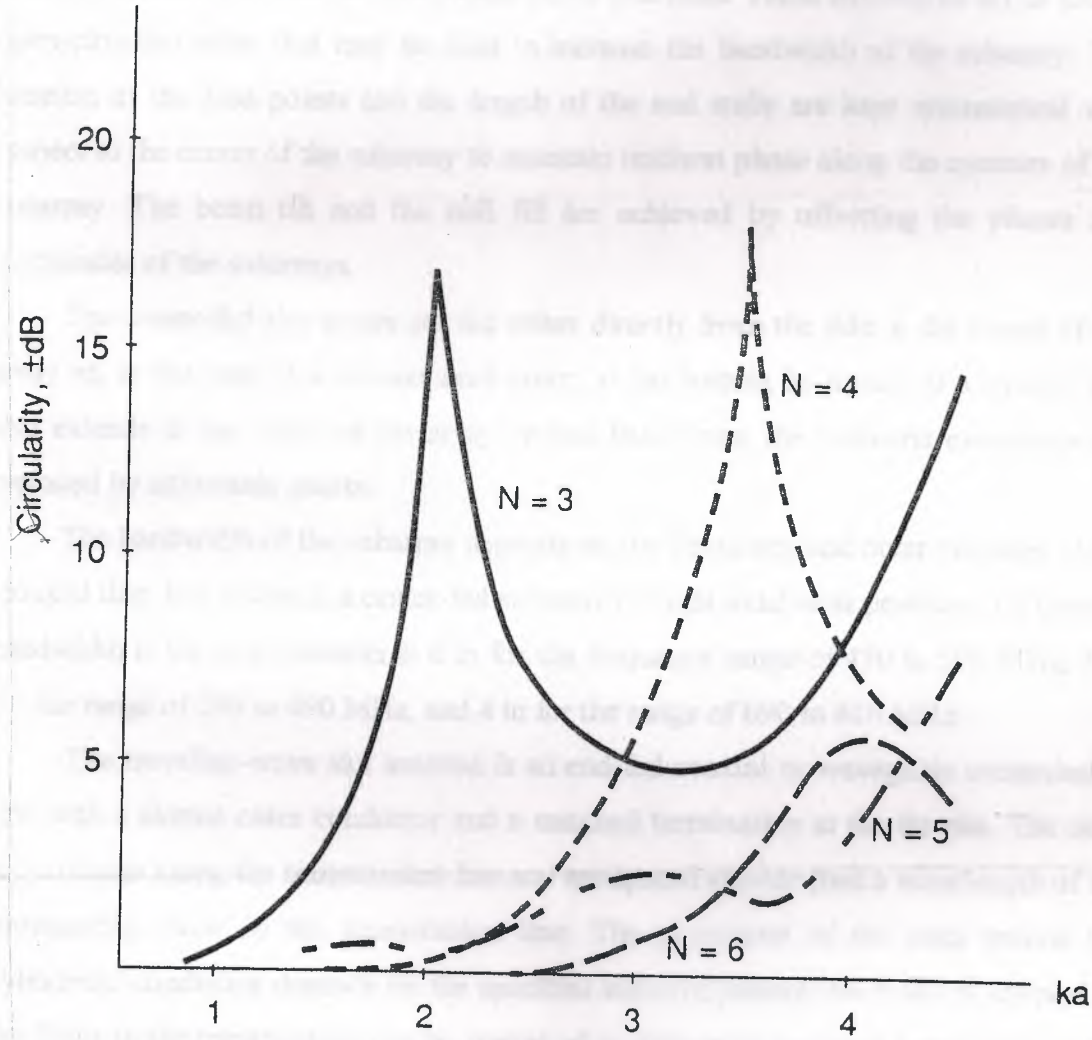


Figure 3.33 Azimuth pattern circularity versus ka.

However, since the coupling between axially displaced slots is much less than the coupling between the circumferentially displaced slots, the admittance of a single layer of the slot antenna is measured directly and the complete array is then modeled as a cascaded circuit using these empirical data. It should be pointed out that the coupling between the azimuthally displaced slots of adjacent layers is significant and should be included for accurate modeling of the array.

Low- to medium-power slot antennas, in general, are coaxial-type standing-wave slot antennas. The array usually is divided into subarrays, each fed independently. Each subarray consists of four to eight layers of slots and is fed at the center. One side of each slot, at its midpoint, is either directly connected to the inner conductor or coupled to the coaxial line by means of a capacitive probe or an inductive loop. The inner conductor is usually extended beyond the feed points of the end slots. These extensions act as shunt-open-circuited stubs that may be used to increase the bandwidth of the subarray. The location of the feed points and the length of the end stubs are kept symmetrical with respect to the center of the subarray to maintain uniform phase along the aperture of the subarray. The beam tilt and the null fill are achieved by offsetting the phases and amplitudes of the subarrays.

The center-fed slot arrays are fed either directly from the side at the center of the array or, in the case of a top-mounted array, at the bottom by means of a triaxial line that extends to the center of the array. In this latter case, the open-end extensions are replaced by adjustable shorts.

The bandwidth of the subarray depends on the frequency and outer diameter of the coaxial line. For instance, a center-fed subarray of eight axial slots produces a 5 percent bandwidth if the pole diameter is 6 in for the frequency range of 470 to 590 MHz, 5 in for the range of 590 to 690 MHz, and 4 in for the range of 690 to 810 MHz.

The traveling-wave slot antenna is an end-fed coaxial or waveguide transmission line with a slotted outer conductor and a matched termination at the far end. The slots are collinear along the transmission line and are spaced shorter than a wavelength of the propagating mode in the transmission line. The placement of the slots around the cylindrical conductor depends on the specified azimuth pattern. Each slot is coupled to the fields in the transmission line by means of an adjustable probe or a minor tilt of the slot from the axial direction. Each column may be considered as an array of progressively phased elements.

The excitation amplitude of each slot is established by an adjustment of the coupling mechanism, and the phase is established by the location of the slot along the transmission line. When coupling is the same for all slots, an equal percentage of power is transmitted from one layer of slots to the next, which results in an exponential power distribution in the array. This produces an excessive null fill in the elevation pattern, and it also may cause breakdown in the first few layers of slots in high-power applications. An approximately uniform excitation is achieved by progressively increased coupling. The last few slots at the end of the array are spaced by one wavelength and are connected directly to the inner conductor. This resonant subarray at the end, when matched serves as the matched termination of the traveling-wave array.

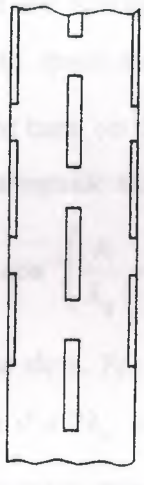


Figure 3.34 Traveling-wave slot antenna.

Figure 3.34 shows a variation of a traveling-wave slot antenna in which the slots are arranged in pairs at each layer, with the pairs separated by a quarter wavelength along the length of the antenna. Adjacent pairs occupy planes at right angles to each other. The slot pairs are fed out of phase by the coaxial line to produce a figure-eight pattern. The adjacent in-line slots are spaced slightly less than a wavelength to provide a downward beam. The quarter-wavelength separation of layers in conjunction with the space quadrature arrangement of successive layers of slots operates as a turnstile-type feed which produces an azimuth pattern with circularity of ± 1 dB. Furthermore, reflections from adjacent layers tend to cancel, which reduces the overall return loss over a wide range of frequencies.

The absence of a center conductor in cylindrical waveguides makes them attractive for high-power applications. The diameter of this antenna is almost twice the diameter of its coaxial counterpart. Consequently, it requires more slots around the pole to produce a circular pattern.

The configuration of the waveguide traveling-wave slot antenna is somewhat different from the coaxial type. The dominant mode of cylindrical waveguide is the TE mode, which is not circularly symmetrical and is not suitable for slot antennas with an omni directional azimuth pattern. This mode is used in slot antennas with skull-shaped, peanut-shaped, or other directional patterns. The next higher-order mode is the TM_{01}, which is similar to the TEM mode of the coaxial line. This mode is symmetrical and is suitable for omnidirectional patterns. For both these modes $\lambda_g > \lambda$. By choosing the inner radius a of the cylinder such that $a \cong 0.44\lambda$ for the TM_{01} mode and $a \cong 0.33\lambda$ for the TE_{11} mode, the wavelength in the waveguide will be twice that of the free space, i.e., $\lambda_g = 2\lambda$ It is then possible to space the slots by one free-space wavelength and alternately place the feed probes, or bars, on the opposite sides of the adjacent slots. The direction of the main beam of a waveguide slot antenna is given by

$$\theta = \cos^{-1}\left(\frac{\lambda}{\lambda_g} - \frac{\lambda}{2d}\right) \tag{3.15}$$

where d is the spacing between the slots. For broadside radiation, $d = \lambda$ and $\lambda_g = 2\lambda$, and the array becomes resonant. When $d \neq 2\lambda_g$ and there is a matched termination at the far end, the array becomes a traveling-wave array with the main beam at some angle other than broadside. The angle of the main beam is determined by Eq. (3.15).

The coupling of power to the slot is accomplished by an L-shaped probe that is placed at the midpoint along the side of the slot. The dimensions of the probe are adjusted to produce a gradual attenuation along the waveguide and uniform excitation of the slots.

An important factor in the design of the waveguide antenna is the single-mode operation. Maintaining the polarity of the mode in the TE), type of antenna is essential to the uniform excitation of all slots. To prevent the azimuthal rotation of the field along the waveguide, use is made of horizontal rods that are positioned transversely in the guide in the plane perpendicular to the E field. In the TM_{01} type of antenna, any deviation from circular symmetry results in excitation of the dominant TE_{11} mode, which affects excitation of the slots. This distorts the azimuth pattern, reduces the gain, and increases the level of the sidelobes.

A combination of vertically polarized radiators and axial slots may be used to achieve circular polarization. Figure 3.35 shows a design in which dipoles are placed in between the adjacent slots and are fed separately from the slots. The dipoles are placed collinear with slots along the axis or in between slots around the cylinder. The power is coupled to each dipole by means of a probe that extends inside the transmission line.

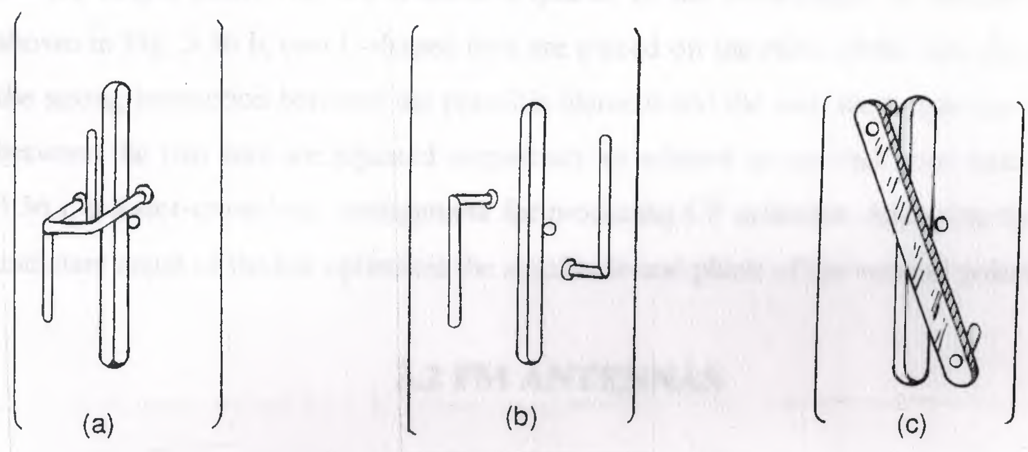


Figure 3.35 Examples of dipole slot circularly polarized antennas.

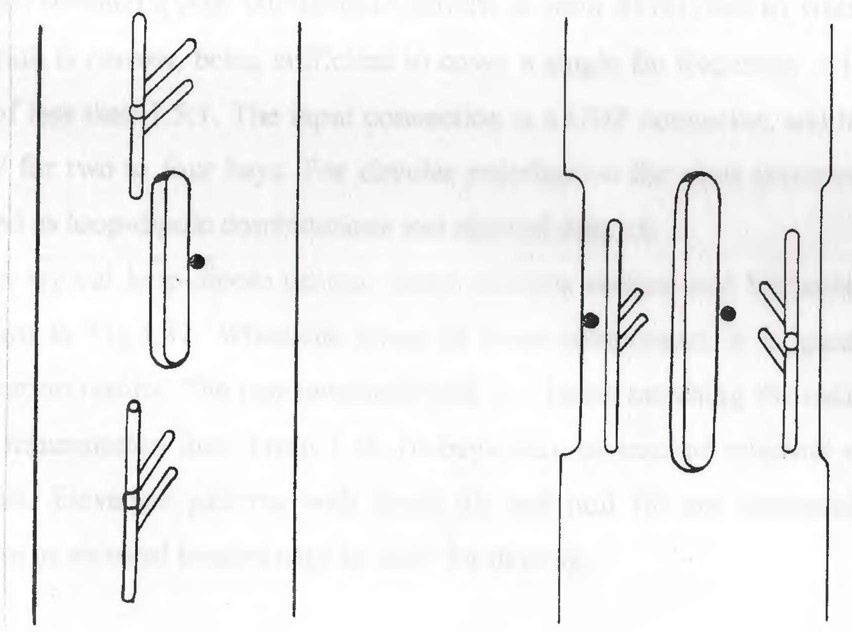


Figure 3.36 Examples of slots combined with parasitic radiators for CP radiation

Figure 3.36 *a* shows another design in which a slant half-wavelength parasitic bar is placed in front of each slot. The parasitic bar is directly connected, near its midpoint, to the sides of the slot. The height above the slot depends on the slant angle. For a vertical bar the height above the slot is about a quarter of the wavelength. In another design, shown in Fig. 3.36 b, two L-shaped bars are placed on the sides of the slot. Because of the strong interaction between the parasitic element and the slot, the height and spacing between the two bars are adjusted empirically to achieve an optimal axial ratio. Figure 3.36 c is a slot-cross-loop arrangement for producing CP radiation. Adjusting the height and slant angle of the bar optimizes the amplitude and phase of the vertical polarization.

3.2 FM ANTENNAS

Aside from panel antennas, there are three types of antennas in general use for omnidirectional side-mount fm broadcasting applications. Two these are circularly polarized, the other is horizontally polarized. The horizontally polarized antenna is primarily used for low-power educational stations and consist of one or more bays of circular loops with a spacing of one wavelength between bays. They are designed for mounting on a pole having an outside diameter of 2 to 2.5 in (5.08 to 6.35 cm). When mounted on such a pole the azimuth pattern is omni directional to within ± 3 db. The bandwidh is narrow, being sufficient to cover a single fm frequency ± 1.2 MHz with a vswr of less than 1.5:1. The input connection is a UHF connector, and the power rating 800 W for two to four bays. For circular polarization the most common types may be grouped as loop-dipole combinations and skewed dipoles.

A typical loop-dipole design, which radiates vertical and horizontal components, is shown in Fig.3.37. When the phase of these components is in quadriture, circular polarization results. The rear terminal block is a balun matching the radiator impedance to the transmission line. From 1 to 16 bays may be stacked nominal one-wavelenght intervals. Elevation patterns with beam tilt and null fill are commonly used. Either radomes or elctrical heaters may be used for deicing.

Figure 3.37 Loop-dipole fm antenna

The azimuth pattern is omnidirectional for both the horizontal and vertical components. The circularity is ± 2 db in free space. When the antenna is side mounted on a tower the circularity is affected. This is common problem with most side-mount fm antennas. The antenna is narrow band, having a vswr over a single fm channel of less than 1.1:1. It is capable of handling 10 kW per bay up to the limits of the 3/8-inch (7.94-cm) EIA feed line.

A typical skewed-dipole antenna design is shown in Fig.3.38.Two skewed dipoles are fed from a common feed line at the middle. The feed point for each dipole is offset from the center. This is a design with very high power-handling capability, radiating up to 40 kW in a single bay. The design is also broadband when compared to many fm designs. Stations having a seperation of up 5 MHz may be diplexed on a single antenna. The vswr for single station operation is less than 1.07:1 for 200 kHz on either side of the carrier. The azimuth pattern is omnidirectional, having a circularity of less than ± 2 db when mounted on a 14-inch (35.56) diameter pole or a 24 inch (67) triangular tower.

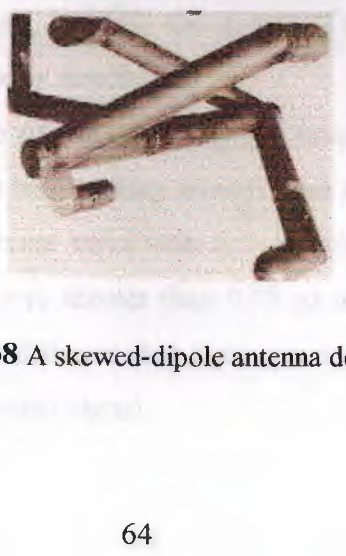


Figure 3.38 A skewed-dipole antenna design.

Another type of fm antenna is a panel antenna, which has been discussed before. The panel antenna is mounted on all side of the tower, permitting a degree of pattern control not afforded by the ordinary side-mount antenna described previously. Another motivation for the use of panel antennas is the need to multiplex several stations into a common antenna. Panels such as cavity-backed radiator are quite broadband, exhibiting an impedance bandwidth over the full fm allocation. When high power-handling capability is designed into the radiator and feed system components, it is possible to multiplex several class C stations reliably. For example, a 12-bay antenna operates with 9 class C stations, each using a 25 kW transmitter. Obviously a key component in a system of this type is the multiplexer.

3.5 MULTIPLE-ANTENNA INSTALLATIONS

The limited number of available antenna sites and financial considerations force many broadcasters to share a single site or a single tower. In some cases, several broadcasters share the same tower top in a candelabra arrangement in which antennas are mounted on the corners of a triangular support structure with a separation of 50 to 100 ft. In other cases, antennas are stacked vertically on several towers all sharing the same site and all in close proximity. A problem with such installations is that reflections from other towers and antennas in the proximity produce ripples in the pattern and/or in some situations echoes of the transmitted signal. The ripples in the pattern are sensitive to frequency. This results in a fairly strong variation, which leads to distortion of the received signal in all directions. In general, it is impractical to theoretically model the scattering from complex antenna-tower structures. However, the scattering characteristics of circular cylinders are well known and can be used, by assuming an equivalent cylinder for each tower, to study the extent of deterioration in the azimuth pattern due to the presence of nearby structures.

Echo (or ghost in the case of television signal) becomes an additional problem when the separation between the transmitting antenna and the reflecting tower exceeds 1000 ft. In general, a 2 to 3 percent echo with a 1-μs delay becomes noticeable as a ghost to a television viewer. Delays shorter than 0.25 μs are hardly noticeable. A 1-μs delay corresponds to 984 ft (the distance that light travels in 1-μs) of path difference between the original and the reflected signal.

CHAPTER FOUR

CONCLUSION

Antenna is one of the most common important and parts in the TV and FM broadcasting. There are various types of antennas with the different shapes and the properties. It is possible to categorize the antennas according to its structure. The structure of the antennas depends upon the type and the destination but in general all antennas have the following structure.

The size of antenna range from microminiature to gigantic, and it depends on the wavelength, which has proportionality with the operations frequency, and this relationship is simple and fast. There must often be some supporting structure to place the radiating element or elements in a clear location (with often is synonymous with a high location). Such devices as towers, masts, and pedestals support antennas.

Metals are the usual conducting materials of antennas. Metals of high conductivity, such as copper and aluminum (and its alloys), are naturally preferred. Brass may be used for machined parts. Magnesium is sometimes used where ultralight weight is important, usually in an alloy and with a protective coating or treatment. The steel may be used, when the strength is of primary importance, either with or without a coating or plating of copper.

The conductivity of unplated steel is adequate when it is used in the form of sheets or other large-surface-area forms (as for the surface of a paraboloidal reflector). Antenna ware is sometimes made with a steel core for strength and to minimize stretching and with a copper coating to increase the conductivity. Such wire is virtually as good a conductor as solid copper. Since the radio frequency RF currents are concentrated near the surfaces of conductors (skin effect). For this reason brass and other metals are sometimes silver plated when exceptionally high conductivity is required. For the same reason large-diameter conductors may be hollow tubes without loss of conductivity. At low radio frequencies the conductivity of large-diameter conductors may be increased, compared to a solid conductor, by interweaving strands of small-diameter insulated wires; the resulting conductor is called Litz wire. This technique is most effective below about 500KHz. At higher frequencies it is not effective because the currents tend to flow only in the outer strands.

The most fundamental properties of antennas are the following The wave polarization refers to the instantaneous component direction on a surface perpendicular to the direction of energy propagation. In the communication system only sinusoidal varying fields are ordinary used. The radiation of an antenna may be linearly, elliptically, or circularly polarized. It is obvious that in communication circuits it is essential that transmission and receiving antennas have the same polarization. Also it is apparent that the polarization properties of any antenna are an important part of its technical description (parameter of its performance). Sometimes it may be desirable to provide polarization pattern of the antenna, that is, a description of the polarization radiated as a function of the direction angles of a spherical coordinate system, although such a complete picture of the polarization is not ordinarily. The radiation pattern of an antenna is one of its most fundamental properties, and many of its performance parameters pertain to various aspects of the pattern.

Antennas have a reciprocal relationship between the processes of radiation and reception, so, it is customary to speak of the antenna pattern as radiation pattern, and a reception pattern as well because it also describes the receiving properties of the antenna. The radiation pattern describes the relative strength of the radiated field in various directions from the antenna, at a fixed or a constant distance.

In our discussion of the antenna gain the concept of an isotropic radiator or isotrope is fundamental. Essentially an isotrope is an antenna that radiates uniformly in all directions of space. This pattern is a perfect spherical surface in space; that is, if the electric intensity of the field radiated by an isotrope is measured at all point on an imaginary spherical surface with the isotrope at the center (in free space), the same value will be measured everywhere.

All antennas are limited in the range of frequency over which they will operate satisfactorily. This range is called the bandwidth of the antenna. Bandwidth is a concept that is probably familiar in other applications, sometimes by another name. For example, a television I-f amplifier must have a bandwidth of approximately 4MHz in order to pass all the frequency components of a television signal. A television-transmitting antenna must have sufficient bandwidth to receive all the channels to which the receiving set can be tuned. If an antenna were capable of operating satisfactory from a minimum frequency of 155MHz to a maximum frequency of 205MHz, its bandwidth would be 10MHz.

In many cases, the supporting structure is a triangular or square tower. Panel antennas are primarily used to control or minimize the reflections from the supporting structure. Some panel antennas are made of a single horizontal dipole or two crossed dipoles (circularly polarized panel) in front of a reflector.

The reflector can be a flat panel, a corner reflector, or a pillbox (commonly referred to as *cavity-backed).* The reflector is usually a wire grid for VHF or a solid sheet for UHF.

The FCC allowance of circularly polarized broadcast transmission has led to the introduction of a wide variety of new transmitting antennas. The antenna types usually take the form of crossed dipoles, circular arrays of slanted dipoles, helical structures, and traveling wave ring configurations.

A common technique for producing circular polarization has been to place two linear dipoles at right angles in front of a reflecting screen and to feed them with equal voltage magnitudes and with a 90°-phase difference. However, the azimuth beam width for horizontal and vertical polarization is about 60 and 120°, respectively.

Many circularly polarized FM and TV broadcast antennas are based on the concept of a circular array of slanted dipoles. The dipoles may be linear, V-shaped, curved, or of similar configuration. Each dipole radiates linear polarization, but the slant angles and diameter of the circular array are adjusted so that an omnidirectional, circularly polarized radiation pattern is obtained.

The multiarm helix is a versatile antenna for radiating circularly polarized waves. For broadside radiation, the turn length of an arm is equal to M wavelengths, where M is an integer and defined as the mode number. For an *N-arm* helix, the arms are fed with equal powers and a phase progression of $360M/N°$ such that the currents in the arms along a directrix of the helical cylinder are in phase. To obtain a low axial ratio and satisfactory radiation patterns, the number of arms N should be larger than the mode number M by a factor in the range of 1.5 to 2.0.

The ring-panel antenna consists of a multiplicity of ring radiators fed in series by a transmission line. The ring circumference is approximately one wavelength, as is the spacing between rings. The antenna is designed so that the characteristic impedance of the transmission-line rod over ground is the same as the strips over ground. A practical value of the characteristic impedance is 140 Ω. By using a resistive termination on the last ring and/or special tuning techniques, it is possible to achieve a traveling-wave type of antenna.

In the dipole-panel antenna, the dipole is usually about a half wavelength long and spaced about a quarter wavelength from the panel. The spacing between the reflector and the dipole may be used to control, to a limited extent, the azimuth beamwidth of the radiation pattern. The dipole may be fed from a balanced transmission line connected to the center of the dipole (not shown) or from a coaxial line entering one of the support arms and extending to the central feed gap, forming a balun. A bandwidth of 10 percent may be achieved with a compensated balun and/or by using open sleeves. Straight dipole panels are placed on square towers to obtain omnidirectional or directional patterns.

Four corner-reflector antennas with a 90°-apex angle, as shown in Fig. 2.27, may be placed around a square tower to obtain horizontally polarized omnidirectional patterns. For the dimensions shown, the azimuth pattern has a 6-dB beamwidth of 90°. When four such corner reflectors are mounted on a square tower with a face dimension of 0.7 wavelength, the resulting omnidirectional pattern has a circularity of ± 1.5 dB. The gain of the corner-reflector antenna shown in Fig. 2.27 is 9 dBd (gain relative to a dipole), which is about 3.5 dB higher than the gain of a dipole panel. The higher gain is an advantage for the comer-reflector antenna when compared with a dipole-panel antenna. For a specified gain, the required number of comer reflectors is less than the required number of panel antennas; this results in a simpler feed system. The VSWR bandwidth of the comer-reflector antenna is typically 10 percent. By using sleeved dipoles and compensated baluns, the bandwidth can be extended to over 20 percent. The finite sizes of reflectors affect the gain, the front-to-back ratio of the field strength, as well as beam shape.

The batwing antenna is the most popular horizontally polarized VHF TV antenna. Typically, a batwing antenna consists of several bays of turnstile configurations of four broadband planar wings labeled as "East-West" and "North-South." A grid of rods forms the wings. Each wing is supported by spacer bars, which are shorted to the supporting mast at the top and bottom of the wing. The four wings are fed in mode 1 (90°-phase progression) via coaxial cables. Opposite wings are fed 180° out of phase with equal-length cables. One wing is fed from a coaxial line grounded to the mast with the center conductor connected to the element at the midpoint between the spacers. The opposite wing is fed from a coaxial line running along the inner edge of the element and with the center conductor connected to the mast midway between the space bars.The 90 ° phasing of the quadrature wings may be obtained by different line lengths or from a quadrature hybrid. In the latter case, the visual and aural transmitters may be diplexed into the antenna system by insertion of the visual signal into one port of the hybrid and the aural signal into the other port.

The panel-type zigzag antenna shown in Fig. 2.30 is a simple type of traveling-wave antenna, which may be placed around triangular or square towers to produce a widevariety of azimuth patterns. The antenna consists of a wire or rod that is bent at half-wavelength intervals to form the zigzag structure. This provides a broadside beam with horizontal polarization The azimuth 6-dB beamwidth is about 90°, which is desirable for a square tower. It is preferable to feed the zigzag in a balanced manner at the center of one rod as shown, which eliminates radiation from the feed structure. If an unbalanced feed is used, e.g., at the bend, the feed radiation will distort the azimuth pattern.The zigzag may be designed on the basis of a leaky-wave antenna, in which the attenuation per wavelength due to radiation increases with distance from the feed point.

Slot antennas are particularly useful for high VHF and UHF applications where high gain and low wind load are of critical importance and the required bandwidth is 1 to 3 percent. The majority of broadcast slot antennas are horizontally polarized. These are vertical arrays of half-wave axial slots placed around a cylindrical pole and fed from one end of the transmission line. Most standing-wave slot antennas are of the coaxial type. However, owing to the limited power capacity of the coaxial lines, their application is limited to powers of 100 kW or less. Owing to the absence of an inner conductor, waveguide slot antennas have higher power capacity than coaxial ones. Circular waveguides are generally used, in traveling-wave designs, for high-power applications.

An important factor in the design of the waveguide antenna is the single-mode operation. Maintaining the polarity of the mode in the TE), type of antenna is essential to the uniform excitation of all slots. To prevent the azimuthal rotation of the field along the waveguide, use is made of horizontal rods that are positioned transversely in the guide in the plane perpendicular to the *E* field. In the TM_{01} type of antenna, any deviation from circular symmetry results in excitation of the dominant TE_{11} mode, which affects excitation of the slots. This distorts the azimuth pattern, reduces the gain, and increases the level of the sidelobes.

There are three types of antennas in general use for omnidirectional side-mount fm broadcasting applications. Two these are circularly polarized, the other is horizontally polarized. The horizontally polarized antenna is primarily used for low-power educational stations and consist of one or more bays of circular loops with a spacing of one wavelength between bays.

Another type of fm antenna is a panel antenna. The panel antenna is mounted on all side of the tower, permitting a degree of pattern control not afforded by the ordinary side-mount antenna described previously. Another motivation for the use of panel antennas is the need to multiplex several stations into a common antenna. Panels such as cavity-backed radiator are quite broadband, exhibiting an impedance bandwidth over the full fm allocation. When high power-handling capability is designed into the radiator and feed system components, it is possible to multiplex several class C stations reliably. For example, a 12-bay antenna operates with 9 class C stations, each using a 25 kW transmitter. Obviously a key component in a system of this type is the multiplexer.

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