RF and Microwave Wireless Systems. Kai Chang

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CHAPTER THREE

Antenna Systems

3.1 INTRODUCTION

The study of antennas is very extensive and would need several texts to cover adequately. In this chapter, however, a brief description of relevant performances and design parameters will be given for introductory purposes.

An antenna is a component that radiates and receives the RF or microwave power. It is a reciprocal device, and the same antenna can serve as a receiving or transmitting device. Antennas are structures that provide transitions between guided and free-space waves. Guided waves are confined to the boundaries of a transmission line to transport signals from one point to another [1], while free-space waves radiate unbounded. A transmission line is designed to have very little radiation loss, while the antenna is designed to have maximum radiation. The radiation occurs due to discontinuities (which cause the perturbation of fields or currents), unbalanced currents, and so on.

The antenna is a key component in any wireless system, as shown in Fig. 3.1. The RF/microwave signal is transmitted to free space through the antenna. The signal propagates in space, and a small portion is picked up by a receiving antenna. The signal will then be amplified, downconverted, and processed to recover the information.

There are many types of antennas; Fig. 3.2 gives some examples. They can be classified in different ways. Some examples are:

1. Shapes or geometries:
   a. Wire antennas: dipole, loop, helix
   b. Aperture antennas: horn, slot
   c. Printed antennas: patch, printed dipole, spiral

2. Gain:
   a. High gain: dish
   b. Medium gain: horn
   c. Low gain: dipole, loop, slot, patch
3. Beam shapes:
   a. Omnidirectional: dipole
   b. Pencil beam: dish
   c. Fan beam: array

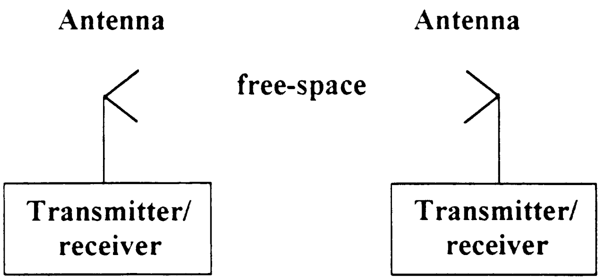


FIGURE 3.1 Typical wireless system.

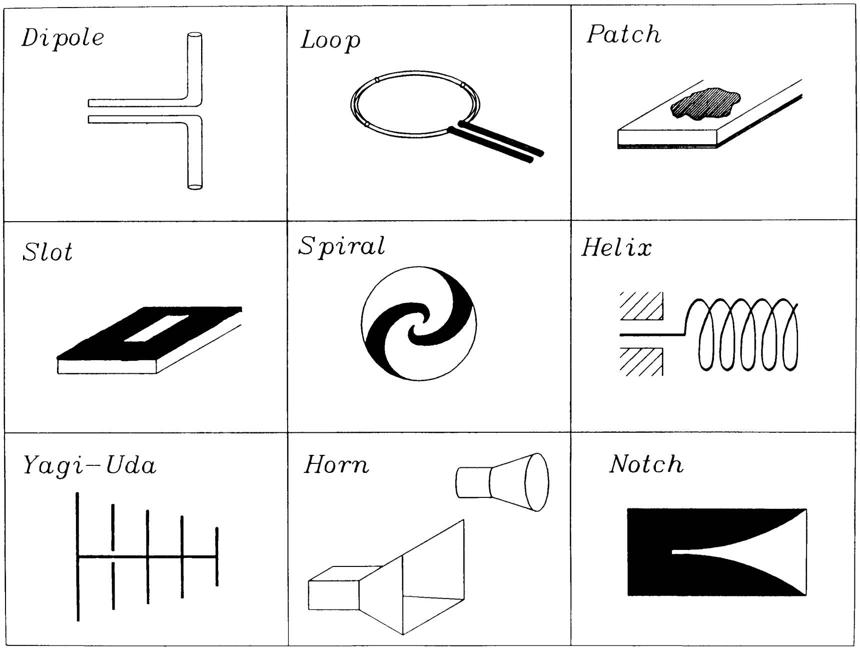


FIGURE 3.2 Various antennas [2].

4. Bandwidth:
   a. Wide band: log, spiral, helix
   b. Narrow band: patch, slot

Since antennas interface circuits to free space, they share both circuit and radiation qualities. From a circuit point of view, an antenna is merely a one-port device with an associated impedance over frequency. This chapter will describe some key antenna properties, followed by the designs of various antennas commonly used in wireless applications.

3.2 ISOTROPIC RADIATOR AND PLANE WAVES

An isotropic radiator is a theoretical point antenna that cannot be realized in practice. It radiates energy equally well in all directions, as shown in Fig. 3.3. The radiated energy will have a spherical wavefront with its power spread uniformly over the surface of a sphere. If the source transmitting power is P_t, the power density P_d in watts per square meters at a distance R from the source can be calculated by

$$P_d = \frac{P_t}{4\pi R^2} \tag{3.1}$$

Although the isotropic antenna is not practical, it is commonly used as a reference with which to compare other antennas.

At a distance far from the point source or any other antenna, the radiated spherical wave resembles a uniform plane wave in the vicinity of the receiving antenna. This can be understood from Fig. 3.4. For a large R, the wave can be approximated by a uniform plane wave. The electric and magnetic fields for plane waves in free space can be found by solving the Helmholtz equation

$$\nabla^2 \vec{E} + k_0^2 \vec{E} = 0 \tag{3.2}$$

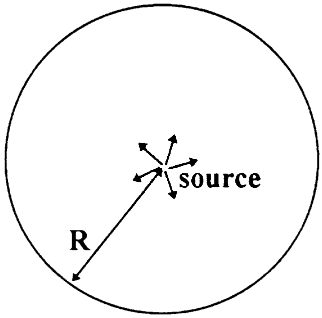


FIGURE 3.3 Isotropic radiator.

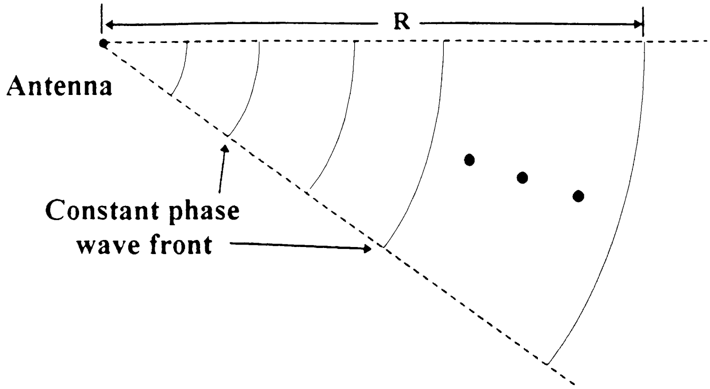


FIGURE 3.4 Radiation from an antenna.

where $k_0 = 2\pi/\lambda_0$. The solution is [1]

$$\vec{E} = \vec{E}_0 e^{-j\vec{k}_0 \cdot \vec{r}} \tag{3.3}$$

The magnetic field can be found from the electric field using the Maxwell equation, given by

$$\vec{H} = -\frac{1}{j\omega\mu_0} \nabla \times \vec{E} = \sqrt{\frac{\varepsilon_0}{\mu_0}}\, \hat{\underline{n}} \times \vec{E} \tag{3.4}$$

where μ_0 is the free-space permeability, ε_0 is the free-space permittivity, ω is the angular frequency, and k_0 is the propagation constant. Here, $\vec{k}_0 = \hat{\underline{n}} k_0$, and $\hat{\underline{n}}$ is the unit vector in the wave propagation direction, as shown in Fig. 3.5. The vector E_0 is perpendicular to the direction of the propagation, and H is perpendicular to $\vec{E}$ and $\hat{\underline{n}}$. Both $\vec{E}$ and $\vec{H}$ lie in the constant-phase plane, and the wave is a TEM wave.

The intrinsic impedance of free space is defined as

$$\eta_0 = \frac{E}{H} = \sqrt{\frac{\mu_0}{\varepsilon_0}} = 120\pi \text{ or } 377\ \Omega \tag{3.5}$$

The time-averaged power density in watts per square meters is given as

$$P_d = \left| \frac{1}{2} \vec{E} \times \vec{H}^* \right| = \frac{1}{2} \frac{E^2}{\eta_0} \tag{3.6}$$

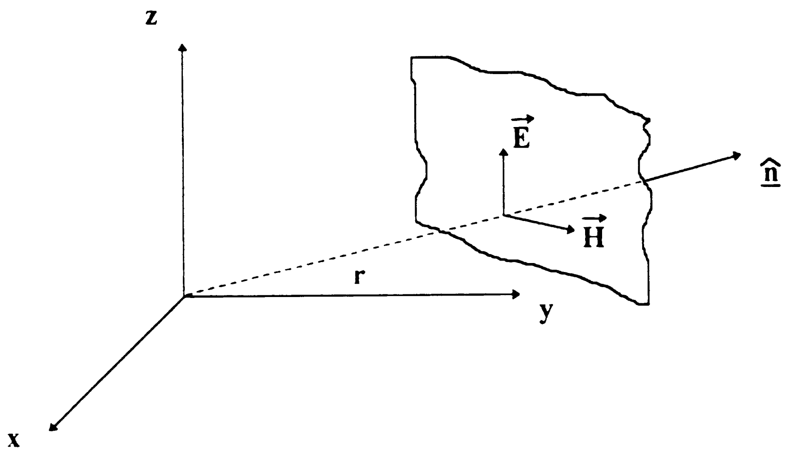


FIGURE 3.5 Plane wave.

where the asterisk denotes the complex conjugate quantity. By equating Eq. (3.1) and Eq. (3.6), one can find the electric field at a distance R from the isotropic antenna as

$$E = \frac{\sqrt{60P_t}}{R} = \sqrt{2}E_{\text{rms}} \tag{3.7}$$

where E is the peak field magnitude and E_{rms} is the root-mean-square (rms) value.

3.3 FAR-FIELD REGION

Normally, one assumes that the antenna is operated in the far-field region, and radiation patterns are measured in the far-field region where the transmitted wave of the transmitting antenna resembles a spherical wave from a point source that only locally resembles a uniform plane wave. To derive the far-field criterion for the distance R, consider the maximum antenna dimension to be D, as shown in Fig. 3.6. We have

$$\begin{aligned} R^2 &= (R - \Delta l)^2 + (\tfrac{1}{2}D)^2 \\ &= R^2 - 2R\,\Delta l + (\Delta l)^2 + (\tfrac{1}{2}D)^2 \end{aligned} \tag{3.8}$$

For $R \gg \Delta l$, Eq. (3.8) becomes

$$2R\,\Delta l \approx \tfrac{1}{4}D^2 \tag{3.9}$$

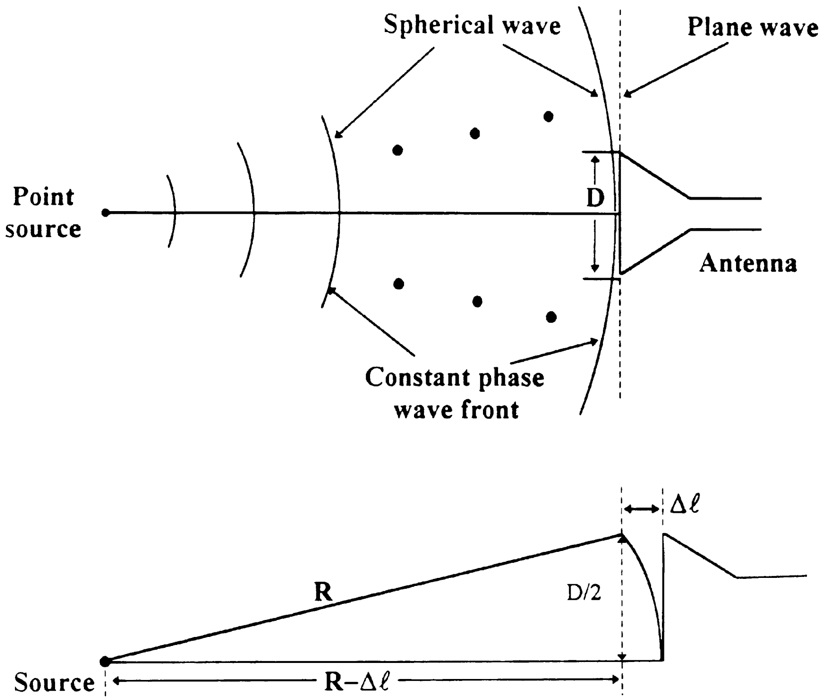


FIGURE 3.6 Configuration used for calculation of far-field region criterion.

Therefore

$$R = \frac{D^2}{8\,\Delta l} \tag{3.10}$$

If we let $\Delta l = \frac{1}{16}\lambda_0$, which is equivalent to 22.5° phase error, be the criterion for far-field operation, we have

$$R_{\text{far-field}} = \frac{2D^2}{\lambda_0} \tag{3.11}$$

where λ_0 is the free-space wavelength. The condition for far-field operation is thus given by

$$R \geq \frac{2D^2}{\lambda_0} \tag{3.12}$$

It should be noted that other criteria could also be used. For example, if $\Delta l = \frac{1}{32}\lambda_0$ or 11.25° phase error, the condition will become $R \geq 4D^2/\lambda_0$ for far-field operation.

3.4 ANTENNA ANALYSIS

To analyze the electromagnetic radiation of an antenna, one needs to work in spherical coordinates. Considering an antenna with a volume V and current $\vec{J}$ flowing in V, as shown in Fig. 3.7, the electric and magnetic fields can be found by solving the inhomogeneous Helmholtz equation [1]:

$$\nabla^2 \vec{A} + k_0^2 \vec{A} = -\mu \vec{J} \tag{3.13}$$

where $\vec{A}$ is the vector potential, defined as

$$\vec{B} = \nabla \times \vec{A} = \mu_0 \vec{H} \tag{3.14}$$

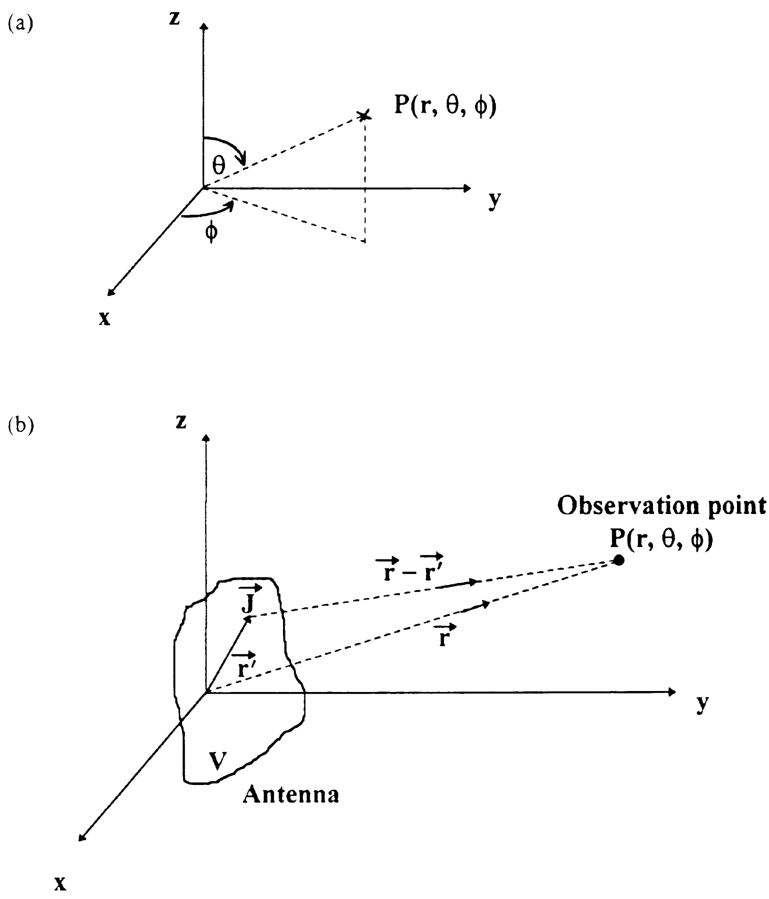


FIGURE 3.7 Antenna analysis: (*a*) spherical coordinates; (*b*) antenna and observation point.

The radiation is due to the current flow on the antenna, which contributes to a vector potential at point $P(r, \theta, \phi)$. This vector potential is the solution of Eq. (3.13), and the result is given by [1]

$$\vec{A}(\vec{r}) = \frac{\mu}{4\pi} \int_V \vec{J}(\vec{r}') \frac{e^{-jk_0|\vec{r}-\vec{r}'|}}{|\vec{r}-\vec{r}'|} dV' \tag{3.15}$$

where r' is the source coordinate and r is the observation point coordinate. The integral is carried over the antenna volume with the current distribution multiplied by the free-space Green's function, defined by

$$\text{Free-space Green's function} = \frac{e^{-jk_0|\vec{r}-\vec{r}'|}}{|\vec{r}-\vec{r}'|} \tag{3.16}$$

If the current distribution is known, then $\vec{A}(\vec{r})$ can be determined. From $\vec{A}(\vec{r})$, one can find $\vec{H}(\vec{r})$ from Eq. (3.14) and thus the electric field $\vec{E}(\vec{r})$. However, in many cases, the current distribution is difficult to find, and numerical methods are generally used to determine the current distribution.

3.5 ANTENNA CHARACTERISTICS AND PARAMETERS

There are many parameters used to specify and evaluate a particular antenna. These parameters provide information about the properties and characteristics of an antenna. In the following, these parameters will be defined and described.

3.5.1 Input VSWR and Input Impedance

As the one-port circuit, an antenna is described by a single scattering parameter S_{11} or the reflection coefficient Γ, which gives the reflected signal and quantifies the impedance mismatch between the source and the antenna. From Chapter 2, the input VSWR and return loss are given by

$$\text{VSWR} = \frac{1+|\Gamma|}{1-|\Gamma|} \tag{3.17}$$

$$\text{RL in dB} = -20\log|\Gamma| \tag{3.18}$$

The optimal VSWR occurs when $|\Gamma| = 0$ or VSWR $= 1$. This means that all power is transmitted to the antenna and there is no reflection. Typically, VSWR $\leqslant 2$ is acceptable for most applications. The power reflected back from the antenna is $|\Gamma|^2$ times the power available from the source. The power coupled to the antenna is $(1 - |\Gamma|^2)$ times the power available from the source.

The input impedance is the one-port impedance looking into the antenna. It is the impedance presented by the antenna to the transmitter or receiver connected to it. The input impedance can be found from Γ by

$$Z_{\text{in}} = Z_0 \frac{1+\Gamma}{1-\Gamma} \tag{3.19}$$

where Z_0 is the characteristic impedance of the connecting transmission line. For a perfect matching, the input impedance should be equal to Z_0 .

3.5.2 Bandwidth

The bandwidth of an antenna is broadly defined as the range of frequencies within which the performance of the antenna, with respect to some characteristic, conforms to a specified standard. In general, bandwidth is specified as the ratio of the upper frequency to the lower frequency or as a percentage of the center frequency. Since antenna characteristics are affected in different ways as frequency changes, there is no unique definition of bandwidth. The two most commonly used definitions are pattern bandwidth and impedance bandwidth.

The power entering the antenna depends on the input impedance locus of the antenna over the frequencies. Therefore, the impedance bandwidth (BW) is the range of frequencies over which the input impedance conforms to a specified standard. This standard is commonly taken to be VSWR ≤ 2 (or $|\Gamma| \leq \frac{1}{3}$) and translates to a reflection of about 11% of input power. Figure 3.8 shows the bandwidth definition [2]. Some applications may require a more stringent specification, such as a VSWR of 1.5 or less. Furthermore, the operating bandwidth of an antenna could be smaller than the impedance bandwidth, since other parameters (gain, efficiency, patterns, etc.) are also functions of frequencies and may deteriorate over the impedance bandwidth.

3.5.3 Power Radiation Patterns

The power radiated (or received) by an antenna is a function of angular position and radial distance from the antenna. At electrically large distances, the power density drops off as $1/r^2$ in any direction. The variation of power density with angular position can be plotted by the radiation pattern. At electrically large distances (i.e., far-field or plane-wave regions), the patterns are independent of distance.

The complete radiation properties of the antenna require that the electric or magnetic fields be plotted over a sphere surrounding the antenna. However, it is often enough to take principal pattern cuts. Antenna pattern cuts are shown in Fig. 3.9. As shown, the antenna has E- and H-plane patterns with co- and cross-polarization components in each. The E-plane pattern refers to the plane containing the electric field vector (E_θ) and the direction of maximum radiation. The parameter E_ϕ is the cross-polarization component. Similarly, the H-plane pattern contains the magnetic field vector and the direction of maximum radiation. Figure 3.10 shows an antenna pattern in either the E- or H-plane. The pattern contains information about half-power beamwidth, sidelobe levels, gain, and so on.

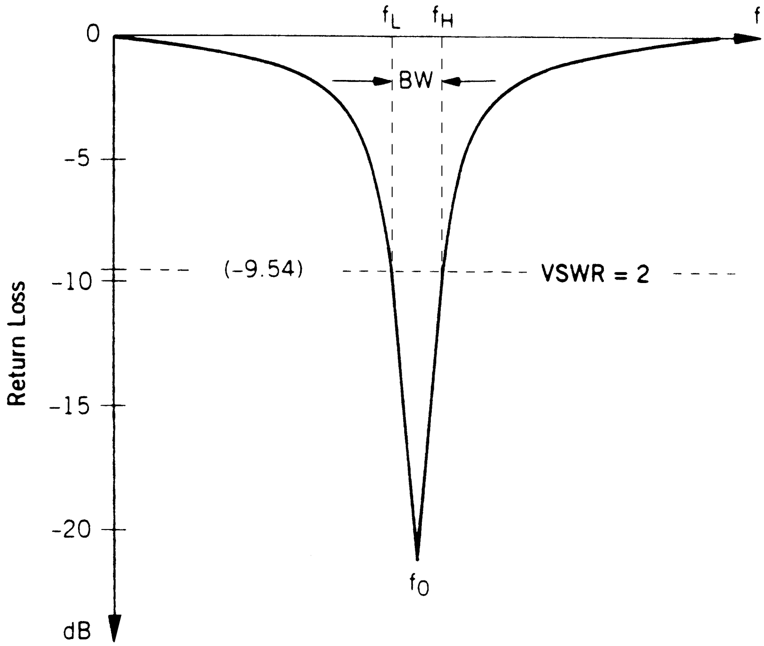


FIGURE 3.8 VSWR = 2 bandwidth [2].

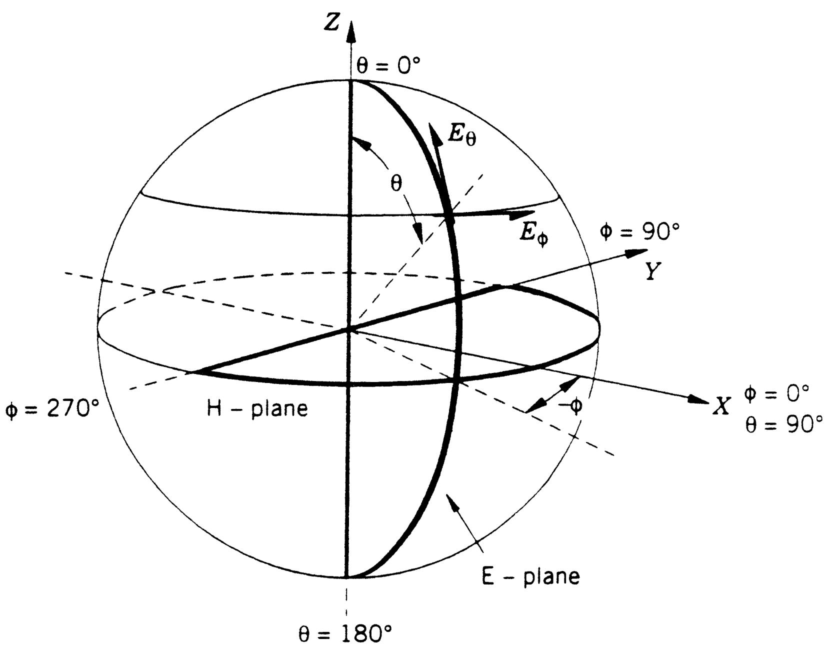


FIGURE 3.9 Antenna pattern coordinate convention [2].

Electric and magnetic fields of an antenna:

$$\mathbf{E} = [E_\theta(\theta,\varphi)\hat{\boldsymbol{\theta}} + E_\varphi(\theta,\varphi)\hat{\boldsymbol{\varphi}}]\frac{e^{-j\mathbf{k}\cdot\mathbf{r}}}{4\pi r} = \mathbf{E}_\theta + \mathbf{E}_\varphi$$

$$\mathbf{H} = \frac{\hat{\mathbf{r}} \times \mathbf{E}}{\eta}$$

$\mathbf{E}_\theta$: Vertically polarized component (Elevation pattern)

$\mathbf{E}_\varphi$: Horizotally polarized component (Azimuth pattern)

Elevation: 고각 (수직방향 각도)

Azimuth: 방위각 (수평방향 각도)

Vertically polarized antenna:

$$\mathbf{E} = E_\theta(\theta,\varphi)\hat{\boldsymbol{\theta}}\frac{e^{-j\mathbf{k}\cdot\mathbf{r}}}{4\pi r}$$

$$\mathbf{H} = \frac{E_\theta(\theta,\varphi)}{\eta}\hat{\boldsymbol{\varphi}}\frac{e^{-j\mathbf{k}\cdot\mathbf{r}}}{4\pi r}$$

$E_\theta(\theta, \varphi = \text{cost})$: E-plane pattern

$E_\theta(\theta = 90°, \varphi)$: H-plane pattern

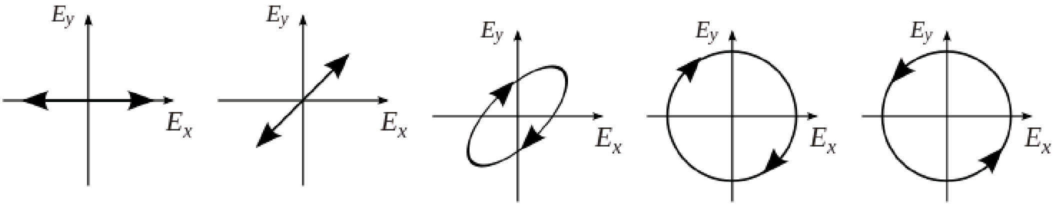
E-plane: 전계면 = 전기장과 평행한 면

H-plane: 자계면 = 자기장과 평행한 면

Antenna Polarization

1. Polarization Design

Polarization: 안테나 편파 = 방사된 전기장의 변화궤적 모양



좌로부터 1 - 5 번. 전파는 +z 방향을 진행

1 = 수평편파, 2 = 45° 편파, 3 = 타원편파, 4 = 좌원편파, 5 = 우원편파

○ 각 경우 전기장의 페이저

1: $E_x = 1, E_y = 0$

2: $E_x = 1/\sqrt{2}, E_y = 1/\sqrt{2}$

3: $E_x = \frac{1}{\sqrt{2}} e^{-j45°}, E_y = \frac{1}{\sqrt{2}}$

4: $E_x = \frac{1}{\sqrt{2}} e^{-j45°}, E_y = \frac{1}{\sqrt{2}} e^{j45°}$

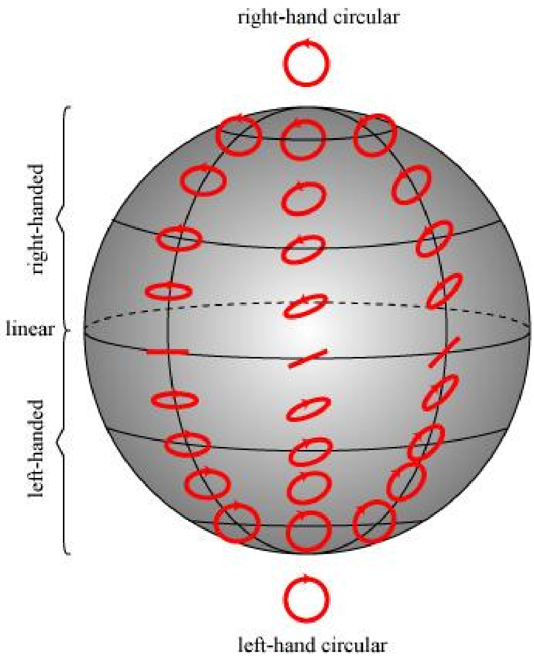
5: $E_x = \frac{1}{\sqrt{2}} e^{j45°}, E_y = \frac{1}{\sqrt{2}} e^{-j45°}$

(중요)

원편파: x 성분 전기장과 y 성분 전기장이 크기는 같고 위상이 90° 차이

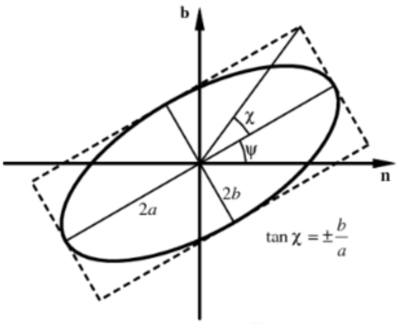
선형편파: : x 성분 전기장과 y 성분 전기장의 위상 차가 0°

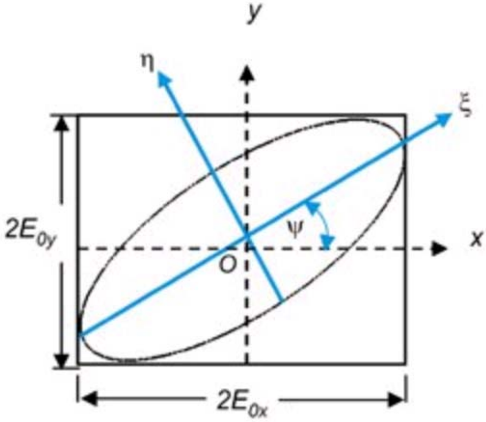
2. Poincare Sphere



3. Polarization Ellipse

$$E_x = A_x e^{j\varphi_x} e^{-jkz};\ E_x = A_y e^{j\varphi_y} e^{-jkz}$$





$\delta = \varphi_y - \varphi_x$

$\tan\alpha = \dfrac{A_y}{A_x}, 0 \le \alpha \le \pi/2$

$\tan 2\psi = (\tan 2\alpha)\cos\delta,\ 0 \le \psi \le \pi$

$\sin 2\chi = (\sin 2\alpha)\sin\delta,\ -\pi/4 \le \chi \le \pi/4$

Axial ratio (축비) = 편파궤적 타원에서 타원의 긴지름을 짧은 지름으로 나눈 값

$$\boxed{\mathrm{AR} = \frac{b}{a},\ 1 \le \mathrm{AR} < \infty;\ \ \mathrm{AR(dB)} = 20\log_{10}\mathrm{AR}}$$

Axial ratio criterion: AR < 3 dB

선형편파 축비: ∞

원편파 축비: 1 (0 dB)

4. Polarization Matching (편파정합)

○ 편파 정합

- 송신 안테나 편파와 수신 안테나 편파가 동일한 경우 = 편파 정합
- 이 경우 편파 손실 없음.
- 편파 정합 사례:

  수직편파와 수직편파

  우원편파와 좌원편파

○ 직교 편파

- 서로 직교하는 (공통점이 0 인) 편파
- 직교 편파 사례:

  수직편파와 수평편파

  좌원편파와 우원편파

- 송신 안테나의 편파와 수신 안테나의 편파가 직교하는 경우 수신 전력이 0

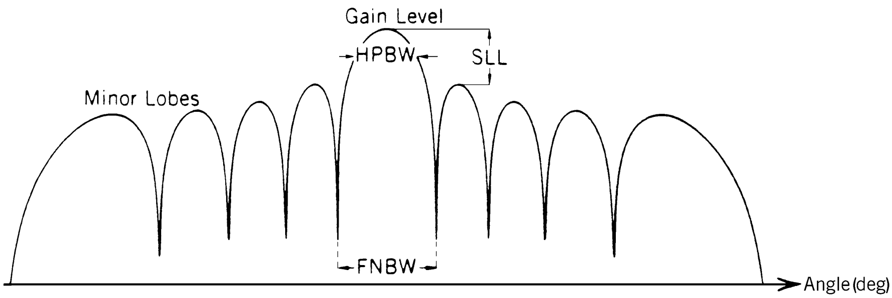


FIGURE 3.10 Antennas pattern characteristics [2].

3.5.4 Half-Power Beamwidth and Sidelobe Level

The half-power beamwidth (HPBW) is the range in degrees such that the radiation drops to one-half of (or 3 dB below) its maximum. The sidelobes are power radiation peaks in addition to the main lobe. The sidelobe levels (SLLs) are normally given as the number of decibels below the main-lobe peak. Figure 3.10 [2] shows the HPBW and SLLs. Also shown is FNBW, the first-null beamwidth.

3.5.5 Directivity, Gain, and Efficiency

The directivity D_{max} is defined as the value of the directive gain in the direction of its maximum value. The directive gain $D(\theta, \phi)$ is the ratio of the Poynting power density $S(\theta, \phi)$ over the power density radiated by an isotropic source. Therefore, one can write

$$D(\theta, \phi) = \frac{S(\theta, \phi)}{P_t/4\pi R^2} \tag{3.20}$$

$$D_{max} = \frac{\text{maximum of } S(\theta, \phi)}{P_t/4\pi R^2} \tag{3.21}$$

where $\vec{S}(\theta, \phi) = \frac{1}{2}\,\text{Re}[\vec{E} \times \vec{H}^*]$.

The directivity of an isotropic antenna equals to 1 by definition, and that of other antennas will be greater than 1. Thus, the directivity serves as a figure of merit relating the directional properties of an antenna relative to those of an isotropic source.

The gain of an antenna is the directivity multiplied by the illumination or aperture efficiency of the antenna to radiate the energy presented to its terminal:

$$\text{Gain} = G = \eta D_{max} \tag{3.22}$$

$$\eta = \text{efficiency} = \frac{P_{rad}}{P_{in}} = \frac{P_{rad}}{P_{rad} + P_{loss}} \tag{3.23}$$

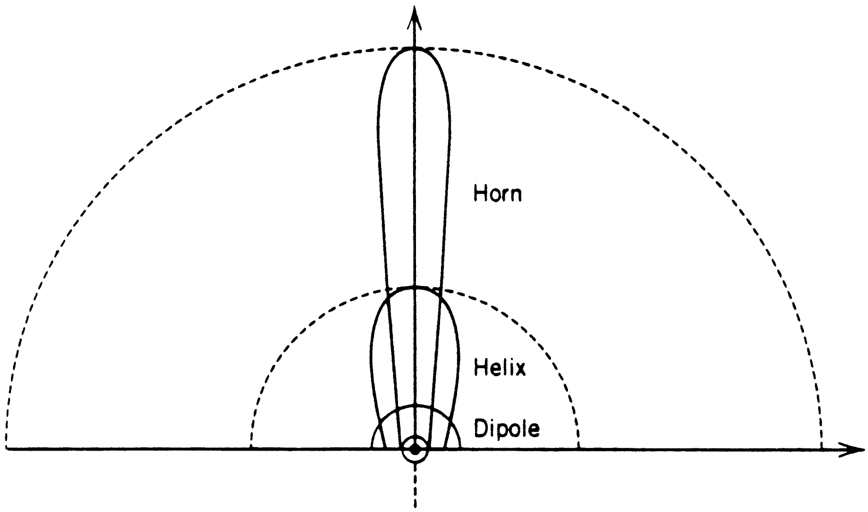


FIGURE 3.11 Gain comparison [2].

where P_{rad} is the actual power radiated, P_{in} is the power coupled into the antenna, and P_{loss} is the power lost in the antenna. The losses could include ohmic or conductor loss, dielectric loss, and so on. In general, the narrower the beamwidth, the higher the gain. Figure 3.11 gives a comparison of gain for three different antennas. From Eqs. (3.21) and (3.22), the radiated power density in the direction of its maximum value is $P_{d,\max} = G(P_t/4\pi R^2)$.

3.5.6 Polarization and Cross-Polarization Level

The polarization of an antenna is the polarization of the electric field of the radiated wave. Antennas can be classified as linearly polarized (LP) or circularly polarized (CP). The polarization of the wave is described by the tip of the E-field vector as time progresses. If the locus is a straight line, the wave is linearly polarized. If the locus is a circle, the wave is circularly polarized. Ideally, linear polarization means that the electric field is in only one direction, but this is seldom the case. For linear polarization, the cross-polarization level (CPL) determines the amount of polarization impurity. As an example, for a vertically polarized antenna, the CPL is due to the E-field existing in the horizontal direction. Normally, CPL is a measure of decibels below the copolarization level.

3.5.7 Effective Area

The effective area (A_e) is related to the antenna gain by

$$G = \frac{4\pi}{\lambda_0^2} A_e \tag{3.24}$$

Friis Transmission Formula (프리스 전송 공식)

송신 안테나가 송신할 경우 멀리 떨어진 수신 안테나에 수신되는 전력을 계산하는 식

$$\boxed{P_r = SA_e}$$

P_r : 수신 전력 (W)

S : 수신 안테나에 입사된 전파의 전력밀도 (W/m^2)

A_e : 수신 안테나의 유효면적 (m^2)

$$\boxed{S = \frac{P_t}{4\pi R^2} G_t}$$

P_t : 송신 전력 (W)

S : 수신 안테나에 입사된 전파의 전력밀도 (W/m^2)

G_t : 송신 안테나 이득 (단위 없음, dB 단위로 주어질 경우 dB 변환 전 선형단위 값 사용)

R : 송신 안테나와 수신 안테나의 이격 거리 (m)

$$\boxed{A_e = \frac{\lambda^2}{4\pi} G_r}$$

A_e : 수신 안테나의 유효면적 (m^2)

λ : 전파의 파장 (m). $\lambda = 3 \cdot 10^8 \ / f$

G_r : 수신 안테나의 이득 (단위 없음, dB 단위로 주어질 경우 dB 변환 전 선형단위 값 사용)

$$\boxed{P_r = P_t \left(\frac{\lambda}{4\pi R} \right)^2 G_t G_r}$$

P_r : 수신 전력 (W)

P_t : 송신 전력 (W)

R : 송신 안테나와 수신 안테나의 이격 거리 (m)

λ : 전파의 파장 (m). $\lambda = 3 \cdot 10^8 \ / f$

G_t : 송신 안테나 이득 (단위 없음, dB 단위로 주어질 경우 dB 변환 전 선형단위 값 사용)

G_r : 수신 안테나의 이득 (단위 없음, dB 단위로 주어질 경우 dB 변환 전 선형단위 값 사용)

It is easier to visualize the concept of effective area when one considers a receiving antenna. It is a measure of the effective absorbing area presented by an antenna to an incident wave [3]. The effective area is normally proportional to, but less than, the physical area of the antenna.

3.5.8 Beam Efficiency

Beam efficiency is another frequently used parameter to gauge the performance of an antenna. Beam efficiency is the ratio of the power received or transmitted within a cone angle to the power received or transmitted by the whole antenna. Thus, beam efficiency is a measure of the amount of power received or transmitted by minor lobes relative to the main beam.

3.5.9 Back Radiation

The back radiation is directed to the backside of an antenna. Normally it is given as the back-to-front ratio in decibels.

3.5.10 Estimation of High-Gain Antennas

There are some convenient formulas for making quick estimates of beamwidths and gains of electrically large, high gain antennas. A convenient equation for predicting a 3-dB beamwidth is [3]:

$$\mathrm{BW} = K_1 \frac{\lambda_0}{D} \tag{3.25}$$

where D is the aperture dimension in the plane of the pattern. For a rough estimate, one can use $K_1 = 70°$. For an example, if the length of an antenna is 10 cm, the beamwidth at 30 GHz, in the plane of length, is 7°.

A convenient equation for predicting gain is given in reference [3]:

$$G = \frac{K_2}{\theta_1 \theta_2} \tag{3.26}$$

where K_2 is a unitless constant and θ_1 and θ_2 are the 3-dB beamwidths in degrees in the two orthogonal principal planes. The correct K_2 value depends on antenna efficiency, but a rough estimate can be made with $K_2 = 30{,}000$.

Example 3.1 The E-plane pattern of an eight-element microstrip patch antenna array is shown in Fig. 3.12 [4]. Describe the characteristics of this pattern.

Solution From the pattern shown in Fig. 3.12, it can be seen that the gain is 11.4 dB. The half-power beamwidth is about 22.2°. The cross-polarization radiation level is over 26 dB below the copolarization radiation in the main beam. The first

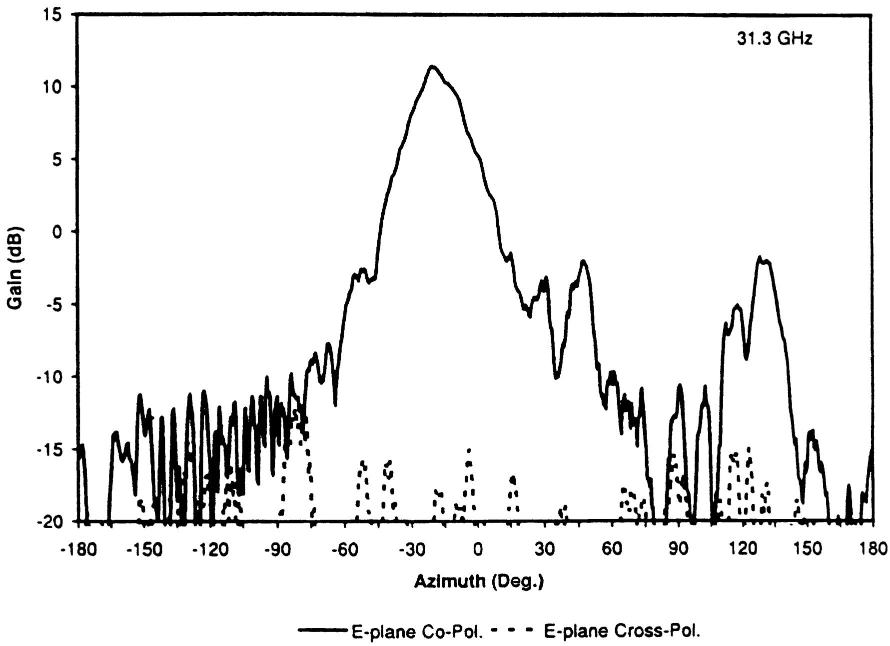


FIGURE 3.12 *E*-plane pattern of an eight-element microstrip patch antenna fed by an image line operating at 31.3 GHz. (From reference [4], with permission from IEEE.)

SLL is about 14 dB below the main lobe. The maximum back radiation occurs at around 135° with a level of about 13 dB below the main-lobe peak. ■

3.6 MONOPOLE AND DIPOLE ANTENNAS

The monopole and dipole antennas are commonly used for broadcasting, cellular phones, and wireless communications due to their omnidirective property. Figure 3.13 shows some examples. A monopole together with its image through a metal or ground plane has radiation characteristics similar to a dipole. A dipole with a length l is approximately equivalent to a monopole with a length of $\frac{1}{2}l$ placed on a metal or ground plane. For a dipole with a length $l < \lambda_0$, the E-plane radiation pattern is a doughnut shape with a hole or figure-eight shape, as shown in Fig. 3.14. The maximum radiation occurs when $\theta = 90°$ and there is no radiation at $\theta = 0°$. The H-plane radiation pattern is a circle, which means the antenna radiates equally in all ϕ directions. Therefore, it is nondirective in the ϕ direction. Since it is only directive in the θ direction, it is called an omnidirective antenna. Because it is nondirective in the ϕ direction, the antenna can receive a signal coming from any direction in the H-plane. This makes the antenna useful for broadcast and wireless applications. The antenna has no gain in the H-plane, that is, $G = 1$. In the E-plane pattern, the gain is fairly low since it has a broad beamwidth.

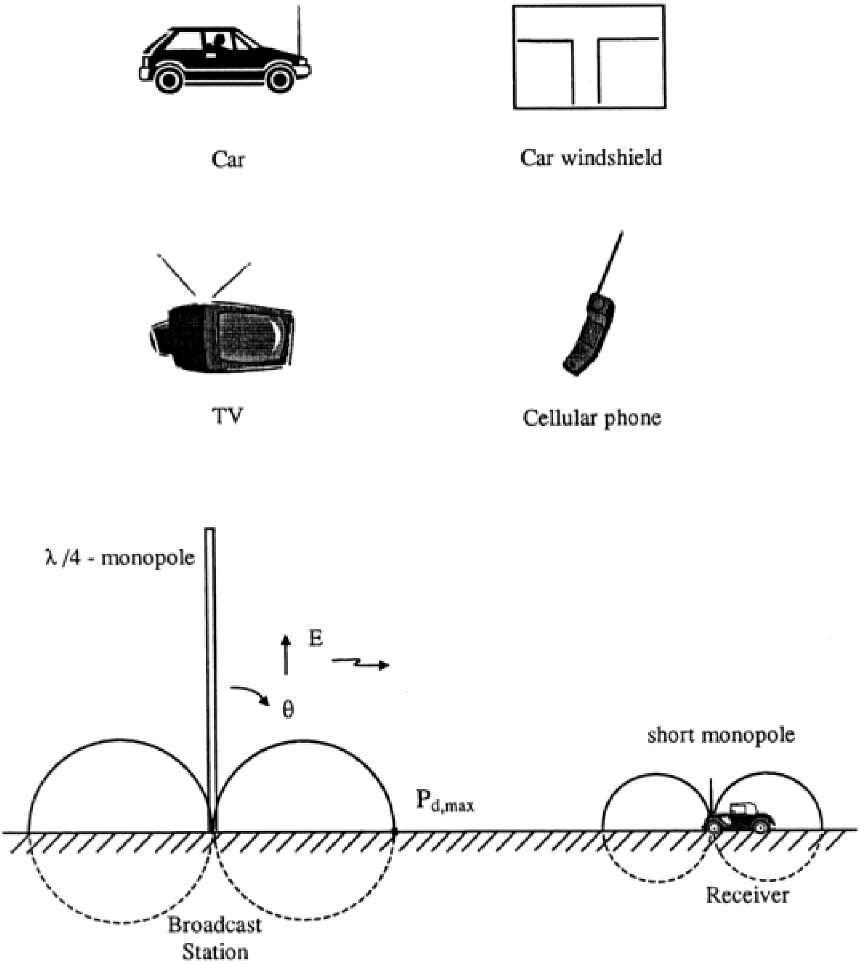


FIGURE 3.13 Examples of dipole and monopole antennas.

By assuming a sinusoidal current distribution on the dipole, the radiated fields can be found from Eqs. (3.14) and (3.15) [5]. Figure 3.15 shows the pattern plots for different dipole length l's [3]. It can be seen that the pattern deteriorates from its figure-eight shape when $l > \lambda_0$. In most cases, a dipole with $l < \lambda_0$ is used.

When $l = \frac{1}{2}\lambda_0$, it is called a half-wave dipole. A half-wave dipole (or a quarter-wavelength monopole) has an input impedance approximately equal to 73 Ω and an antenna gain of 1.64. For a very short dipole with $l \ll \lambda_0$, the input impedance is very small and difficult to match. It has low efficiency, and most power is wasted. The gain for a short dipole is approximately equal to 1.5 (or 1.7 dB).

In practice, the antenna is always fed by a transmission line. Figure 3.16 shows a quarter-wavelength monopole for mobile communication applications. This type of antenna normally uses a flexible antenna element and thus is called the quarter-

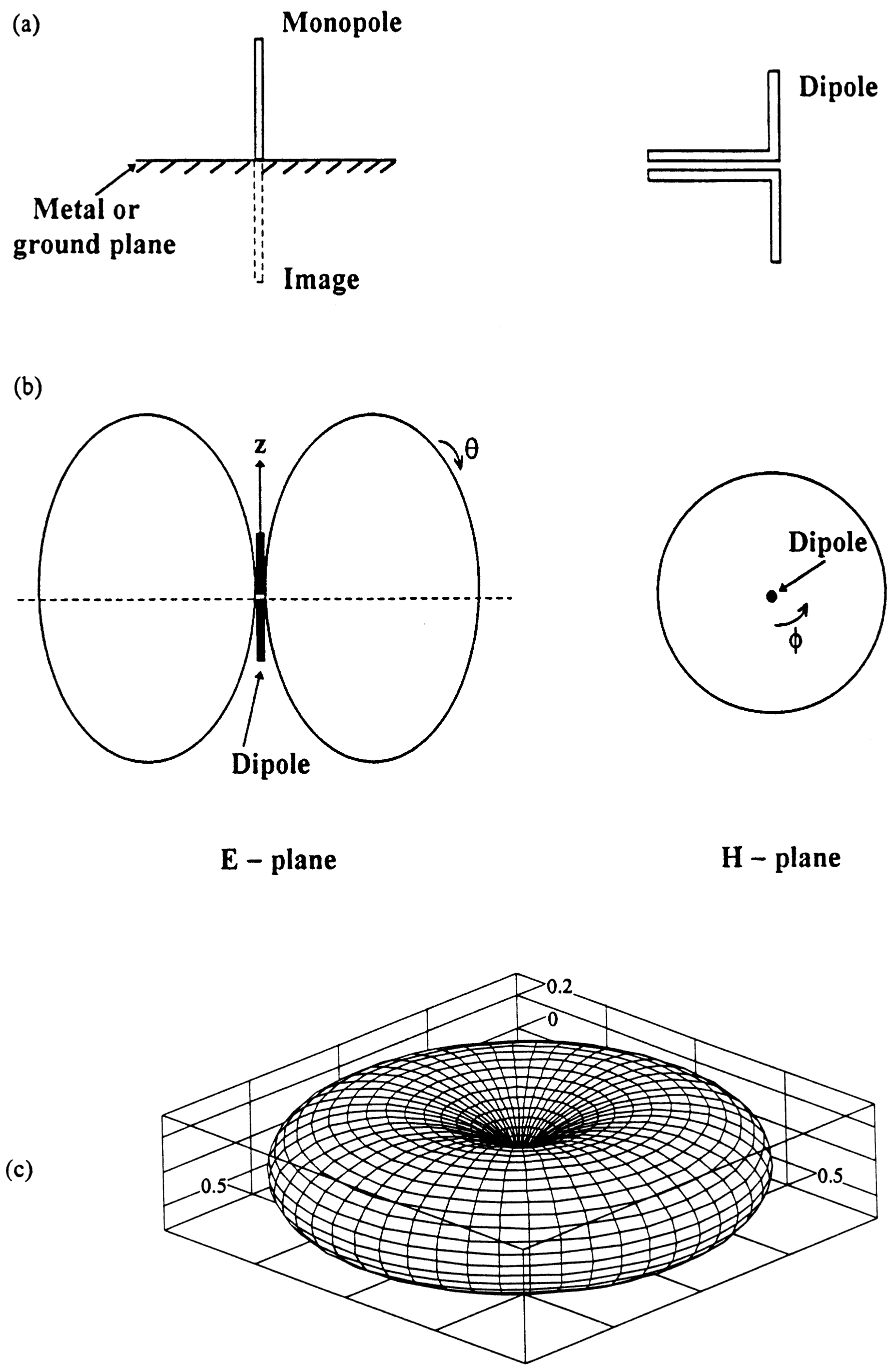


FIGURE 3.14 (*a*) Thin-wire dipole and monopole antenna. (*b*) *E*- and *H*-plane radiation patterns. (*c*) Three-dimensional view of the radiation pattern.

FIGURE 3.15 Radiation patterns of center-driven dipole assuming sinusoidal current distribution and very thin dipole. (From reference [3], with permission from McGraw-Hill.)

wavelength "whip" antenna. The antenna is mounted on the ground plane, which is the roof of a car. If the ground plane is assumed to be very large and made of a perfect conductor, the radiation patterns of this antenna would be the same as that of a half-wave dipole. However, the input impedance is only half that of a half-wave dipole.

There are various types of dipoles or monopoles used for wireless applications. A folded dipole is formed by joining two cylindrical dipoles at the ends, as shown in Fig. 3.17. The excitation of a folded dipole can be considered as a superposition of two modes, a symmetrical mode and an asymmetrical mode [6].

Figure 3.18 shows a sleeve antenna [7]. The coaxial cylindrical skirt behaves as a quarter-wavelength choke, preventing the antenna current from leaking into the outer surface of the coaxial cable. The choke on the lower part of the coaxial cable is used to improve the radiation pattern by further suppressing the current leakage. This antenna does not require a ground plane and has almost the same characteristics as that of a half-wavelength dipole antenna. The feeding structure is more suitable for vehicle mounting than the center-fed dipole antenna.

Other variations of monopole antennas are inverted L and inverted F antennas, as shown in Figs. 3.19*a* and *b*. These are low-profile antennas formed by bending a quarter-wavelength monopole element mounted on a ground plane into an L-shape or F-shape. The wire can be replaced by a planar element with wide-band characteristics [7]. The modified structure is shown in Fig. 3.19*c*. These types of antennas are used on hand-held portable telephones.

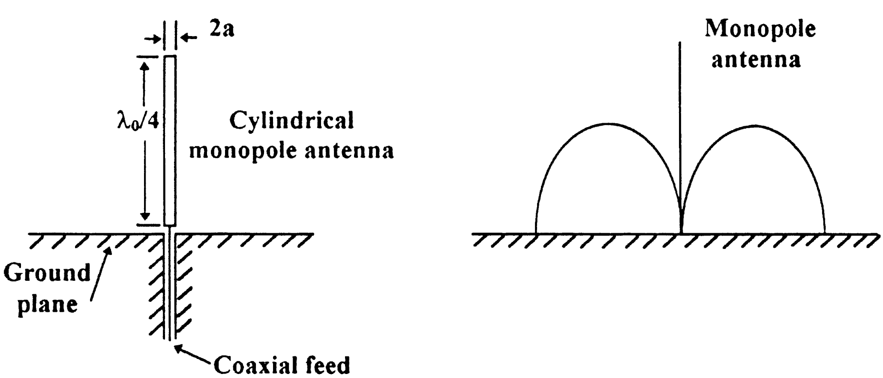


FIGURE 3.16 Quarter-wavelength whip antenna and its pattern.

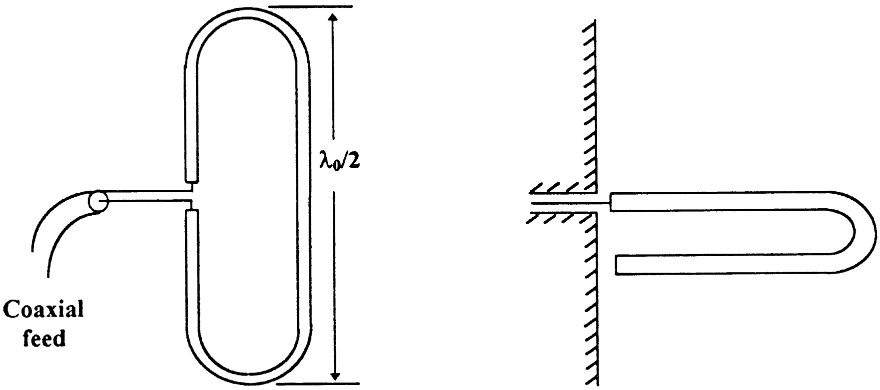


FIGURE 3.17 Folded dipoles and monopoles.

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FIGURE 3.18 Sleeve antenna for vehicle application. (From reference [7], with permission from McGraw-Hill.)

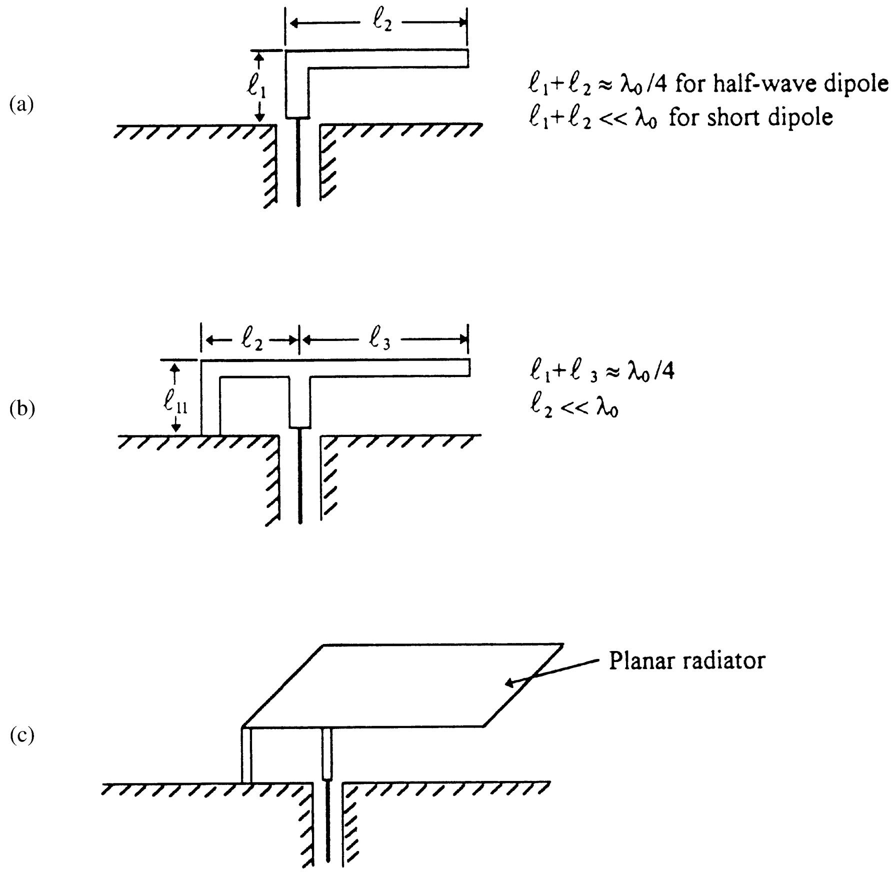


FIGURE 3.19 (*a*) Inverted *L*-antenna (ILA). (*b*) Inverted *F*-antenna (IFA). (*c*) Planar inverted *F*-antenna.

Example 3.2 An AM radio station operates at 600 kHz transmitting an output power of 100 kW using a monopole antenna as shown in Fig. 3.20. (a) What is the length of l if the antenna is an equivalent half-wave dipole? (b) What is the maximum rms electric field in volts per meter at a distance 100 km away from the station? The half-wave dipole has an antenna gain of 1.64.

Solution

(a)

$$\lambda_0 = \frac{c}{f} = \frac{3 \times 10^8 \text{ m/sec}}{600 \times 10^3 \text{ sec}^{-1}} = 500 \text{ m}$$

$$2l = \text{monopole and its image} = \tfrac{1}{2}\lambda_0$$

$$l = \tfrac{1}{4}\lambda_0 = 125 \text{ m}$$

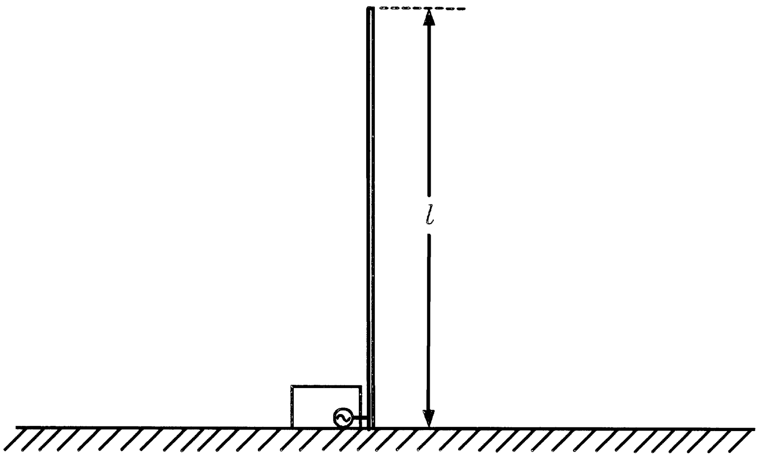


FIGURE 3.20 Broadcast station uses a monopole antenna.

(b) The power density for an isotropic antenna at a distance R is equal to $P_t/4\pi R^2$, as shown in Eq. (3.1). For a directive antenna with a gain G, the maximum power density is

$$P_{d,\max} = G\frac{P_t}{4\pi R^2} = 1.64 \times \frac{100 \times 10^3 \text{ W}}{4 \times 3.14 \times (100 \times 10^3 \text{ m})^2}$$
$$= 1.31 \times 10^{-6} \text{ W/m}^2$$

The maximum power density occurs when $\theta = 90°$, that is, on the ground.

From Eqs. (3.6) and (3.7),

$$P_d = \frac{E_{\text{rms}}^2}{\eta_0} \qquad \eta_0 = 377\ \Omega$$

The maximum E_{rms} is

$$E_{\text{rms}} = \sqrt{P_{d,\max}\eta_0} = 22.2 \text{ mV/m}$$ ■

3.7 HORN ANTENNAS

The horn antenna is a transition between a waveguide and free space. A rectangular waveguide feed is used to connect to a rectangular waveguide horn, and a circular waveguide feed is for the circular waveguide horn. The horn antenna is commonly used as a feed to a parabolic dish antenna, a gain standard for antenna gain measurements, and as compact medium-gain antennas for various systems. Its gain can be calculated to within 0.1 dB accuracy from its known dimensions and is therefore used as a gain standard in antenna measurements.

For a rectangular pyramidal horn, shown in Fig. 3.21a, the dimensions of the horn for optimum gain can be designed by setting [3]

$$A = \sqrt{3\lambda_0 l_h}, \qquad B = \sqrt{2\lambda_0 l_e} \tag{3.27}$$

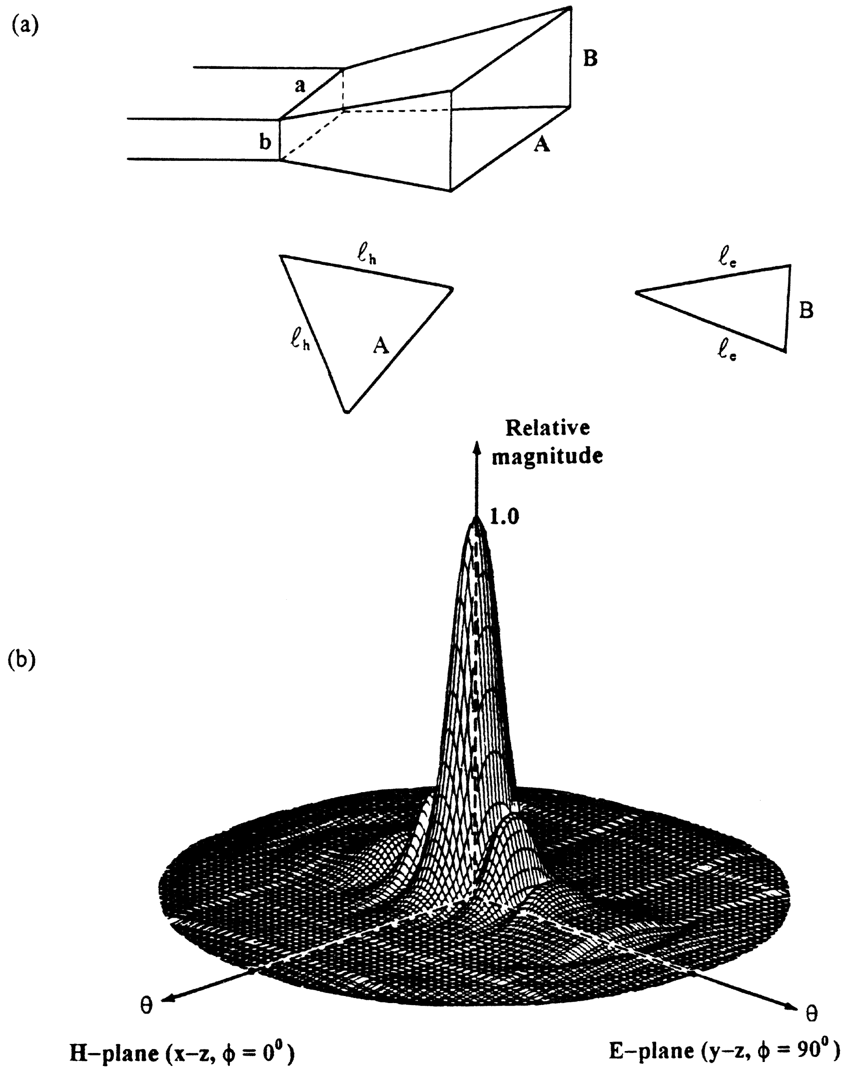


FIGURE 3.21 Rectangular pyramidal horn fed by a rectangular waveguide: (a) configuration and (b) typical three-dimensional radiation pattern.

where A and B are dimensions of the horn and l_e and l_h are the slant lengths of the horn, as shown in Fig. 3.21. The effective area is close to 50% of its aperture area, and its gain is given as

$$\text{Gain (in dB)} = 8.1 + 10\log\frac{AB}{\lambda_0^2} \tag{3.28}$$

As an example, a horn with $A = 9$ in. and $B = 4$ in. operating at 10 GHz will have a gain of 22.2 dB. A typical three-dimensional pattern is shown in Fig. 3.21*b* [8].

For an optimum-gain circular horn, shown in Fig. 3.22, the diameter should be designed as

$$D = \sqrt{3l_c\lambda_0} \tag{3.29a}$$

and

$$\text{Gain (in dB)} = 20\log\frac{\pi D}{\lambda_0} - 2.82 \tag{3.29b}$$

3.8 PARABOLIC DISH ANTENNAS

A parabolic dish is a high-gain antenna. It is the most commonly used reflector antenna for point-to-point satellites and wireless links.

A parabolic dish is basically a metal dish illuminated by a source at its focal point. The spherical wavefront illuminated by the source is converted into a planar wavefront by the dish, as shown in Fig. 3.23 [9].

For an illumination efficiency of 100%, the effective area equals the physical area:

$$A_e = \pi\left(\frac{D}{2}\right)^2 = A \tag{3.30}$$

where D is the diameter of the dish.

In practice, the illumination efficiency η is typically between 55 and 75% due to the feed spillover, blockage, and losses. Using a 55% efficiency for the worst case, we have

$$A_e = \eta A = 0.55\pi(\tfrac{1}{2}D)^2 \tag{3.31}$$

and the gain of the antenna is, from (3.24),

$$G = \frac{4\pi}{\lambda_0^2}A_e = 0.55\left(\frac{\pi D}{\lambda_0}\right)^2 \tag{3.32}$$

The half-power beamwidth is approximately given by [from Eq. (3.25) with $K_1 = 70°$]

$$\text{HPBW} = 70\frac{\lambda_0}{D} \quad \text{(deg)} \tag{3.33}$$

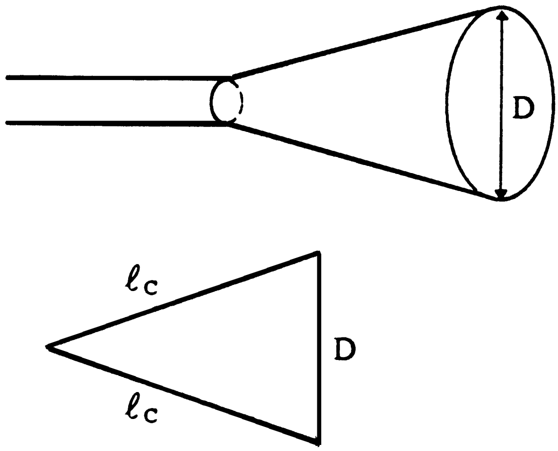


FIGURE 3.22 Circular horn fed by a circular waveguide.

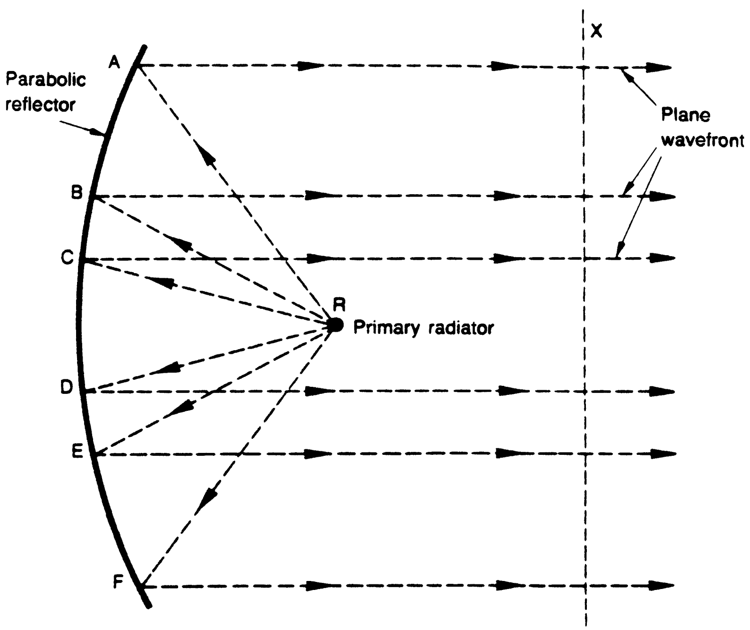


FIGURE 3.23 Radiation from a parabolic dish antenna. (From reference [9], with permission from Longman Scientific & Technical.)

Example 3.3 A parabolic dish antenna with a diameter of 3 ft is operated at 10 GHz. Determine the approximate gain, beamwidth, and distance for the far-field region operation. The illumination efficiency is 55%.

Solution

$$D = 3 \text{ ft} = 36 \text{ in.}$$

$$\lambda_0 = \frac{c}{f} = 3 \text{ cm} = 1.18 \text{ in.}$$

$$\text{Gain} = 10 \log\left[0.55\left(\frac{\pi D}{\lambda_0}\right)^2\right] = 37 \text{ dB or } 5047$$

$$\text{Beamwidth} = 70\frac{\lambda_0}{D} = 2.29^\circ$$

For far-field region operation

$$R > \frac{2D^2}{\lambda_0} = 2196 \text{ in. or } 183 \text{ ft.}$$ ■

It can be seen that the dish antenna provides a very high gain and narrow beam. The alignment of the dish antenna is usually very critical. The parabolic dish is generally fed by a horn antenna connected to a coaxial cable. There are four major feed methods: front feed, Cassegrain, Gregorian, and offset feed. Figure 3.24 shows these arrangements [9]. The front feed is the simplest method. The illumination efficiency is only 55–60%. The feed and its supporting structure produce aperture blockage and increase the sidelobe and cross-polarization levels. The Cassegrain method has the advantages that the feed is closer to other front-end hardware and a shorter connection line is needed. The Gregorian method is similar to the Cassegrain feed, but an elliptical subreflector is used. An illumination efficiency of 76% can be achieved. The offset feed method avoids the aperture blockage by the feed or subreflector. The sidelobe levels are smaller, and the overall size is smaller for the same gain.

At low microwave frequencies or ultrahigh frequencies (UHFs), a parabolic dish is big, so only a portion of the dish is used instead. This is called a truncated parabolic dish, commonly seen on ships. To make the dish lighter and to withstand strong wind, a dish made of metal mesh instead of solid metal can be used.

3.9 MICROSTRIP PATCH ANTENNAS

Microstrip patch antennas are widely used due to the fact that they are highly efficient, structurally compact, and conformal. One of the most common types of microstrip antenna is the rectangular patch.

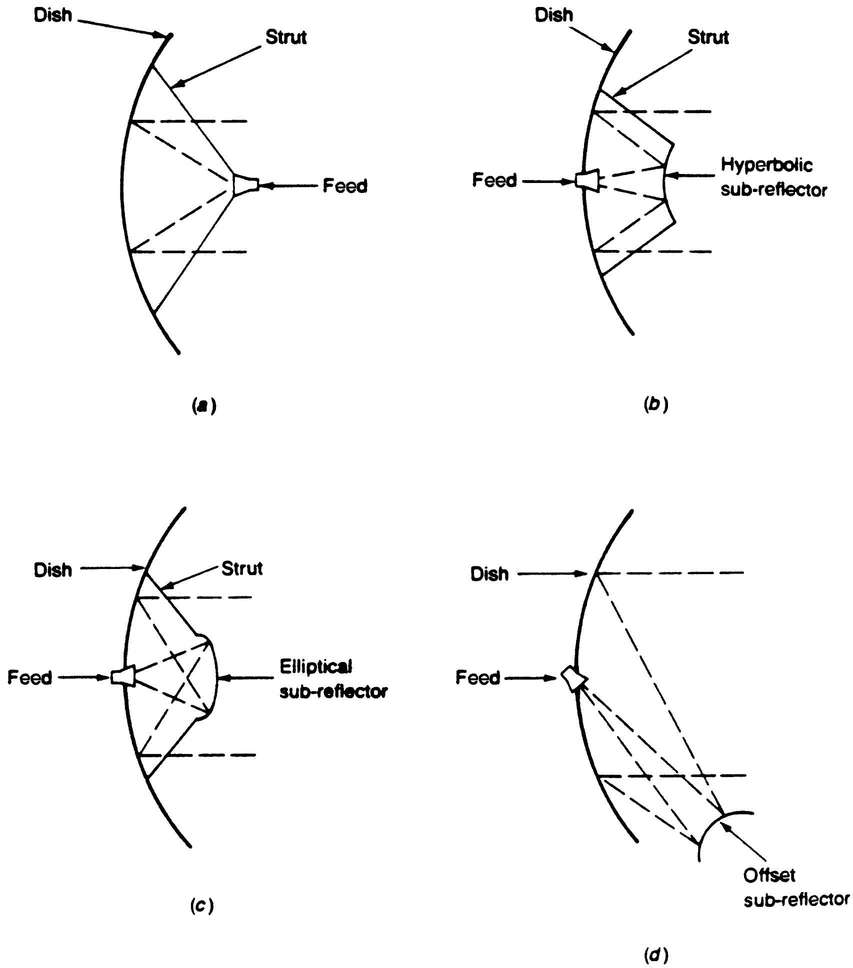


FIGURE 3.24 Parabolic dish aerials: (*a*) front feed; (*b*) Cassegrain; (*c*) Gregorian; and (*d*) offset-feed. (From reference [9], with permission from Longman Scientific & Technical.)

Figure 3.25 shows a typical rectangular patch antenna with width W and length L over a grounded dielectric plane with dielectric constant ε_r. Ideally, the ground plane on the underside of the substrate is of infinite extent. Normally, the thickness of the dielectric substrate, h, is designed to be $\leq 0.02\lambda_g$, where λ_g is the wavelength in the dielectric.

There are several theories that can be used for the analysis and design of microstrip patch antennas. The first one is the transmission line model [10]. This theory is based on the fact that the rectangular patch is simply a very wide transmission line terminated by the radiation impedance. The transmission line model can predict properties only approximately, because the accepted formulas that

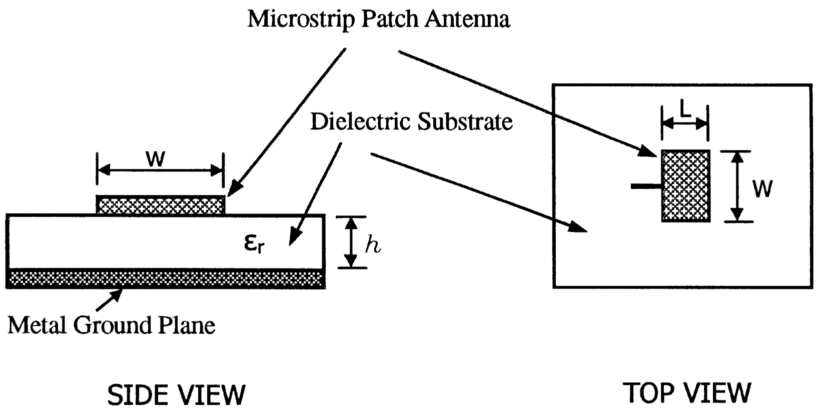


FIGURE 3.25 Rectangular patch antenna.

describe microstrip transmission line characteristics are either approximations or empirically fit to measured data. The second method is the cavity model [11]. This model assumes the rectangular patch to be essentially a closed resonant cavity with magnetic walls. The cavity model can predict all properties of the antenna with high accuracy but at the expense of much more computation effort than the transmission line model.

The patch antenna can be approximately designed by using the transmission line model. It can be seen from Fig. 3.26 that the rectangular patch with length L and width W can be viewed as a very wide transmission line that is transversely resonating, with the electric field varying sinusoidally under the patch along its resonant length. The electric field is assumed to be invariant along the width W of the patch. Furthermore, it is assumed that the antenna's radiation comes from fields leaking out along the width, or radiating edges, of the antenna.

The radiating edges of the patch can be thought of as radiating slots connected to each other by a microstrip transmission line. The radiation conductance for a single slot is given as

$$G = \frac{W^2}{90\lambda_0^2} \quad \text{for } W < \lambda_0 \tag{3.34a}$$

$$G = \frac{W}{120\lambda_0} \quad \text{for } W > \lambda_0 \tag{3.34b}$$

Similarly, the radiation susceptance of a single slot [12] is given as

$$B = \frac{k_0\ \Delta l\sqrt{\varepsilon_{\text{eff}}}}{Z_0} \tag{3.35a}$$

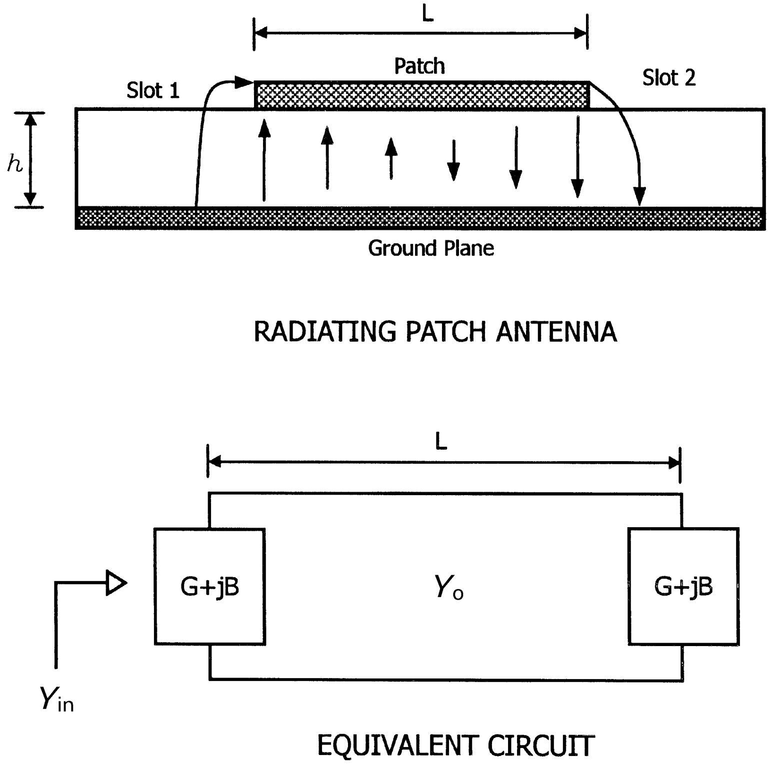


FIGURE 3.26 Transmission line model of a patch antenna.

where

$$Z_0 = \frac{120\pi h}{W\sqrt{\varepsilon_{\mathrm{eff}}}} \tag{3.35b}$$

with

$$\varepsilon_{\mathrm{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left(1 + \frac{12h}{W}\right)^{-1/2} \tag{3.35c}$$

$$\Delta l = 0.412h\left(\frac{\varepsilon_{\mathrm{eff}} + 0.3}{\varepsilon_{\mathrm{eff}} - 0.258}\right)\frac{(W/h) + 0.264}{(W/h) + 0.8} \tag{3.35d}$$

where $k_0 = 2\pi/\lambda_0$ is the free-space wavenumber, Z_0 is the characteristic impedance of the microstrip line with width W, $\varepsilon_{\mathrm{eff}}$ is referred to as the effective dielectric constant, and Δl is the correction term called the edge extension, accounting for the fringe capacitance. In Fig. 3.26 it can be seen that the fields slightly overlap the edges of the patch, making the electrical length of the patch slightly larger than its physical length, thus making the edge extension necessary.

To determine the radiation impedance of the antenna, we combine the slot impedance with the transmission line theory. The admittance of a single slot is given in Eqs. (3.34) and (3.35). The microstrip patch antenna is merely two slots in parallel separated by a transmission line with length L, which has a characteristic admittance Y_0. The input admittance at the radiating edge can be found by adding the slot admittance to the admittance of the second slot by transforming it across the length of the patch using the transmission line equation. The result is

$$Y_{\text{in}} = Y_{\text{slot}} + Y_0 \frac{Y_{\text{slot}} + jY_0 \tan \beta(L + 2\,\Delta l)}{Y_0 + jY_{\text{slot}} \tan \beta(L + 2\,\Delta l)} \tag{3.36}$$

where $\beta = 2\pi\sqrt{\varepsilon_{\text{eff}}}/\lambda_0$ is the propagation constant of the microstrip transmission line. At resonance $(L + 2\,\Delta l = \frac{1}{2}\lambda_g)$, this reduces to two slots in parallel, giving an input admittance twice that of Eq. (3.34),

$$Y_{\text{in}} = 2G \tag{3.37}$$

More generally, the input admittance at a point inside the patch at a given distance y_1 from the radiating edge can be found by using the transmission line equation to transform the slot admittance across the patch by a distance y_1 [13]. We add this to the admittance from the other slot, which is transformed a distance $y_2 = (L + 2\,\Delta l) - y_1$ so the two admittances are at the same point. This result gives

$$Y_{\text{in}} = Y_0 \frac{Y_{\text{slot}} + jY_0 \tan \beta y_1}{Y_0 + jY_{\text{slot}} \tan \beta y_1} + Y_0 \frac{Y_{\text{slot}} + jY_0 \tan \beta y_2}{Y_0 + jY_{\text{slot}} \tan \beta y_2} \tag{3.38}$$

The patch antenna resonates when the imaginary part of (3.38) disappears.

Perhaps a more intuitive picture of resonance can be seen from field distribution under the patch in Fig. 3.26. In order to resonate, the effective length [adding twice the length extension found in Eq. (3.35d) onto the physical length] must be equal to half a transmission line wavelength. In other words,

$$(L + 2\,\Delta l) = \frac{\lambda_g}{2} = \frac{\lambda_0}{2\sqrt{\varepsilon_{\text{eff}}}} \tag{3.39}$$

from which we can determine the resonant frequency (which is the operating frequency) in terms of patch dimensions:

$$f_r = \frac{c}{2\sqrt{\varepsilon_{\text{eff}}}(L + 2\,\Delta l)} \tag{3.40}$$

Here, W is not critical but can be selected as

$$W = \frac{c}{2f_r}\left(\frac{\varepsilon_r + 1}{2}\right)^{-1/2} \tag{3.41}$$

From Eqs. (3.41), (3.35c), (3.35d), and (3.40), one can determine W and L if ε_r, h, and f_r are known. Equations (3.34) and (3.37) give the input admittance at the resonant or operating frequency.

Example 3.4 Design a microstrip patch antenna operating at 3 GHz. The substrate is Duroid 5880 ($\varepsilon_r = 2.2$) with a thickness of 0.030 in. The antenna is fed by a 50-Ω line, and a quarter-wavelength transformer is used for impedance matching, as shown in Fig. 3.27.

Solution

$$f_r = 3 \text{ GHz}, h = 0.030 \text{ in.} = 0.0762 \text{ cm}$$

$$\lambda_0 = \frac{c}{f_r} = 10 \text{ cm}$$

From Eq. (3.41)

$$W = \frac{c}{2f_r}\left(\frac{\varepsilon_r + 1}{2}\right)^{-1/2} = 3.95 \text{ cm}$$

From Eq. (3.35c)

$$\varepsilon_{\text{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[1 + \frac{12h}{W}\right]^{-1/2} = 2.14$$

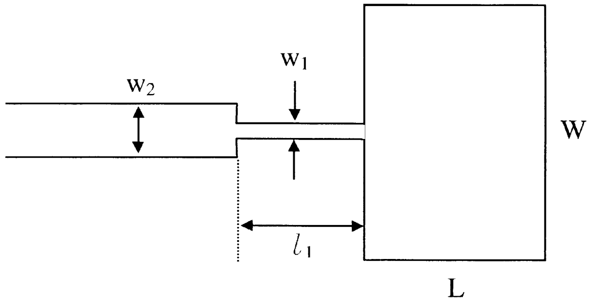


FIGURE 3.27 Design of a rectangular patch antenna.

From Eq. (3.35d)

$$\Delta l = 0.412h\left(\frac{\varepsilon_{\text{eff}} + 0.3}{\varepsilon_{\text{eff}} - 0.258}\right)\left(\frac{W/h + 0.264}{W/h + 0.8}\right) = 0.04 \text{ cm}$$

From Eq. (3.40)

$$L = \frac{c}{2f_r\sqrt{\varepsilon_{\text{eff}}}} - 2\,\Delta l = 3.34 \text{ cm}$$

Since $W < \lambda_0$, $G = W^2/90\lambda_0^2$ from (3.34),

$$Y_{\text{in}} = 2G = \frac{1}{45}\frac{W^2}{\lambda_0^2} = \frac{1}{R_{\text{in}}}$$
$$R_{\text{in}} = 288\ \Omega = \text{input impedance}$$

The characteristic impedance of the transformer is

$$Z_{0T} = \sqrt{R_{\text{in}} \times 50} = 120\ \Omega$$

From Figs. 2.24 and 2.25,

$$\frac{w_1}{h} \approx 0.58 \quad \text{and} \quad \frac{\lambda_{g1}\sqrt{\varepsilon_r}}{\lambda_0} \approx 1.13$$

Therefore, the impedance transformer has a width W_1 and length l_1 given by

$$w_1 = 0.0442 \text{ cm} \qquad l_1 = \tfrac{1}{4}\lambda_{g1} = 1.90 \text{ cm}$$

For a 50-Ω line w_2 can be estimated from Fig. 2.24 and found to be 0.228 cm.

Similar results can be obtained by using Eqs. (2.84) and (2.86); we have $w_1 = 0.044$ cm, $\varepsilon_{\text{eff}} = 1.736$, and $l_1 = 1.89$ cm. For the 50-Ω line, $w_2 = 0.238$ cm from Eq. (2.84) ■

Patch antennas can be fed by many different ways [13–16]. The most common feed methods are the coaxial probe feed and microstrip line edge feed, as shown in Fig. 3.28. The probe feed method is simple but is not attractive from the fabrication point of view. The edge feed method has the advantage that both patch and feed line can be printed together. Other feed methods are electromagnetically coupled microstrip line feed, aperture-coupling feed, slot line feed, coplanar waveguide feed, and so on.

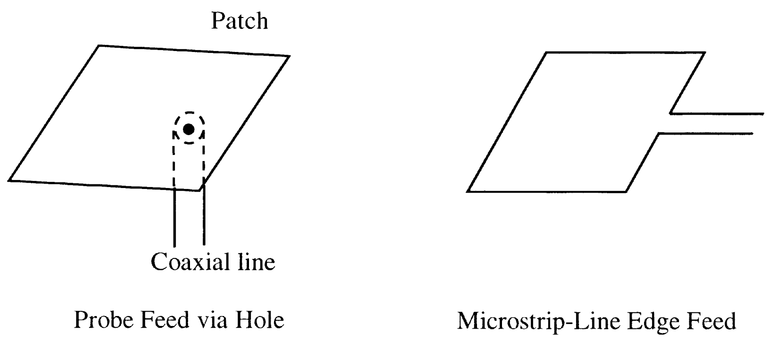


FIGURE 3.28 Microstrip patch antenna feed methods.

The radiation patterns of a microstrip patch antenna can be calculated based on electromagnetic analysis [13–16]. Typical radiation patterns are shown in Fig. 3.29 [17]. Typical half-power beamwidth is 50°–60°, and typical gain ranges from 5 to 8 dB.

In many wireless applications, CP antennas are required. A single square patch can support two degenerate modes at the same frequency with the radiated fields linearly polarized in orthogonal directions. Circular polarization can be accomplished by using a proper feed network with a 90° hybrid coupler, as shown in Fig. 3.30*a*. Other methods can also be employed to obtain circular polarization by perturbing the patch dimensions without the need of the hybrid coupler feed network. Some of these methods are shown in Fig. 3.30*b*.

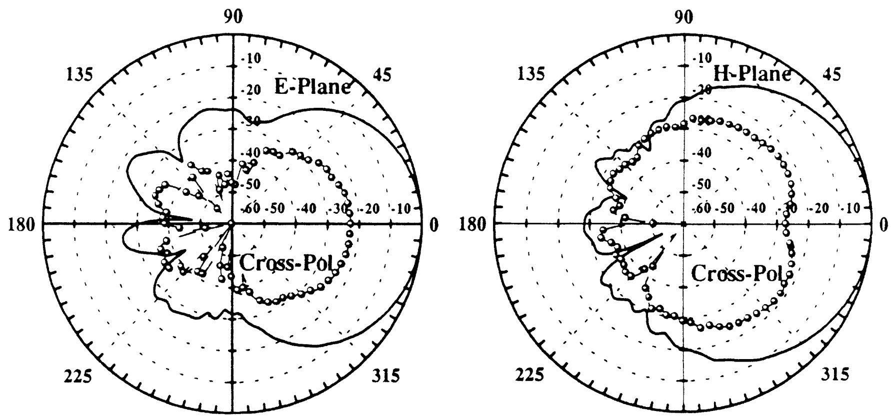


FIGURE 3.29 Radiation patterns of an inverted microstrip patch antenna. (From reference [17], with permission from IEEE.)

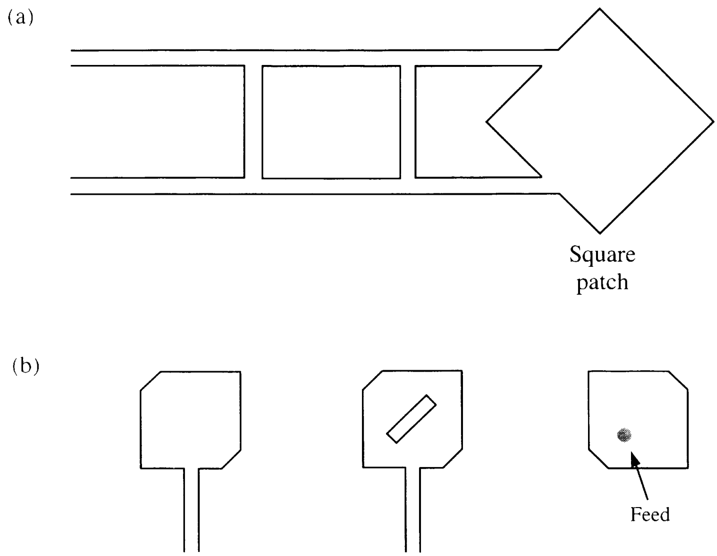


FIGURE 3.30 Circularly polarized square patch antennas.

3.10 ANTENNA ARRAYS AND PHASED ARRAYS

Single antennas are often limited for many applications because of a large HPBW and, consequently, a lower gain. For many applications, a high-gain, narrow pencil beam is required. Since most antennas have dimensions that are on the order of one wavelength, and since beamwidth is inversely proportional to antenna size, more than one antenna is required to sharpen the radiation beam. An array of antennas working simultaneously can focus the reception or transmission of energy in a particular direction, which increases the useful range of a system.

Considering the one-dimensional linear array shown in Fig. 3.31, the radiated field from a set of sources can be given by

$$E_{\text{total}} = I_1 f_1(\theta, \phi)\rho_1 \frac{e^{-j(k_0 r_1 - \Phi_1)}}{4\pi r_1} + I_2 f_2(\theta, \phi)\rho_2 \frac{e^{-j(k_0 r_2 - \Phi_2)}}{4\pi r_2} + \cdots + I_i f_i(\theta, \phi)\rho_i \frac{e^{-j(k_0 r_i - \Phi_i)}}{4\pi r_i} + \cdots \tag{3.42}$$

where I_i, ρ_i, and ϕ_i are the ith element's magnitude, polarization, and phase, respectively; $f_i(\theta, \phi)$ is the radiation pattern of the ith element and r_i is the distance from the ith element to an arbitrary point in space; and k_0 is the propagation constant, equal to $2\pi/\lambda_0$.

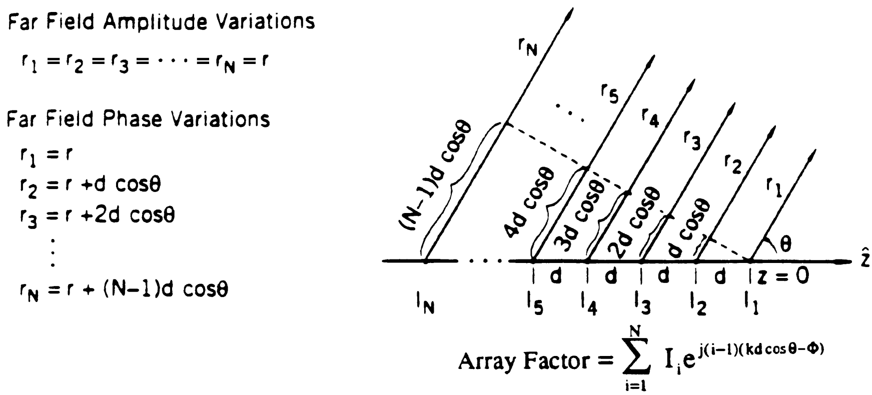


FIGURE 3.31 An N-element linear array along z axis.

Typically, the polarization of every element is aligned for copolarization (i.e., $\rho_i \approx \rho = 1$). The array has N elements with uniform spacing d. It is oriented along the z axis with a phase progression Φ. The first element is placed at the origin, and the distance r_i in the phase term is approximated using the following:

$$\begin{aligned} r_1 &\cong r \\ r_2 &\cong r + d\cos\theta \\ &\vdots \\ r_N &\cong r + (N-1)d\cos\theta \end{aligned} \tag{3.43}$$

These approximations allow the total field to be given by

$$\begin{aligned} E_{\text{total}} &= f(\theta,\phi)\frac{e^{-jk_0 r}}{4\pi r}\sum_{i=1}^{N} I_i e^{-j(i-1)(k_0 d\cos\theta-\Phi)} \\ &= \text{element pattern} \times \text{array factor} \end{aligned} \tag{3.44}$$

The total field from the array described by the equations above is made up of an element pattern $f(\theta,\phi)(e^{-jk_0 r}/4\pi r)$ and the array factor (AF). This is known as pattern multiplication.

Now let us consider beam scanning. The AF can be formulated without consideration of the type of element used in the array. For the sake of convenience we will assume the array elements are isotropic radiators and $I_i = 1$. For an array such as the one shown in Fig. 3.32, the AF is given as

$$\text{AF} = 1 + e^{-j(k_0 d\cos\theta-\Phi)} + e^{-j2(k_0 d\cos\theta-\Phi)} + \cdots + e^{-j(N-1)(k_0 d\cos\theta-\Phi)} \tag{3.45a}$$

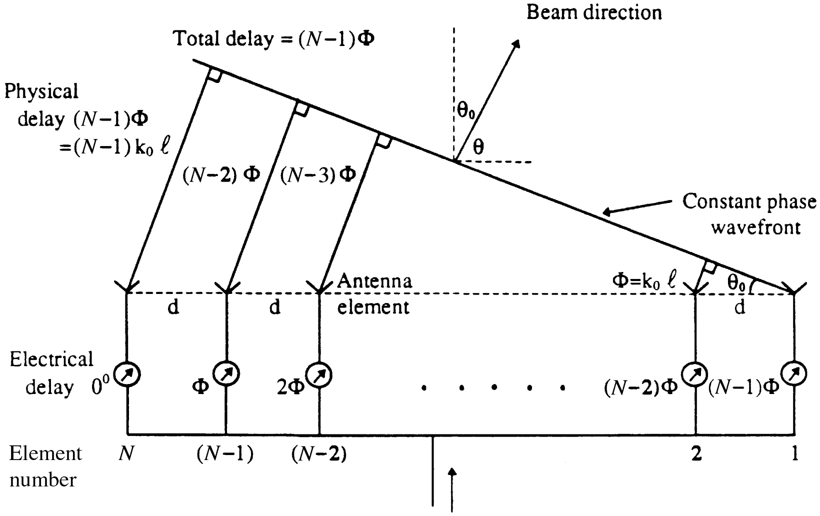


FIGURE 3.32 An N-element linear array with progressively larger phase delay from left to right.

or

$$\mathrm{AF} = \sum_{n=0}^{N-1} e^{-jn\psi} \tag{3.45b}$$

where $\psi = k_0 d \cos\theta - \Phi$.

The parameter Φ is the progressive phase shift across the array, which means that there is a phase difference of Φ between the currents on adjacent elements. The progressive phase shift causes the radiation emitted from the array to have a constant phase front that is pointing at the angle θ_0 (where $\theta_0 = 90^\circ - \theta$), as seen in Fig. 3.32. By varying the progressive phase shift across the array, the constant phase front is also varied.

To see how this is accomplished, note that the array factor in (3.45b) is a maximum when the exponential term equals 1. This happens when $\psi = 0$ or when

$$\Phi = k_0 d \sin(\theta_0) \tag{3.46a}$$

or

$$\theta_0 = \text{scanning angle} = \sin^{-1}\left(\frac{\Phi}{k_0 d}\right) \tag{3.46b}$$

As Φ is varied, it must satisfy (3.46) in order to direct the constant phase front of the radiation at the desired angle θ_0 (scanning the main beam). This is the basic concept used in a phased array.

Alternatively, Eq. (3.46) can be derived by examining Fig. 3.32. For each element, at the constant phase wavefront, the total phase delay should be the same for all elements. The total phase delay equals the summation of the electrical phase delay due to the phase shifter and the physical phase delay. From any two neighboring elements, elements 1 and 2 for example, we have

$$l = d \sin \theta_0 \tag{3.47}$$

Therefore,

$$\Phi = k_0 l = k_0 d \sin \theta_0 \tag{3.48}$$

Equation (3.48) is the same as (3.46a).

For an array antenna, the beamwidth and gain can be estimated from the number of elements. With the elements spaced by half-wavelengths to avoid the generation of grating lobes (multiple beams), the number of radiating elements N for a pencil beam is related to the half-power (or 3-dB) beamwidth by [18]

$$N \approx \frac{10{,}000}{(\theta_{BW})^2} \tag{3.49}$$

where θ_{BW} is the half-power beamwidth in degrees. From (3.49), we have

$$\theta_{BW} \approx \frac{100}{\sqrt{N}}$$

The corresponding antenna array gain is

$$G \approx \eta \pi N \tag{3.50}$$

where η is the aperture efficiency.

In a phased array, the phase of each antenna element is electronically controllable. One can change the phase of each element to make the array electronically steerable. The radiation beam will point to the direction that is normal to the constant phase front. This front is adjusted electronically by individual control of the phase of each element. In contrast to the mechanically steerable beam, the beam in a phased array can be steered much faster, and the antenna array is physically stationary.

The array factor is a periodic function. Hence, it is possible to have a constant phase front in several directions, called grating lobes. This can happen when the argument in the exponential in Eq. (3.45b) is equal to a multiple of 2π. To scan to a given angle, θ_0, as in Fig. 3.32, Φ must be chosen to satisfy $\Phi = k_0 d \sin(\theta_0)$ as before. Thus, $\psi = -2\pi = k_0 d(\cos\theta - \sin\theta_0)$. For the given scan direction, a grating lobe will begin to appear in the end-fire direction ($\theta = 180°$) when $-k_0 d[1 + \sin\theta_0] = -2\pi$. Dividing out 2π from the equation, we come up with the condition that

$$\frac{d}{\lambda_0} = \frac{1}{1 + \sin\theta_0} \tag{3.51}$$

Grating lobes reduce the array's ability to focus the radiation in a specific area of angular space (directivity) and are undesirable in the antenna pattern. The spacing between adjacent elements should be less than the distance defined in Eq. (3.51) to avoid grating lobes.

By adjusting the phase and spacing between elements in two dimensions, we can extend this theory to a two-dimensional array. Such an array would make scanning possible in two perpendicular directions. The array factor in (3.45b) is modified to

$$\mathrm{AF} = \sum_{m=0}^{M-1} e^{-jm(k_0 d_x \cos\theta\cos\phi - \Phi_x)} \sum_{n=0}^{N-1} e^{-jn(k_0 d_y \cos\theta\sin\phi - \Phi_y)} \tag{3.52}$$

where d_x and d_y define the element spacing in the x and y directions, respectively, and Φ_x and Φ_y are the progressive phase shifts in the x and y directions, respectively.

The arrays can be fed by two major methods: corporate feed method and traveling-wave feed method, as shown in Fig. 3.33. In the corporate feed method, 3-dB power dividers are used to split the input power and deliver it to each element.

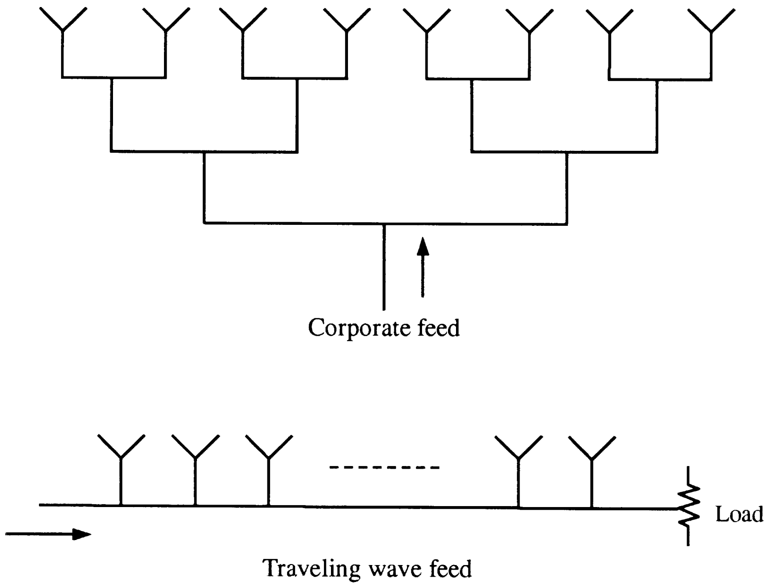


FIGURE 3.33 Antenna array feeding methods.

Each power splitter introduces some losses. In the traveling-wave feed method, antenna elements are coupled to a transmission line. The power coupled to each element is controlled by the coupling mechanisms.

Figure 3.34 shows a 16 × 16 microstrip patch antenna array with 256 elements using a corporate feed. A coaxial feed is connected to the center of the array from the other side of the substrate. Power dividers and impedance-matching sections are used to couple the power to each element for radiation.

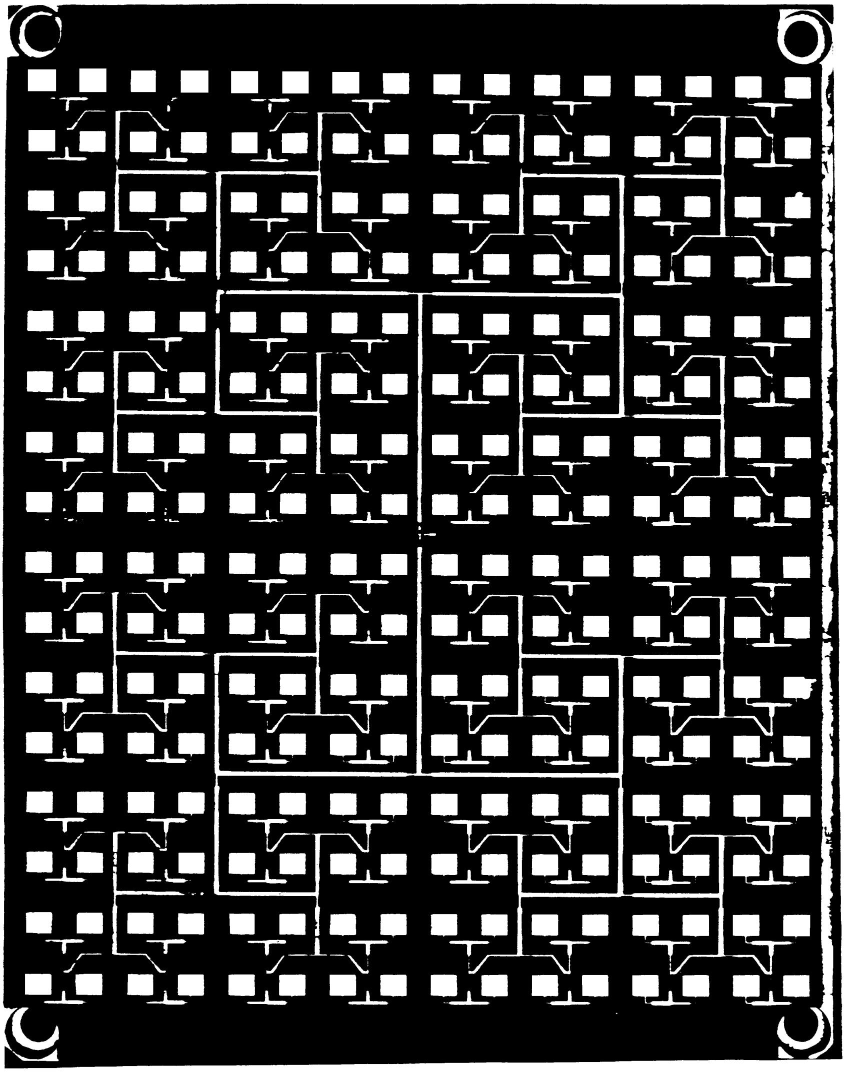


FIGURE 3.34 A 16 × 16 microstrip patch antenna array. (Courtesy of Omni-Patch Designs.)

3.11 ANTENNA MEASUREMENTS

Antennas can be measured in an indoor antenna chamber or an outdoor antenna range. Figure 3.35 shows a typical setup for the antenna testing [2]. The antenna under test (AUT) is located at the far-field region on top of a rotating table or positioner. A standard gain horn is used as a transmitting antenna, and the AUT is normally used as a receiving antenna. For the indoor range, the setup is placed inside an anechoic chamber. The chamber walls are covered with wave-absorbing material to isolate the AUT from the building structure and simulate the free-space unbounded medium.

The system is first calibrated by two standard gain horns. Parameters to be measured include antenna gain, antenna patterns, sidelobes, half-power beamwidth, directivity, cross-polarization, and back radiation. In most cases, external objects, finite ground planes, and other irregularities change the radiation patterns and limit the measurement accuracies.

Figure 3.36 shows a standard anechoic chamber for antenna measurement for far-field pattern measurements. Near-field and compact indoor ranges are also available for some special purposes.

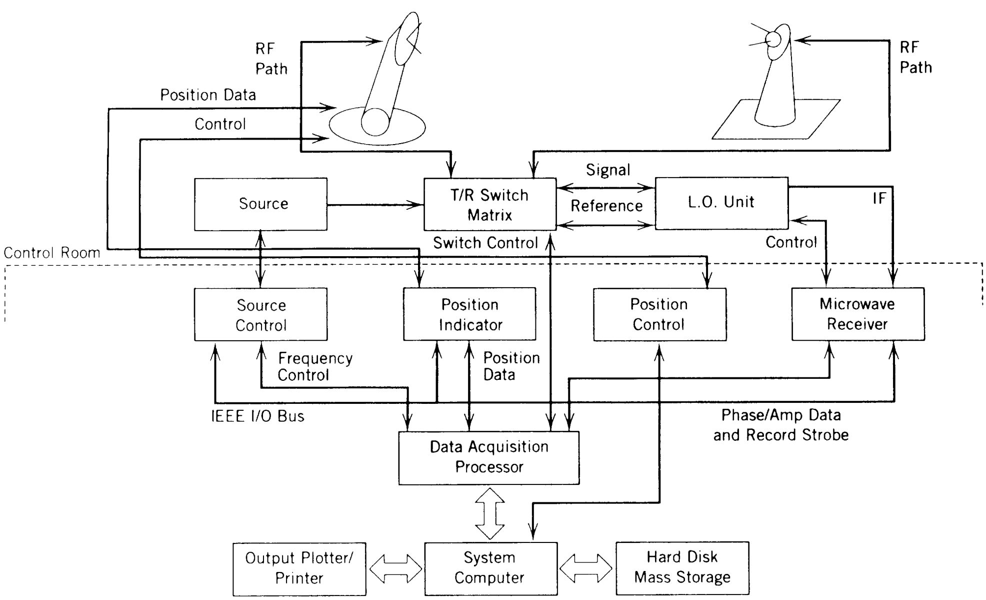


FIGURE 3.35 Typical antlenna range setup [2].

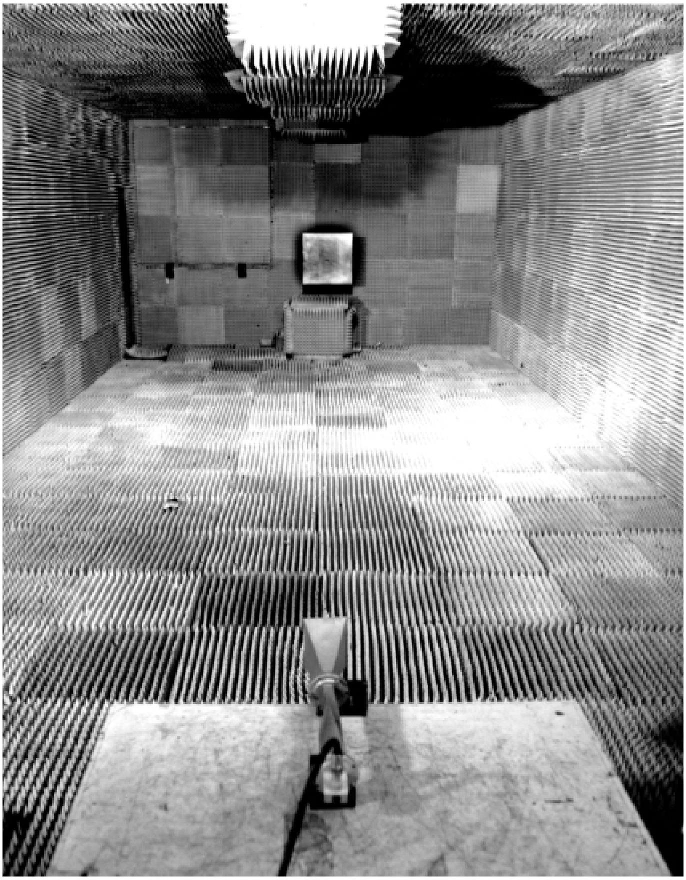


FIGURE 3.36 Standard indoor anechoic chamber at Texas A&M University.

PROBLEMS

3.1 Design a half-wave dipole at 1 GHz. Determine (a) the length of this dipole, (b) the length of the monopole if a monopole is used instead of a dipole, and (c) the power density (in millwatts per square centimeters) and rms E-field (in volts per meter) at a distance of 1 km if the transmit power $P_t = 10\,\text{W}$.

3.2 At 2 GHz, what is the length in centimeters of a half-wavelength dipole? What is the length of the equivalent monopole?

3.3 An FM radio station operating at 10 MHz transmits 1000 W using a monopole antenna, as shown in Fig. P3.3.

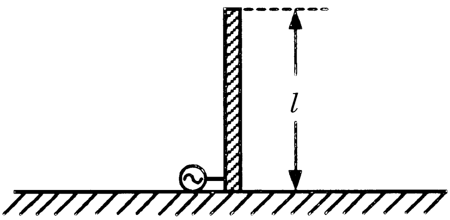


FIGURE P3.3

(a) Design l in meters such that the antenna is an equivalent half-wave dipole.

(b) The half-wave dipole has a gain of 1.64. What is the maximum rms electric field (in volts per meter) at a distance of 100 km away from the station?

3.4 A transmitter has a power output of 100 W and uses a dipole with a gain of 1.5. What is the maximum rms electric field at a distance of 1 km away from the transmitter?

3.5 A rectangular horn antenna has $A = 6$ in. and $B = 4$ in. operating at 12 GHz. Calculate (a) the antenna gain in decibels and (b) the maximum rms E-field at a distance of 100 m from the antenna if the transmit power is 100 W (E is in volts per meter).

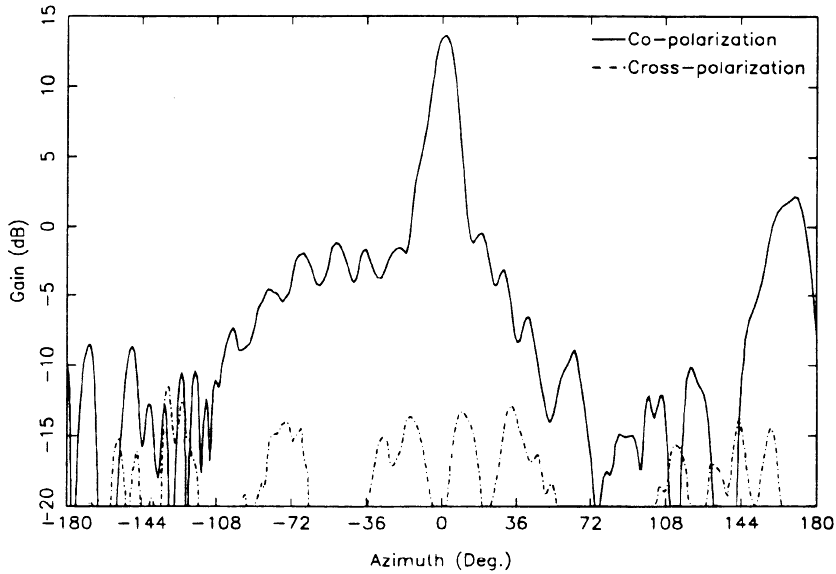
3.6 Calculate the gain (in decibels) and 3-dB beamwidth (in degrees) for a dish antenna with the following diameters at 10 GHz: (a) 5 ft and (b) 10 ft. Determine the far-field zones for both cases ($\eta = 55\%$).

3.7 At 10 GHz, design a dish antenna with a gain of 60 dB. What is the diameter of this antenna? What is the beamwidth in degrees? Note that this is a very high gain antenna ($\eta = 55\%$).

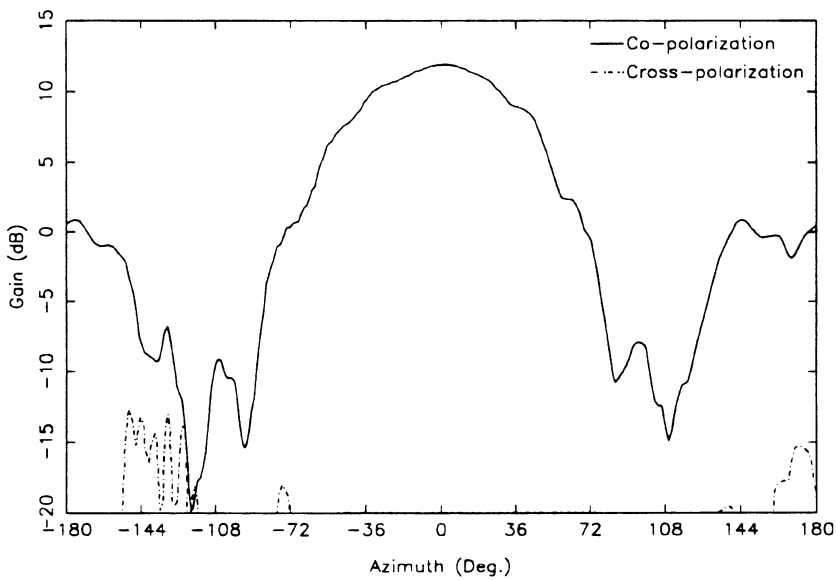
3.8 A dish antenna assuming a 55% illumination efficiency with a diameter of 3 m is attached to a transmitter with an output power of 100 kW. The operating frequency is 20 GHz. (a) Determine the maximum power density and electric (rms) field strength at a distance of 10 km away from this antenna. (b) Is this distance located at the far-field zone? (c) Is it safe to have someone inside the beam at this distance? (Note that the U.S. safety standard requires $P_{d,\max} < 10\ \mathrm{mW/cm^2}$.)

3.9 A high-power radar uses a dish antenna with a diameter of 4 m operating at 3 GHz. The transmitter produces 200 kW CW power. What is the minimum safe distance for someone accidentally getting into the main beam? (The U.S. standard requires power density $< 10\ \mathrm{mW/cm^2}$ for safety.)

3.10 A parabolic dish antenna has a diameter of 1 m operating at 10 GHz. The antenna efficiency is assumed to be 55%. (a) Calculate the antenna gain in decibels. (b) What is the 3-dB beamwidth in degrees? (c) What is the maximum power density in watts per square meter at a distance of 100 m away from the antenna? The antenna transmits 10 W. (d) What is the power density at 1.05° away from the peak?



E-plane radiation pattern of eight element array at X-band



H-plane radiation pattern of eight element array at X-band

FIGURE P3.13

3.11 A parabolic dish antenna has a diameter of 2 m. The antenna is operating at 10 GHz with an efficiency of 55%. The first sidelobe is 20 dB below the main-beam peak. (a) Calculate the antenna gain in decibels and the antenna's effective area in square meters. (b) Calculate the maximum power density in watts per square meter and E-field in volts per meter at a distance of 1 km away from the antenna. The power transmitted by the antenna is 100 W. (c) Compute the half-power beamwidth in degrees. (d) Calculate the power density in watts per square meter at the first-sidelobe location.

3.12 A direct-TV parabolic dish antenna has a diameter of 12 in. operating at 18 GHz with an efficiency of 60%. Calculate (a) the antenna gain in decibels, (b) the effective area in square centimeters, and (c) the half-power beamwidth in degrees. (d) Sketch the antenna pattern.

3.13 The E- and H-plane patterns shown in Fig. P3.13 are obtained from a dielectric waveguide fed microstrip patch array [4]. The array consists of 1×8 elements and operates at 10 GHz. Determine (a) the beamwidths for both patterns, (b) the maximum sidelobe levels for both patterns, (c) the backscattering levels for both patterns, and (d) the cross-polarization levels for both patterns.

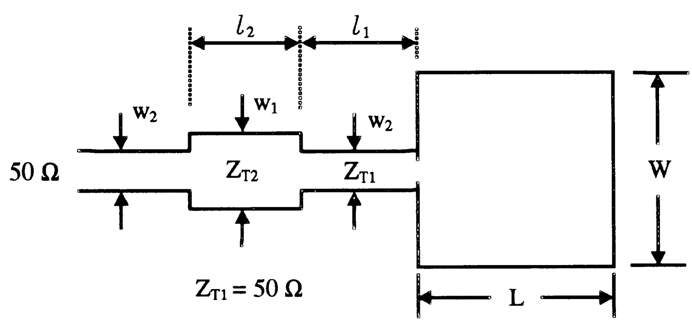


FIGURE P3.14

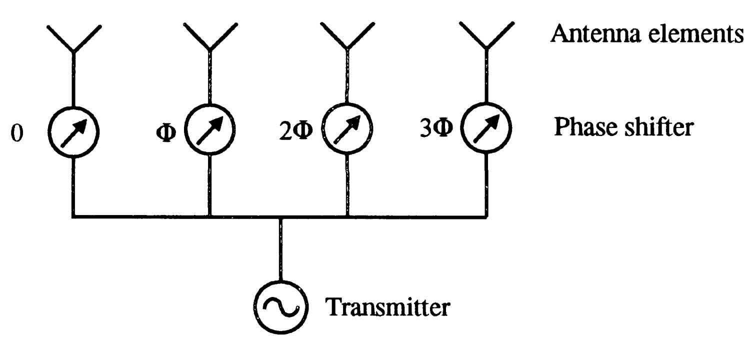


FIGURE P3.15

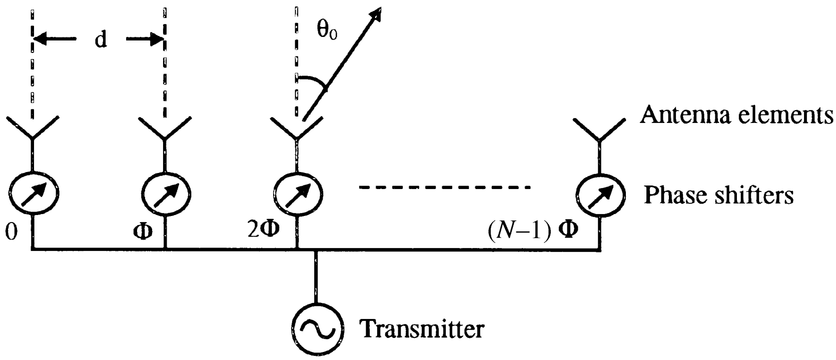


FIGURE P3.16

3.14 In the design of a matching circuit for a patch antenna in Example 3.4, since W_1 is small and a narrow microstrip line is lossy, an alternative design is to use a quarter-wavelength 50-Ω line followed by a quarter-wavelength low-impedance line as shown in Fig. P3.14. Determine the dimensions for W_1, W_2, l_1 and l_2 (see figure).

3.15 A four-element antenna array is shown in Fig. P3.15 with a phase shifter on each element. The separation between neighboring elements is $\lambda_0/2$. What is the scan angle θ_0 when Φ is equal to 90 degrees?

3.16 An N-element phased array has an antenna element separation d of $0.75\lambda_0$ (see Fig. P3.16). What are the scan angles when the progressive phase shift Φ equals 45 degrees and 90 degrees?

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3 장 용어 해설

RF (radio frequency): 전파

microwave: 마이크로파, 초고주파

receiving, receive, receiver, Rx: 수신, 수신기

transmitting, transmit, transmitter, Tx: 송신

free space: 자유공간, 진공으로 채워진 무한히 큰 공간

transmission line: 전송선

radiation: 전파(파동의) 방사, 파동이 공간으로 퍼져 나가는 현상

loss: 전력, 신호의 손실

wire antenna: 도선 안테나, 가는 금속 도선으로 구현된 안테나

aperture antenna: 개구 안테나, 뚫린 구멍을 통해 전파가 방사되는 안테나

printed antenna: 인쇄회로 기판 형태의 안테나

gain: 안테나 이득

omnidirectional antenna: 등방성 안테나, 수평면에서 모든 방향으로 같은 세기로 방사하는 안테나

beamwidth: 안테나 빔폭, 안테나의 전파가 대부분이 방사되는 각도 범위

one-port device: 입력단자가 1 개 (1 개의 전송선)인 장치

isotropic: 등방성, 모든 방향으로 동일한 세기로 방사하는

wavefront: 파면, 전파가 동시에 도달하는 면, 구면, 원통면, 평면으로 분류

power density: 전력밀도, 안테나에서 방사된 단위면적당 전력 (W/m^2)

magnetic field: 자기장, H (A/m)

electric field: 전기장, E (V/m)

permeability: 투자율, μ = 특정 물질의 투자율, μ_0 = 진공의 투자율 = $4\pi \cdot 10^{-7}$ H/m

permittivity: 유전율, ε = 특정 물질의 유전율, ε_0 = 진공의 유전율 = $8.854 \cdot 10^{-12}$ F/m

angular frequency: 각속도 (ω), $\omega = 2\pi f$ (rad/s) propagation constant: 전파상수, $k_0 = 2\pi / \lambda_0$ = 진공에서의 전파상수,

$\lambda_0 = 3 \cdot 10^8 / f$ = 진공에서의 파장

intrinsic impedance: 고유 임피던스, 특정 물질 내에서 전기장을 자기장으로 나눈 값

far-field region: 원거리 영역, 안테나로부터 충분히 떨어진 위치, 원거리에서 안테나의 방사전력의 관측각도에 따른 변화는 거리에 관계 없다.

spherical wave: 구면파, 한 점에서 모든 방향으로 동시에 방사되어 생성되는 파동

point source: 한 점에서 방사하는 이상적인 방사체

spherical coordinates: 구좌표계, (r, θ, φ), r = 거리, θ : 수직 회전각, φ : 수평 회전각

scattering parameter: 산란계수, 어떤 장치에서 반사파, 투과파를 입사파로 나눈 값, S_{11}, S_{21}, S_{12}, S_{22}

reflection coefficient: 반사계수, 어떤 장치에서 반사파 전압을 입사파 전압으로 나눈 값. Γ, S_{11}

impedance mismatch: 임피던스 부정합

VSWR (voltage standing wave ratio): 전압 정재파비, 전송선 상에서 최대 전압을 최소 전압으로 나눈 값

RL (return loss): 전송선 입력단에서 입사파를 반사파로 나눈 후 10 을 base 로 log 를 취한 후 20 을 곱한 값

input impedance: 전송선 입력단에서의 임피던스

characteristic impedance: 전송선의 특성임피던스, 전송선에서 부하 방향으로 진행하는 전압파동을 부하방향으로 진행하는 전류파동으로 나눈 값

bandwidth: 대역폭, 안테나가 요구되는 성능 규격을 만족하는 주파수 범위

impedance match (VSWR) bandwidth: 입력단 VSWR 이 2 이하인 주파수 범위, VSWR 이 2 이면 약 입사전력의 약 10%가 반사된다.

axial ratio bandwidth: 안테나에서 방사된 전파의 축비가 3 dB 이하인 주파수 범위

radiation pattern: 방사패턴, 안테나에서 방사된 전력의 밀도를 각도에 따라 표현한 것

E-plane pattern: 안테나의 방사패턴 중에서 전기장과 평행한 면 상에서의 방사패턴

H-plane pattern: 안테나의 방사패턴 중에서 자기장과 평행한 면 상에서의 방사패턴

co-polarization: 주편파, 방사된 파동 전기장 벡터 성분 중에서 가장 큰 성분

cross-polarization: 주편파 전기장 벡터와 수직인 전기장 벡터 성분

half-power beamwidth (BW, HPBW): 빔폭, 반치각, 각도에 따른 안테나의 방사전력 세기가 최대값에서 1/2 로 감소하는 각도 범위

SLL (sidelobe level): 데시벨 단위의 이득에서 주엽(main lobe) 최대 이득 값과 부변(sidelobe)의 최대 이득 값의 차이

directivity: 지향도, 각도에 따라 방사전력 밀도의 각도에 따른 변화

(maximum) directivity: (최대) 지향도, 그냥 지향도라 하면 최대 지향도를 의미

gain: 안테나 이득, 지향도에 안테나 효율을 곱한 값

efficiency: 안테나 효율, 안테나의 방사전력을 입력전력으로 나눈 값

polarization: 안테나의 편파, 전기장 벡터의 시간에 따른 궤적

LP (linear polarization): 전기장 벡터가 직선상에서 변화, 수직편파, 수평편파, 45° 편파

CP (circular polarization): 전기장 벡터의 변화궤적이 원, 우원편파, 좌원편파

effective area: 안테나의 유효면적, 수신 안테나에 입사된 전파의 전력밀도와 수신 안테나의 유효면적을 곱하면 수신 안테나 출력단에서의 수신된 전력이 된다.

Friis transmission formula: 1 개 안테나는 송신하고 다른 안테나는 수신할 경우 수신전력을 구하는 공식

half-wave dipole antenna: 반파장 다이폴 안테나, 가장 단순하며 기초적인 안테나, 길이가 약 0.45 파장인 도선의 중간을 절단 후 급전선을 연결한 안테나, 편파 순도가 높고 구조가 간단하여 표준 안테나로 사용됨. 이득은 2.2 dB 이며 입력 임피던스는 73 Ω 이다.

array antenna, antenna array: 배열 안테나, 안테나 여러 개를 일정한 규칙에 따라 배치한 후 이를 급전 회로망으로 연결하여 이득을 높인 안테나

phased array antenna: 배열 안테나에 있어서 각 소자의 위상을 전자적으로 조정하여 고속으로 안테나 main beam 의 위치를 변화시키는 안테나, 첨단 통신, 군용 레이더 등에 사용

3 장 문제풀이 시 참고사항

1. 단위

○ 모든 계산공식 사용시 MKS (meter, kilogram, second) 단위의 물리량 사용

전압: V (V, volt)

전류: I (A, ampere)

임피던스: Z (Ω, ohm)

전기장: E (V/m)

자기장: H (A/m)

전력: P (W, watt)

주파수: f (Hz)

파장: λ (m)

2. dB 단위

○ 모든 공식에는 특별히 표시되는 않은 한 linear 단위 (dB 값 취하기 전의 단위) 사용

$$P_s = S = \frac{P_t}{4\pi R^2} G_t$$

$P_s = S$: 전력밀도 (W/m^2)

P_t : 송신전력 (W)

R : 송신 안테나로부터 관측점까지의 거리 (m)

G_t : 송신 안테나 이득 (단위 없음, 비율 값)

○ dB 단위

1) 전력 또는 전력단위의 물리량: 10 을 베이스로 log 를 취한 후 10 을 곱한다.

P (dBW) = 10log10[P (W)]

예: 20 W = 10log$_{10}$(20 W/ 1W) = 13 dBW

예: 50 dBW = $10^{50/10} = 10^5$ W

P (dBm) = 10log10[P (mW)]

예: 0.02 W = $10\log_{10}$(20 mW/1 mW) = 13 dBm

예: -115 dBm = $10^{-115/10} = 10^{-11.5} = 3.2\cdot10^{-12}$ mW

2) 전압, 전류, 전기장, 자기장: 10 을 베이스로 log 를 취한 후 20 을 곱한다.

V(dBμV) = $20\log_{10}$[V(μV)]

예: 0.016 V = $1.6\cdot10^{4}\,\mu$V = $20\log_{10}(1.6\cdot10^{4})$ = 84.1 dBμV

예: 50 dBμV = $10^{50/20}$ = 316.2 μV

3. 예제

5 dBm 전력을 4 dB 이득 안테나로 방사할 경우 1.2 km 에서 전력밀도 (W/m^2)와 전기장 세기 (V/m)를 구하라.

(해답)

1) 전력밀도

$$S = \frac{P_t}{4\pi R^2} G_t$$

$P_t = 10^{5/10}$ mW = 3.16 mW = 0.00316 W

R = 1.2 km = 1200 m

$G_t = 10^{4/10} = 2.51$

$S = 0.00316\times2.51 / [4\cdot3.14\cdot(1200)^2] = 4.44\cdot10^{-10}$ W

2) 전기장 세기

$$S = \frac{1}{2}\frac{E^2}{377}$$

$E = (4.44\cdot10^{-10}\times2\times377)^{1/2} = 5.79\cdot10^{-4}$ V/m