5.1.1 The Electric (Hertzian) Dipole

The Hertzian dipole consists of an infinitesimal current element of length dl carrying a phasor current $\hat{I}$ that is assumed to be the same (in magnitude and phase) at all points along the element length, as illustrated in Fig. 5.1. A spherical coordinate system is commonly used to describe antennas. The location of a point in this coordinate system is described by the radial distance to the point, r, and the angular positions of a radial line to the point from the z axis, θ, and between its projection on the xy plane and the x axis, ϕ, as shown in Fig. 5.1. Orthogonal unit vectors $\vec{a}_r$, $\vec{a}_\theta$, $\vec{a}_\phi$, point in the directions of increasing values of these coordinates. The components of the magnetic field intensity fector become [1]

$$\hat{H}_r = 0 \tag{5.1a}$$

$$\hat{H}_\theta = 0 \tag{5.1b}$$

$$\hat{H}_\phi = \frac{\hat{I}\,dl}{4\pi}\beta_0^2 \sin\theta\left(j\frac{1}{\beta_0 r} + \frac{1}{\beta_0^2 r^2}\right)e^{-j\beta_0 r} \tag{5.1c}$$

Similarly, the components of the electric field intensity vector are [1]

$$\hat{E}_r = 2\frac{\hat{I}\,dl}{4\pi}\eta_0\beta_0^2 \cos\theta\left(\frac{1}{\beta_0^2 r^2} - j\frac{1}{\beta_0^3 r^3}\right)e^{-j\beta_0 r} \tag{5.2a}$$

$$\hat{E}_\theta = \frac{\hat{I}\,dl}{4\pi}\eta_0\beta_0^2 \sin\theta\left(j\frac{1}{\beta_0 r} + \frac{1}{\beta_0^2 r^2} - j\frac{1}{\beta_0^3 r^3}\right)e^{-j\beta_0 r} \tag{5.2b}$$

$$\hat{E}_\phi = 0 \tag{5.2c}$$

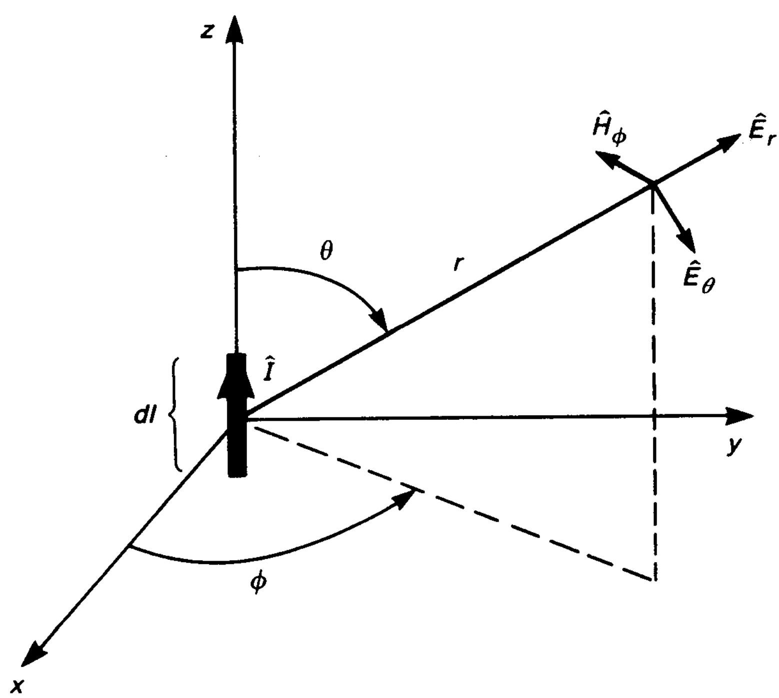

FIGURE 5.1 The electric (Hertzian) dipole.

where $\eta_0 = \sqrt{\mu_0/\epsilon_0}$ is the intrinsic impedance of free space. Note that the fields in (5.1) and (5.2) can be viewed as functions of *electrical distance from the antenna,* since $\beta_0 r = 2\pi r/\lambda_0$ and $\lambda_0 = v_0/f$ is the wavelength at the frequency of the antenna current.

The complete fields in (5.1) and (5.2) are quite complicated [2]. Our main interest is in the *far fields* where the field point is sufficiently distant from the antenna. How far is "far enough?" Note that the $1/r^3$ and $1/r^2$ terms dominate at very close distances to the antenna. As we move further from the antenna, the $1/r$ terms begin to dominate. The point where the $1/r^3$ and $1/r^2$ terms become insignificant compared with the $1/r$ terms is referred to the boundary between the *near field* and the *far field.* This occurs where $r = \lambda_0/2\pi \cong \frac{1}{6}\lambda_0$. The reader is cautioned that *the boundary between the near and far fields for other antennas is not simply $\lambda_0/2\pi$,* as is frequently assumed. A more realistic choice for the boundary between the near and far fields will be discussed later, but can be summarized as being the larger of $3\lambda_0$ or $2D^2/\lambda_0$, where D is the largest dimension of the antenna. Typically the first criterion is used for "wire-type" antennas and the second for "surface-type" antennas such as parabolics or horns. This boundary between the near and far fields is not meant to be a precise criterion, but is only intended to indicate a general region where the fields transition from complicated to quite simple structure. In the use of antennas for communication this question of whether the receiving antenna is in the near or far field of the transmitting antenna never arises because these antennas are used for communication over large distances. However, in the area of EMC and interference caused by emissions the receiver (which may be an intentional antenna used for compliance measurement) is frequently in the near field of the transmitting antenna (which may be the product). This is particularly true for the lower frequencies of the FCC, Class B radiated emission measurement, and is investigated in detail in [2]. Nevertheless, for the moment we will assume that the field point is in the far field of the Hertzian dipole. Retaining only the $1/r$ terms in the field expressions in (5.1) and (5.2) gives the *far-field vectors*:

$$\vec{\hat{E}}_{\text{far field}} = j\eta_0\beta_0\frac{\hat{I}\,dl}{4\pi}\sin\theta\frac{e^{-j\beta_0 r}}{r}\vec{a}_\theta \tag{5.3a}$$

$$\vec{\hat{H}}_{\text{far field}} = j\beta_0\frac{\hat{I}\,dl}{4\pi}\sin\theta\frac{e^{-j\beta_0 r}}{r}\vec{a}_\phi \tag{5.3b}$$

The time-domain fields are obtained by multiplying the phasor fields by $e^{j\omega t}$ and taking the real part of the results:

$$\begin{aligned}\vec{E}_{\text{far field}} &= \mathscr{Re}\{\vec{\hat{E}}_{\text{far field}}e^{j\omega t}\} \qquad (5.4\text{a})\\ &= \frac{E_m}{r}\cos\left[\omega\left(t-\frac{r}{v_0}\right)+90^\circ\right]\vec{a}_\theta\\ &= -\frac{E_m}{r}\sin\left[\omega\left(t-\frac{r}{v_0}\right)\right]\vec{a}_\theta\end{aligned}$$

$$\vec{H}_{\text{far field}} = \mathcal{Re}\{\vec{\hat{H}}_{\text{far field}} e^{j\omega t}\} \tag{5.4b}$$
$$= \frac{E_m}{\eta_0 r} \cos\left[\omega\left(t - \frac{r}{v_0}\right) + 90^\circ\right] \vec{a}_\phi$$
$$= -\frac{E_m}{\eta_0 r} \sin\left[\omega\left(t - \frac{r}{v_0}\right)\right] \vec{a}_\phi$$

where

$$E_m = \frac{\eta_0 \beta_0 I\, dl}{4\pi} \sin\theta \tag{5.4c}$$

The far fields of the Hertzian dipole satisfy many of the properties of uniform plane waves. In fact, "locally" the fields resemble uniform plane waves, although they are more correctly classified as spherical waves. These properties are as follows:

1. The fields are proportional to $1/r$, $\hat{I}$, dl, and $\sin\theta$.
2. $|\vec{\hat{E}}_{\text{far field}}|/|\vec{\hat{H}}_{\text{far field}}| = \eta_0$.
3. $\vec{\hat{E}}_{\text{far field}}$ and $\vec{\hat{H}}_{\text{far field}}$ are *locally orthogonal.*
4. $\vec{\hat{E}}_{\text{far field}} \times \vec{\hat{H}}_{\text{far field}} = K\vec{a}_r$.
5. A phase term $e^{-j\beta_0 r}$ translates to a time delay in the time domain of $\sin[\omega(t - r/v_0)]$.

This is the origin of the technique of translating fields using the inverse-distance relationship. For example, the electric and magnetic fields at distances D_1 and D_2 are related by $|\vec{E}_{D_2}| = (D_1/D_2)|\vec{E}_{D_1}|$; that is, the fields decay inversely with increasing distance away from the radiator. It is important to remember that *the inverse-distance rule holds only if both D_1 and D_2 are in the far field of the radiating element.* This important restriction is frequently not adhered to. If either of the two distances are in the near field, the inverse-distance rule cannot be used.

We next obtain the total average power radiated by integrating the average power Poynting vector over a suitable closed surface surrounding the antenna. First we compute the Poynting vector using the *total* phasor fields in (5.1) and (5.2) as

$$\vec{S}_{\text{av}} = \tfrac{1}{2}\mathcal{Re}\{\vec{\hat{E}} \times \vec{\hat{H}}^*\} \tag{5.5}$$
$$= \tfrac{1}{2}\mathcal{Re}\{\hat{E}_\theta \hat{H}_\phi^* \vec{a}_r - \hat{E}_r \hat{H}_\phi^* \vec{a}_\theta\}$$
$$= 15\pi\left(\frac{dl}{\lambda_0}\right)^2 |\hat{I}|^2 \frac{\sin^2\theta}{r^2} \vec{a}_r \quad (\text{in W/m}^2)$$

This shows that average power is flowing away from the current element: our first hint of "radiation." It is instructive to note that this average power density vector could have been obtained solely from the far-field expressions given in (5.3). Integrating this result over a sphere of radius r enclosing the current element gives the total average power radiated by the current element [1]:

$$P_{\text{rad}} = \oint_S \vec{S}_{\text{av}} \cdot d\vec{s} \tag{5.6}$$

$$= 80\pi^2 \left(\frac{dl}{\lambda_0}\right)^2 \frac{|\hat{I}|^2}{2} \quad \text{(in W)}$$

Denoting $\hat{I}/\sqrt{2} = \hat{I}_{\text{rms}}$, we can compute a *radiation resistance*:

$$R_{\text{rad}} = \frac{P_{\text{rad}}}{|\hat{I}_{\text{rms}}|^2} \tag{5.7}$$

$$= 80\pi^2 \left(\frac{dl}{\lambda_0}\right)^2 \quad (\text{in } \Omega)$$

The radiation resistance represents a fictitious resistance that dissipates the same amount of power as that radiated by the Hertzian dipole when both carry the same value of rms current.

The Hertzian dipole is a very ineffective radiator. For example, for a length $dl = 1$ cm and a frequency of 300 MHz ($\lambda_0 = 1$ m), the radiation resistance is 79 mΩ. In order to radiate 1 W of power, we require a current of 3.6 A! If the frequency is changed to 3 MHz ($\lambda_0 = 100$ m), the radiation resistance is 7.9 μΩ and the current required to radiate 1 W is 356 A! Nevertheless the Hertzian dipole serves a useful purpose in that the *far fields of a Hertzian dipole are virtually identical to the far fields of most other practical antennas.*

5.1.2 The Magnetic Dipole (Loop)

A dual to the elemental electric dipole is the *elemental magnetic dipole or current loop* shown in Fig. 5.2. A very small loop of radius b lying in the xy plane carries a phasor current $\hat{I}$. This loop constitutes a magnetic dipole moment

$$\hat{m} = \hat{I}\pi b^2 \quad (\text{in A m}^2) \tag{5.8}$$

where πb^2 is the area enclosed by the loop. The radiated fields are [1]

$$\hat{E}_r = 0 \tag{5.9a}$$

$$\hat{E}_\theta = 0 \tag{5.9b}$$

$$\hat{E}_\phi = -j\frac{\omega\mu_0 \hat{m}\beta_0^2}{4\pi} \sin\theta \left(j\frac{1}{\beta_0 r} + \frac{1}{\beta_0^2 r^2}\right) e^{-j\beta_0 r} \tag{5.9c}$$

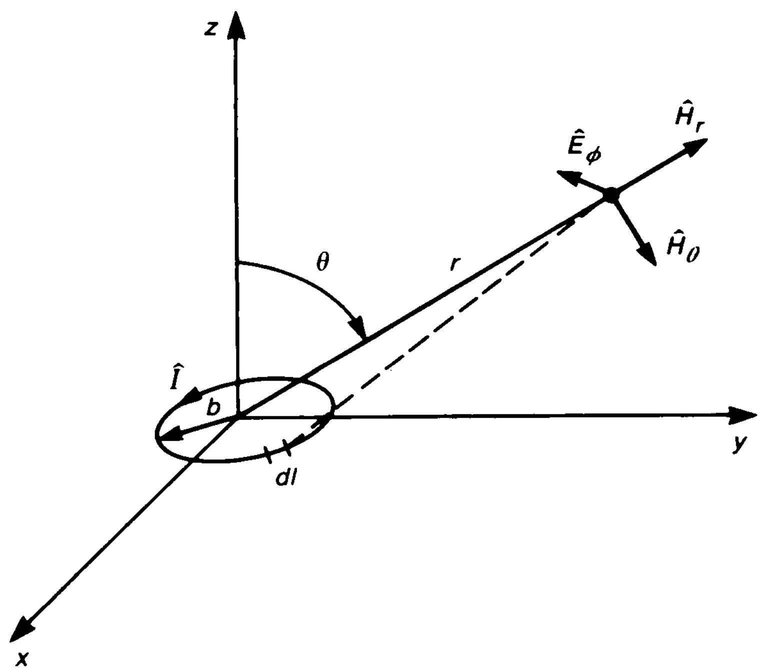

FIGURE 5.2 The magnetic dipole.

and

$$\hat{H}_r = j2\frac{\omega\mu_0\hat{m}\beta_0^2}{4\pi\eta_0}\cos\theta\left(\frac{1}{\beta_0^2 r^2} - j\frac{1}{\beta_0^3 r^3}\right)e^{-j\beta_0 r} \tag{5.10a}$$

$$\hat{H}_\theta = j\frac{\omega\mu_0\hat{m}\beta_0^2}{4\pi\eta_0}\sin\theta\left(j\frac{1}{\beta_0 r} + \frac{1}{\beta_0^2 r^2} - j\frac{1}{\beta_0^3 r^3}\right)e^{-j\beta_0 r} \tag{5.10b}$$

$$\hat{H}_\phi = 0 \tag{5.10c}$$

Comparing (5.9) and (5.10) with the fields of the Hertzian dipole given in (5.1) and (5.2), we see that duality exists between the fields of these structures. Observations about the electric (magnetic) field of the electric dipole apply to the magnetic (electric) field of the magnetic dipole.

The far field of the magnetic dipole is characterized by the $1/r$-dependent terms:

$$\vec{\hat{E}}_{\text{far field}} = \frac{\omega\mu_0\hat{m}\beta_0}{4\pi}\sin\theta\frac{e^{-j\beta_0 r}}{r}\vec{a}_\phi \tag{5.11a}$$

$$\vec{\hat{H}}_{\text{far field}} = -\frac{\omega\mu_0\hat{m}\beta_0}{4\pi\eta_0}\sin\theta\frac{e^{-j\beta_0 r}}{r}\vec{a}_\theta \tag{5.11b}$$

As was the case for the Hertzian (electric) dipole, the far field of the magnetic dipole is such that the fields (1) decay as $1/r$, (2) lie in a (local) plane perpendicular to the radial direction, and (3) are related by η_0.

As was done for the Hertzian dipole, we may determine a radiation resistance of the magnetic dipole, which becomes [1]

$$R_{\text{rad}} = \frac{P_{\text{av}}}{|\hat{I}_{\text{rms}}|^2} \tag{5.12}$$
$$= 31{,}170\left(\frac{A}{\lambda_0^2}\right)^2$$

where $A = \pi b^2$ is the area of the loop. Like the Hertzian dipole, the magnetic dipole is not an efficient radiator. Consider a loop of radius 1 cm. At 300 MHz the radiation resistance is 3.08 mΩ. In order to radiate 1 W, the loop requires a current of 18 A. At 3 MHz the radiation resistance is 3.08×10^{-11} Ω, and the current required to radiate 1 W is 1.8×10^5 A!

If a loop is electrically small, its shape is not important with respect to the far fields that it generates [1]. In order to illustrate the application of this result to radiated emissions, consider a 1 cm $\times$ 1 cm current loop on a PCB (an equivalent loop radius of 5.64 mm). Suppose the loop carries a 100 mA current at a frequency of 50 MHz. At a measurement distance of 3 m (FCC Class B) the electric field is a maximum in the plane of the loop and is $|\hat{E}| = 109.6$ μV/m = 40.8 dBμV/m. Recall that the FCC Class B limit from 30 MHz to 88 MHz is 40 dBμV/m. Therefore a 1 cm $\times$ 1 cm loop carrying a 50 MHz, 100 mA current will cause a radiated emission that will fail to comply with the FCC Class B regulatory limit! This should serve to illustrate to the reader that passing the regulatory requirements on radiated emissions is not a simple matter, since the above dimensions and current levels are quite representative of those found on PCBs of electronic products.

5.2 THE HALF-WAVE DIPOLE AND QUARTER-WAVE MONOPOLE

The Hertzian dipole considered in Section 5.1.1 is an obviously impractical antenna for several reasons. Primarily, the length of the dipole was assumed to be infinitesimal in order to simplify the computation of the fields. Also, the current along the Hertzian dipole was assumed to be constant along the dipole. This latter assumption required the current to be nonzero at the endpoints of the dipole—an unrealistic and, moreover, physically impossible situation since the surrounding medium, free space, is nonconductive. Also, the Hertzian dipole is a very inefficient radiator since the radiation resistance is quite small, requiring large currents in order to radiate significant power. The magnetic dipole suffers from similar problems. In this section we will consider two practical and more frequently used antennas: the long-dipole and monopole antennas.

The long-dipole antenna (or, simply, the dipole antenna) consists of a thin wire that is fed or excited via a voltage source inserted at the midpoint, as shown in Fig. 5.3(a). Each leg is of length $\frac{1}{2}l$.

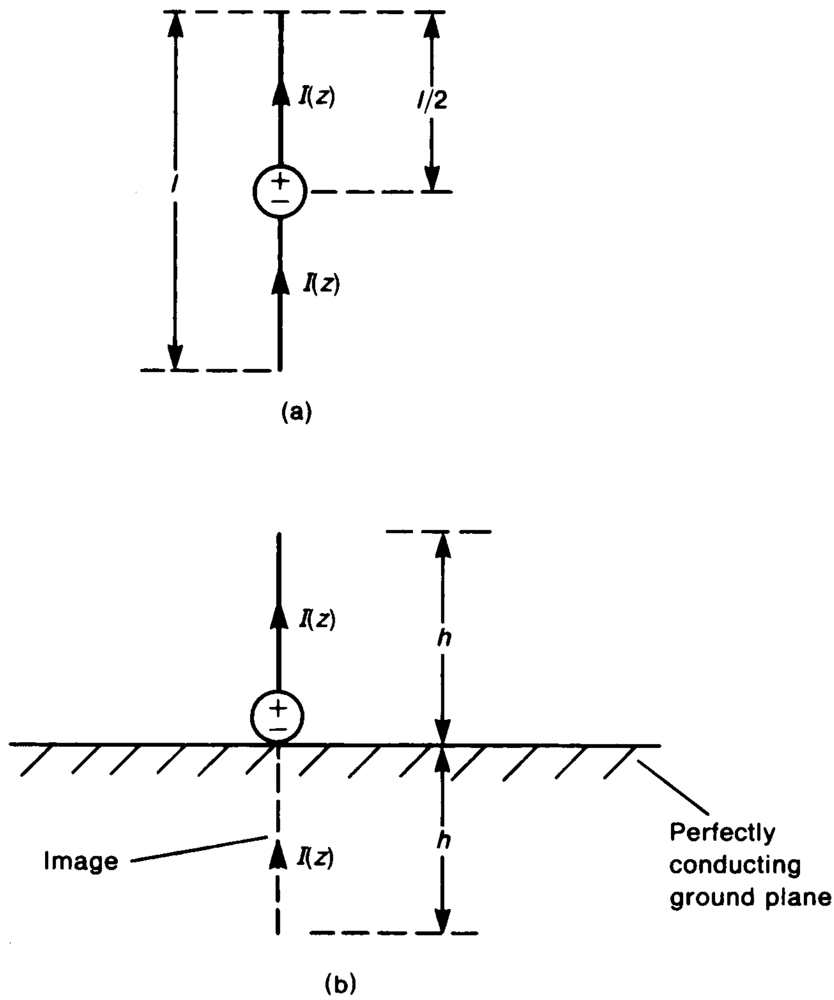

FIGURE 5.3 Illustration of (a) the dipole antenna and (b) the monopole antenna.

The monopole antenna shown in Fig. 5.3(b) consists of a single leg perpendicular to a ground plane. The monopole is fed at its base with respect to the ground plane. For purposes of analysis, the ground plane is considered to be infinite and perfectly conducting. In practice, this ideal ground plane is approximated. On aircraft, for example, the metallic fuselage simulates this ground plane. For ground-based stations the earth simulates, to some degree, this ground plane. Since the earth is much less of an approximation to a perfectly conducting plane that is metal, ground-based stations are usually augmented by a grid of wires lying on the ground to simulate the ground plane. The monopole antenna can be analyzed by replacing the ground plane with the image of the current element that is above the ground plane, as indicated in Fig. 5.3(b). Images are discussed in Section 5.6.1. Once the ground plane is replaced with the image, the problem reduces to the dipole problem so that a separate analysis of the monopole is not needed.

We observed previously that if we know the current distribution over the surface of an antenna, we may compute the resulting radiated fields. In practice, one often makes a reasonable guess as to the *current distribution* over the antenna surface. It can be shown that the current distribution of the long-dipole antenna follows (approximately) the same distribution as on a transmission line; that is, $\hat{I}(z)$ is proportional to sin $\beta_0 z$ [1]. Placing the center of the dipole at the origin of a spherical coordinate system as shown in Fig. 5.4(a), with the dipole directed along the z axis, we may therefore write an expression for the current distribution along the wire is

$$\hat{I}(z) = \begin{cases} \hat{I}_m \sin[\beta_0(\frac{1}{2}l - z)] & \text{for } 0 < z < \frac{1}{2}l \\ \hat{I}_m \sin[\beta_0(\frac{1}{2}l + z)] & \text{for } -\frac{1}{2}l < z < 0 \end{cases} \tag{5.13}$$

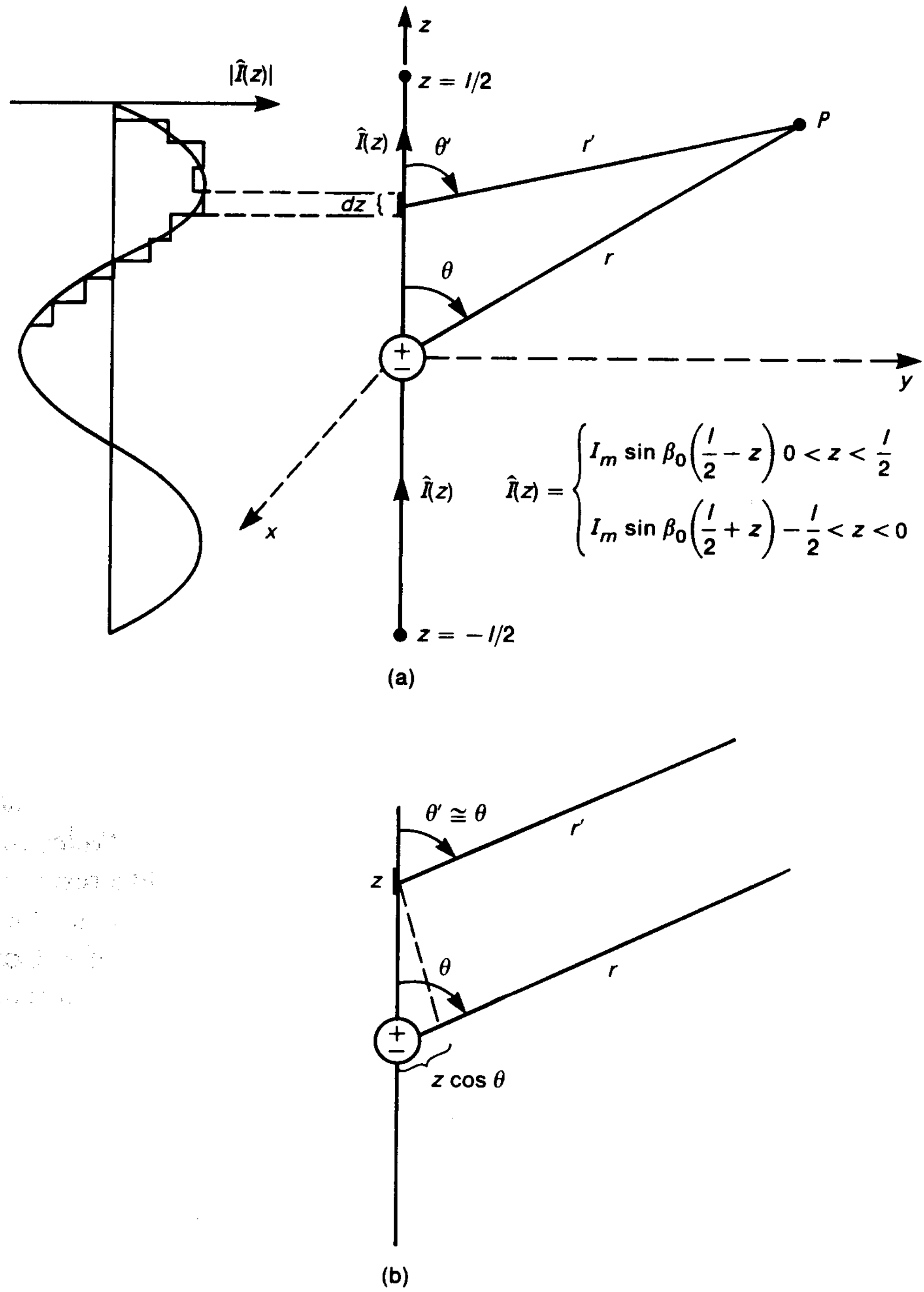

FIGURE 5.4 Computation of the radiated fields of the dipole antenna: (a) superposition of the Hertzian dipole fields; (b) the far-field approximation.

Note that this current distribution satisfies the necessary criteria: (1) the variation with z is proportional to $\sin \beta_0 z$ and (2) the current goes to zero at the endpoints, $z = -\frac{1}{2}l$ and $z = \frac{1}{2}l$.

Now that we have assumed the current distribution along the dipole, we may compute the fields of the long dipole as being the superpositions of the fields due to many small Hertzian dipoles of length dz having a current that is constant and equal to the value of $\hat{I}(z)$ at that point along the dipole, as shown in Fig. 5.4(a). We also assume that the desired field point is in the far field of these current elements, so that we only need the far-field expressions for the Hertzian dipole given in (5.3). For example, the field at point P in Fig. 5.4(a) due to the segment dz is

$$d\hat{E}_\theta = j\eta_0 \beta_0 \frac{\hat{I}(z) \sin \theta'}{4\pi r'} e^{-j\beta_0 r'}\, dz \tag{5.14}$$

Our ultimate objective is to obtain the fields at point P in terms of the radial distance r from the midpoint of the dipole and angle θ. Since we are considering only the far field, the radial distance r from the center of the dipole at point P and the distance r' from the current element to point P will be approximately equal ($r \cong r'$) and the angles θ and θ' will be approximately equal ($\theta \cong \theta'$), as shown in Fig. 5.4(b).

We may substitute $r' = r$ into the denominator of (5.14), but we should not make this substitution into the $e^{-j\beta_0 r'}$ term for the following reason. This term may be written as $e^{-j\beta_0 r'} = \underline{/2\pi r'/\lambda_0}$, and its value depends not on physical distance r' but on electrical distance r'/λ_0. Therefore, even though r' and r may be approximately equal, the term may depend significantly on the difference in electrical distances. For example, suppose $r = 1000$ m and $r' = 1000.5$ m and the frequency is $f = 300$ MHz. We have ($\lambda_0 = 1$ m) $\beta_0 r = 2\pi(1000) = 360{,}000°$ and $\beta_0 r' = 2\pi(1000.5) = 360{,}180°$. Note that the fields at 1000 m and those only 0.5 m farther away are 180° out of phase! We will see a more striking example of this in the next section, where the far fields from two antennas that are widely spaced physically but separated on the order of a wavelength may actually be completely out of phase and add destructively to yield a result of zero.

Thus it is not a reasonable approximation to substitute r for r' in the phase term in (5.14). However, we may still write the result in terms of r. Consider Fig. 5.4(b), which shows the two radial distances r and r' as being approximately parallel. Thus we are assuming that the field point is sufficiently far, physically, from the antenna. From Fig. 5.4(b) we may obtain

$$r' \cong r - z \cos \theta \tag{5.15}$$

Substituting (5.15) into the phase term in (5.14) and r into the denominator gives

$$d\hat{E}_\theta = j\eta_0 \beta_0 \frac{\hat{I}(z) \sin \theta}{4\pi r} e^{-j\beta_0 (r - z\cos\theta)}\, dz \tag{5.16}$$

The total electric field is the sum of these contributions:

$$\hat{E}_\theta = \int_{z=-l/2}^{z=l/2} j\eta_0\beta_0 \frac{\hat{I}(z)\sin\theta}{4\pi r} e^{-j\beta_0 r} e^{j\beta_0 z\cos\theta}\, dz \tag{5.17}$$

Substituting the expressions for $\hat{I}(z)$ given in (5.13), we obtain [1]

$$\begin{aligned}\hat{E}_\theta &= j\frac{\eta_0 \hat{I}_m e^{-j\beta_0 r}}{2\pi r} F(\theta) \\ &= j\frac{60\hat{I}_m e^{-j\beta_0 r}}{r} F(\theta)\end{aligned} \tag{5.18}$$

where the θ-variation term in this result is denoted by

$$\begin{aligned}F(\theta) &= \frac{\cos[\beta_0(\frac{1}{2}l)\cos\theta] - \cos\beta_0(\frac{1}{2}l)}{\sin\theta} \\ &= \frac{\cos[(\pi l/\lambda_0)\cos\theta] - \cos(\pi l/\lambda_0)}{\sin\theta}\end{aligned} \tag{5.19}$$

since $\beta_0 = 2\pi/\lambda_0$. The magnetic field in the far-field region of the Hertzian dipole is orthogonal to the electric field and related by η_0. If we carry through the above development for the magnetic field, we obtain

$$\hat{H}_\phi = \frac{\hat{E}_\theta}{\eta_0} \tag{5.20}$$

where $\hat{E}_\theta$ is given by (5.18).

The most frequently encountered case is the half-wave dipole, in which the total dipole length is $l = \frac{1}{2}\lambda_0$. Substituting into (5.19), we obtain

$$F(\theta) = \frac{\cos(\frac{1}{2}\pi\cos\theta)}{\sin\theta} \quad (\text{half-wave dipole, } l = \tfrac{1}{2}\lambda_0) \tag{5.21}$$

The electric field will be a maximum for $\theta = 90°$ (broadside to the antenna). For this case, $F(90°) = 1$, and the maximum electric field for the half-wave dipole becomes

$$|\hat{E}|_{\max} = 60\frac{|\hat{I}_m|}{r} \quad (\theta = 90°) \tag{5.22}$$

The field is directed in the θ direction and is independent of ϕ, which makes sense from symmetry considerations.

Now let us compute the average power radiated by the dipole. Once agai integration of the Poynting vector over a sphere of radius r gives the tot radiated power [1]:

$$P_{\text{rad}} = 73|\hat{I}_{\text{in,rms}}|^2 \quad \text{(in W)} \quad \text{(half-wave dipole)} \qquad (5.23$$

Thus, if we know the rms value of the *input current* at the terminals of a half-wav dipole, we may find the total average power radiated by multiplying the squar of the rms current by 73 Ω. This suggests that we define a radiation resistanc of the half-wave dipole as

$$R_{\text{rad}} = 73\ \Omega \quad \text{(half-wave dipole)} \qquad (5.24$$

There is one important difference between the dipole and the monopol Although the field patterns are the same, the monopole radiates only half th power of the dipole: the power radiated out of the half-sphere above the groun plane. Thus the radiation resistance for the monopole is half that of th corresponding dipole. In particular, for a quarter-wave monopole of lengt $h = \frac{1}{4}\lambda_0$ (which corresponds to a half-wave dipole), we have

$$R_{\text{rad}} = 36.5\ \Omega \quad \text{(quarter-wave monopole)} \qquad (5.25$$

Up to this point, we have not considered the total input impedance $\hat{Z}_{\text{in}}$ see at the terminals of the dipole or monopole antenna. The input impedance wil in general, have a real and an imaginary part as

$$\hat{Z}_{\text{in}} = R_{\text{in}} + jX_{\text{in}} \qquad (5.26$$

The input resistance will consist of the sum of the radiation resistance and th resistance of the imperfect wires used to construct the dipole, so that

$$\hat{Z}_{\text{in}} = R_{\text{loss}} + R_{\text{rad}} + jX_{\text{in}} \qquad (5.27$$

Figure 5.5 shows the radiation resistance and reactance referred to the base c a monopole antenna for various lengths of the antenna [3]. Figure 5.5 can b used to give the input impedance for a dipole by doubling the values given i the figure. The input reactance for a half-wave dipole (quarter-wave monopole is $X_{\text{in}} = 42.5\ \Omega$ ($X_{\text{in}} = 21.25\ \Omega$). Note that for monopoles that are shorter tha one-quarter wavelength (or dipoles shorter than one-half wavelength) th radiation resistance becomes much smaller and the reactive part become negative, symbolizing a capacitive reactance. Thus monopoles that are shorte than one-quarter wavelength appear at their input as a small resistance in serie with a capacitance, which is, intuitively, a sensible result. Also observe i Fig. 5.5 that the reactive part of the input impedance is zero for a monopole tha is slightly shorter than a quarter wavelength. Having a zero reactive part

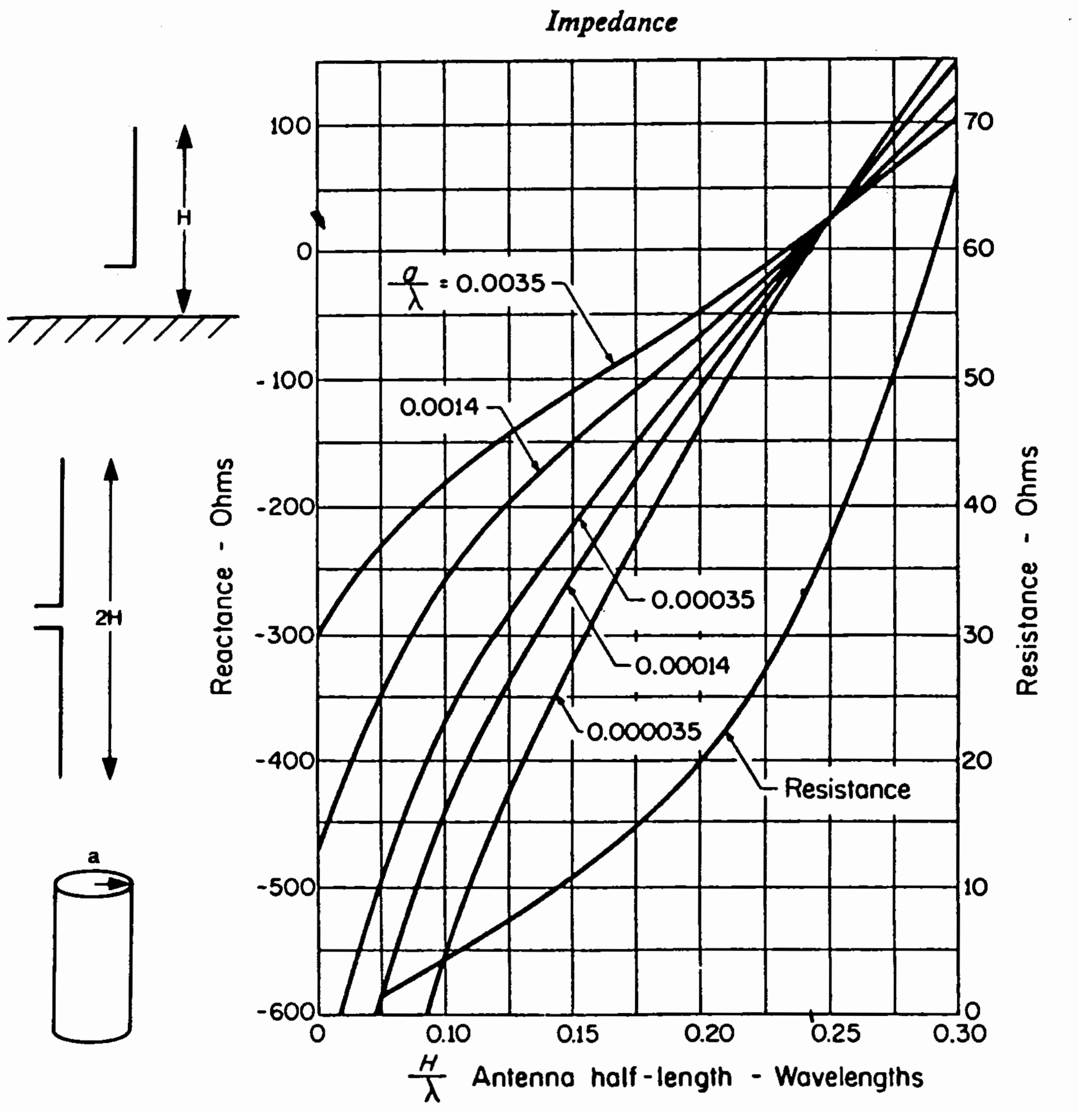

FIGURE 5.5 The radiation resistance and reactance of a dipole antenna as a function of length and wire radius [3].

obviously desirable from the standpoint of maximizing the power that is delivered from a source that has a real source resistance (such as 50 Ω) to the antenna and subsequently radiated. This is why monopoles are cut to lengths of slightly shorter than a quarter wavelength. If the physical length of a quarter-wave monopole is excessive for the intended installation, it can be shortened, but that introduces a large capacitive reactance to the input impedance, which necessitates a larger value of source excitation voltage to produce the same level of radiated power (dissipated in R_{rad}). In order to overcome this problem, short antennas have "loading coils" or inductors inserted in series with their input to cancel this capacitive reactance and increase the radiated power. This is sometimes referred to as "tuning the antenna."

The important point to realize here is that, knowing the input impedance of the antenna, we can compute the total average power radiated by the antenna by computing the average power dissipated in R_{rad}. For example, consider the half-wave dipole driven by a 100 V rms, 150 MHz, 50 Ω source as shown in Fig. 5.6(a). Replacing the antenna with its equivalent circuit at its input terminals gives the circuit shown in Fig. 5.6(b). The input current to the antenna is

$$\hat{I}_{ant} = \frac{\hat{V}_S}{R_{loss} + R_{rad} + jX}$$

Assume the wires are #20 AWG solid copper. The wires have radii much larger than a skin depth at the operating frequency of 150 MHz ($\delta = 5.4 \times 10^{-6}$ m),

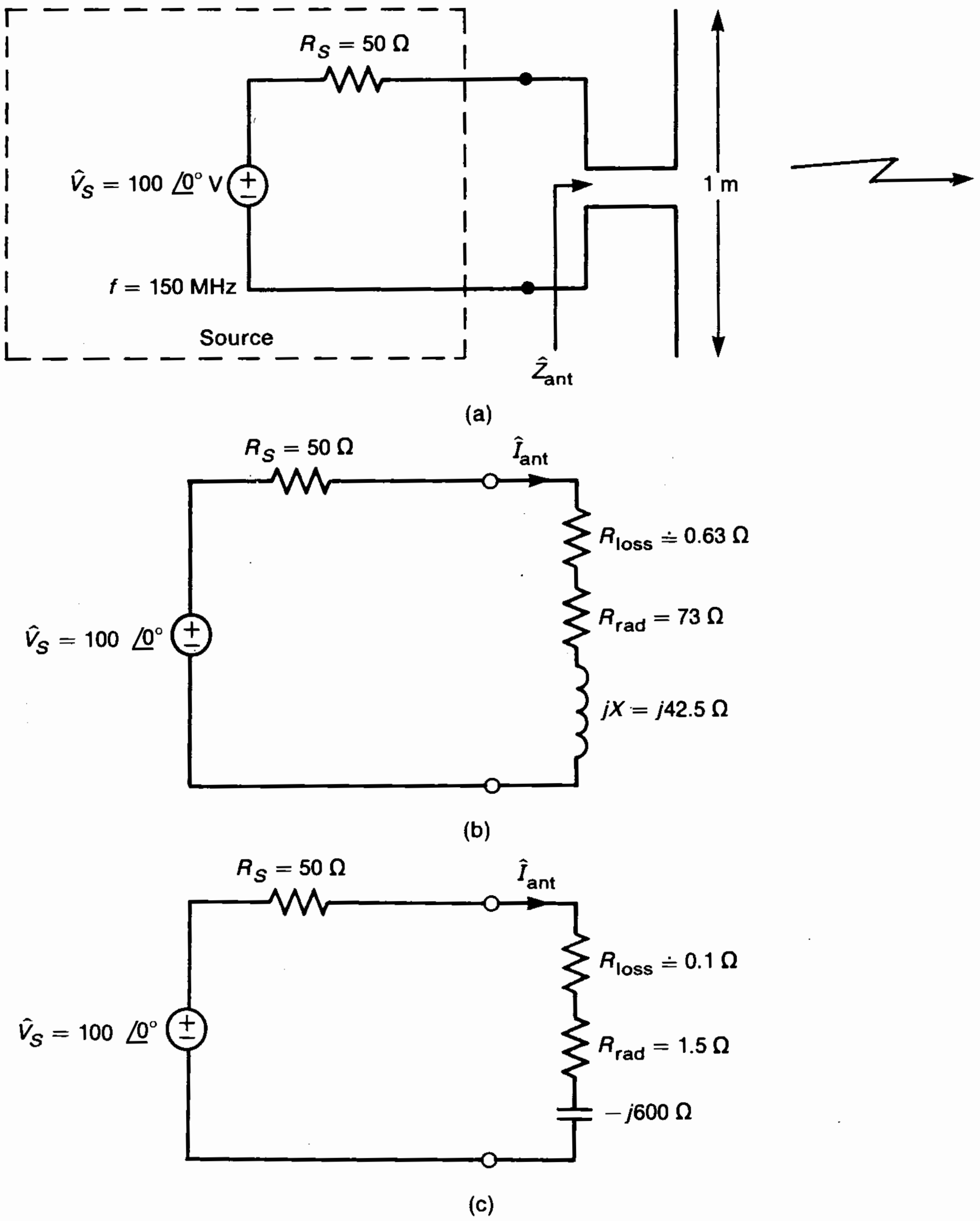

FIGURE 5.6 An example illustrating the computation of the radiated power of a dipole antenna: (b) equivalent circuit of a half-wave dipole; (c) equivalent circuit of a $\frac{1}{8}\lambda_0$ dipole.

so the high-frequency approximation for wire resistance developed in Chapter 4 can be used to compute R_{loss} as

$$r_{\text{wire}} = \frac{1}{2\pi r_w \sigma \delta}$$
$$= 1.25\ \Omega/\text{m}$$

Using this result, the net ohmic resistance of the wires used to construct the dipole can be obtained as [1]

$$R_{\text{loss}} = r_{\text{wire}} \tfrac{1}{2} l$$
$$= 0.63\ \Omega$$

Since the dipole is a half-wave dipole, its impedance is $(73 + j42.5)\ \Omega$ (see Fig. 5.5), so that

$$\hat{Z}_{\text{ant}} = R_{\text{loss}} + R_{\text{rad}} + jX_{\text{in}}$$
$$= (0.63 + 73 + j42.5)\ \Omega$$

Thus the current at the antenna input terminals is

$$\hat{I}_{\text{ant}} = \frac{100\ \underline{/0^\circ}}{50 + 73.63 + j42.5}$$
$$= 0.765\ \underline{/-18.97^\circ}\ \text{A}$$

The total average power dissipated in antenna losses is

$$P_{\text{loss}} = \tfrac{1}{2} |\hat{I}_{\text{ant}}|^2 R_{\text{loss}}$$
$$= 184\ \text{mW}$$

The total average power radiated is

$$P_{\text{rad}} = \tfrac{1}{2} |\hat{I}_{\text{ant}}|^2 R_{\text{rad}}$$
$$= 21.36\ \text{W}$$

In order to illustrate the effect of short antennas on radiated power, consider the previous problem where the dipole antenna is shortened to $\frac{1}{8}\lambda_0$ in total length. From Fig. 5.5, $R_{\text{rad}} \cong 1.5\ \Omega$ and $X_{\text{in}} \cong -600\ \Omega$. The equivalent circuit is shown in Fig. 5.6(c). The current at the input to the antenna is

$$\hat{I}_{\text{ant}} = \frac{100\ \underline{/0^\circ}}{50 + 0.16 + 1.5 - j600}$$
$$= 0.166\ \underline{/85.1^\circ}\ \text{A}$$

Thus the radiated power is

$$P_{\text{av,rad}} = \tfrac{1}{2}|\hat{I}_{\text{ant}}|^2 R_{\text{rad}}$$
$$= 20.7 \text{ mW}$$

Thus the reduced radiation resistance along with the large increase in reacti part in the input impedance caused by shortening the length of the dipole h significantly reduced the radiated power of the antenna. If an inductor havi an inductance of 0.637 μH giving a reactance of $+j600$ is inserted in series wi this antenna, the radiated power is increased to 2.81 W! This illustrates t extreme effect of the reactive part of the input impedance.

5.3 ANTENNA ARRAYS

The radiation characteristics of the Hertzian dipole, the magnetic dipole, t long dipole, and the monopole are evidently omnidirectional in any pla perpendicular to the antenna axis, since all fields are independent of ϕ. Th characteristic follows from the symmetry of these structures about the z ax From the standpoint of communication, we may wish to focus the radiat signal since any of the radiated power that is not transmitted in the directic of the receiver is wasted. On the other hand, from the standpoint of EMC, v may be interested in directing the radiated signal away from another receiv in order to prevent interference with that receiver. If the transmitting antenn has an omnidirectional pattern, we do not have these options. In this sectic we will investigate how to use two or more omnidirectional antennas to produ maxima and/or nulls in the resulting pattern. This results from phasing t currents to the antennas and separating them sufficiently such that the combin fields will add constructively or destructively to produce these resulting maxin or nulls. This result, although applied to the emission patterns of communicatio antennas, has direct application in the radiated emissions of products, since illustrates how multiple emissions may combine. In addition, we will use t simple results obtained here to obtain simple models for predicting the radiat emissions from wires and PCB lands in Chapter 8.

Consider two omnidirectional antennas such as half-wave dipoles in fr space or quarter-wave monopoles above ground, as shown in Fig. 5.7(a). Th current elements lie on the y axis and are directed in the z direction. They a separated by a distance d and are equally spaced about the origin. Assumi the field point P is in the far field of the antennas, $\theta_1 \cong \theta_2 \cong \theta$, the far fiel at point P due to each antenna are of the form

$$\hat{E}_{\theta 1} = \frac{\hat{M} I \underline{/\alpha}}{r_1} e^{-j\beta_0 r_1} \qquad (5.28$$

$$\hat{E}_{\theta 2} = \frac{\hat{M} I \underline{/0}}{r_2} e^{-j\beta_0 r_2} \qquad (5.28$$

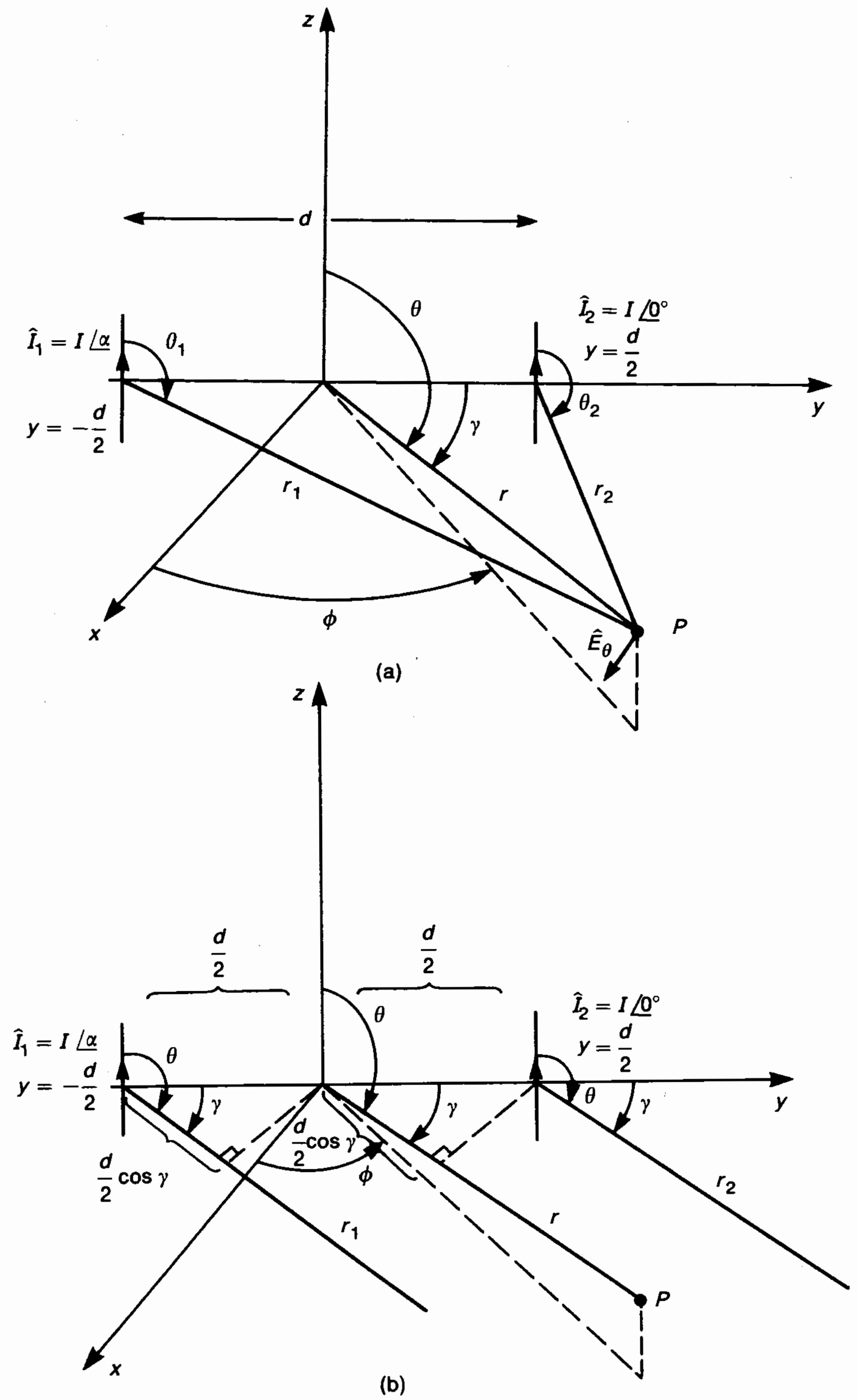

FIGURE 5.7 Computation of the radiated fields of an array of two dipoles: (a) definitions; (b) the far-field approximation.

where $\hat{I}_1 = I\,\angle\alpha$ and $\hat{I}_2 = I\,\angle 0$, and we assume the currents of the two antennas are equal in magnitude but the current in antenna #1 leads that of antenna #2 by α. The factor $\hat{M}$ depends on the type of antennas used. For Hertzian dipoles $\hat{M} = j\eta_0\beta_0(dl/4\pi)\sin\theta$ (see (5.3a)). For long dipoles $\hat{M} = j60F(\theta)$ (see

(5.18)). The total field at point P is the sum of the fields of the two antennas

$$\hat{E}_\theta = \hat{E}_{\theta 1} + \hat{E}_{\theta 2} \tag{5.29}$$

$$= \hat{M}I\left(\frac{e^{-j\beta_0 r_1}}{r_1}e^{j\alpha} + \frac{e^{-j\beta_0 r_2}}{r_2}\right)$$

$$= \hat{M}Ie^{j\alpha/2}\left(\frac{e^{-j\beta_0 r_1}e^{j\alpha/2}}{r_1} + \frac{e^{-j\beta_0 r_2}e^{-j\alpha/2}}{r_2}\right)$$

In order to simplify this equation, we observe that, since P is sufficiently far from the origin, $r_1 \cong r_2 \cong r$, where r is the distance from the midpoint of the array to point P. This approximation can be used in the denominator terms of (5.29) but cannot be used in the phase terms for reasons discussed in the previous section. If we draw the radius vectors r_1 and r_2 parallel to the radius vector r as shown in Fig. 5.7(b) then we can obtain a reasonable approximation for the phase terms. To do this, we observe that the path lengths can be written in terms of the angle γ between the radius vector and the y axis as

$$r_1 \cong r + \tfrac{1}{2}d\cos\gamma \tag{5.30a}$$

$$r_2 \cong r - \tfrac{1}{2}d\cos\gamma \tag{5.30b}$$

The angle γ can be obtained as the dot product of the unit vector in the radial direction and the unit vector in the y direction as [1]

$$\cos\gamma = \vec{a}_r \cdot \vec{a}_y \tag{5.31}$$

$$= \sin\theta\sin\phi$$

Substituting gives

$$r_1 \cong r + \tfrac{1}{2}d\sin\theta\sin\phi \tag{5.32a}$$

$$r_2 \cong r - \tfrac{1}{2}d\sin\theta\sin\phi \tag{5.32b}$$

where d is the separation between the two antennas. Substituting (5.32) into (5.29) gives

$$\hat{E}_\theta = \frac{\hat{M}I}{r}e^{j\alpha/2}e^{-j\beta_0 r}\left(e^{j(\beta_0(d/2)\sin\theta\sin\phi - \alpha/2)}\right. \tag{5.33}$$

$$\left. + e^{-j(\beta_0(d/2)\sin\theta\sin\phi - \alpha/2)}\right)$$

$$= 2e^{j\alpha/2}\underbrace{\left[\hat{M}\frac{Ie^{-j\beta_0 r}}{r}\right]}_{\substack{\text{pattern of}\\ \text{individual}\\ \text{elements}}} \times \underbrace{\cos\left(\frac{\pi d}{\lambda_0}\sin\theta\sin\phi - \frac{\alpha}{2}\right)}_{F_{\text{array}}(\theta,\phi)}$$

We have substituted $\beta_0 = 2\pi/\lambda_0$ and $\cos\psi = \frac{1}{2}(e^{j\psi} + e^{-j\psi})$. Observe that the resultant field is the *product* of the pattern of the individual (identical) antenna elements and the array factor $F_{\text{array}}(\theta, \phi)$, which depends only on the antenna spacing in electrical dimensions, d/λ_0, and phasing of the currents, α. This is referred to as *the principle of pattern multiplication.*

The antenna spacing and/or the relative phase of the currents can be chosen to give maxima/minima in the radiation pattern of the array. For example,

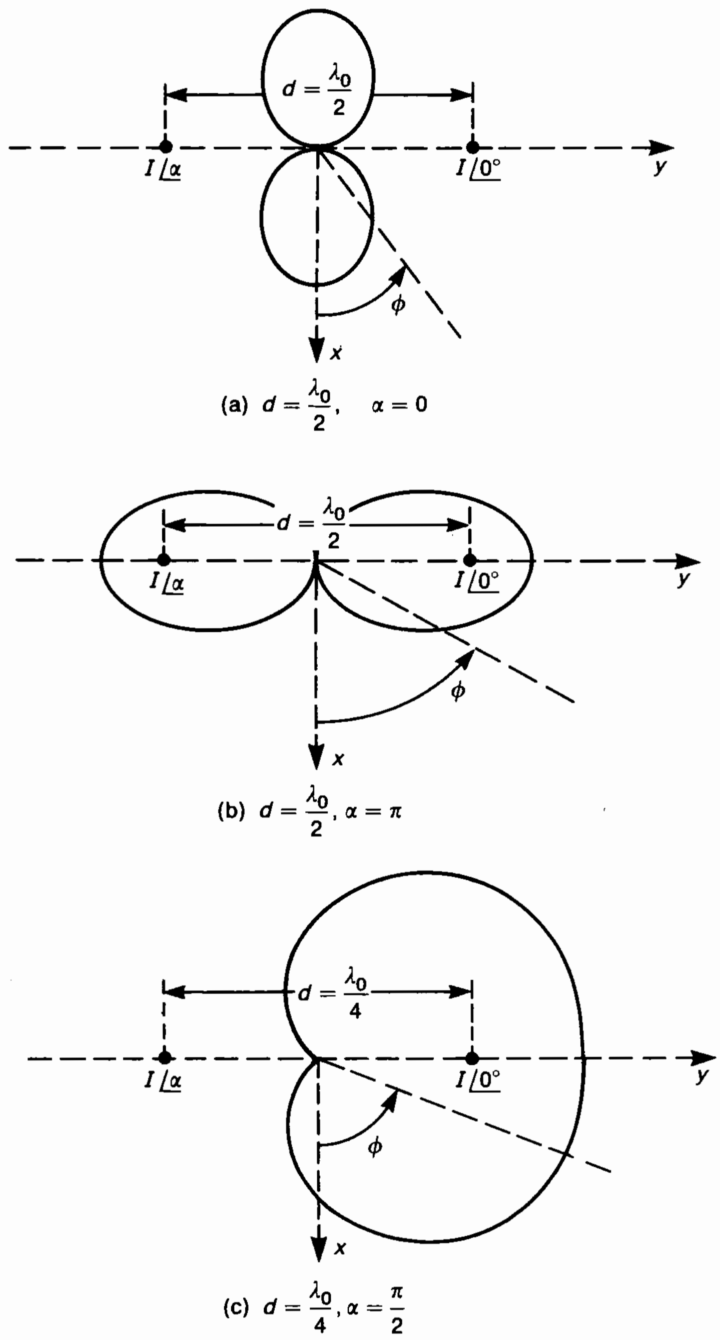

FIGURE 5.8 Patterns of a two-element array: (a) $d = \frac{1}{2}\lambda_o$, $\alpha = 0°$; (b) $d = \frac{1}{2}\lambda_o$, $\alpha = 180°$; (c) $d = \frac{1}{4}\lambda_o$, $\alpha = 90°$.

suppose the spacing between the antennas is one-half wavelength and $\alpha = 0°$. The array factor in the plane perpendicular to the array, $\theta = 90°$, is

$$F_{\text{array}}(\theta = 90°, \phi) = \cos(\tfrac{1}{2}\pi \sin \phi)$$

This pattern is plotted in Fig. 5.8(a). We have produced a pattern that has nulls at $\phi = 90°$ and $\phi = 270°$. As another example, suppose the antenna spacing is again one-half wavelength but $\alpha = 180°$. The array factor is

$$F_{\text{array}}(\theta = 90°, \phi) = \cos(\tfrac{1}{2}\pi \sin \phi - \tfrac{1}{2}\pi)$$

The pattern is plotted in Fig. 5.8(b). This has produced nulls at $\phi = 0°$ and $\phi = 180°$. As a final example, suppose the antennas are spaced one-quarter wavelength and $\alpha = 90°$. The array factor is

$$F_{\text{array}}(\theta = 90°, \phi) = \cos(\tfrac{1}{4}\pi \sin \phi - \tfrac{1}{4}\pi)$$

The pattern is plotted in Fig. 5.8(c), and we have produced a null at $\phi = 270°$ and a maximum at $\phi = 90°$.

5.4 CHARACTERIZATION OF ANTENNAS

The antennas we have considered previously are quite simple to analyze. The analysis of other antennas to determine their total radiated power and, more importantly, the shapes of their radiated emission patterns is not so simple. These more complicated antennas are more commonly characterized by measured parameters such as directivity and gain, effective aperture, and/or antenna factor. The purpose of this section is to investigate these criteria.

5.4.1 Directivity and Gain

The *directive gain* of an antenna, $D(\theta, \phi)$, is a measure of the concentration of the radiated power in a particular θ, ϕ direction at a fixed distance r away from the antenna. For the elemental dipoles, the long dipole, and the monopole, we noted that the radiated power is a maximum for $\theta = 90°$ and is zero for $\theta = 0°$ and $\theta = 180°$. To obtain a more quantitative measure of this concentration of radiated power, we will define the radiation intensity $U(\theta, \phi)$.

We found that the far-field, radiated average-power densities for the Hertzian dipole, the magnetic dipole, the long dipole, and the monopole are of the form

$$\vec{S}_{\text{av}} = \frac{|\vec{\hat{E}}_{\text{far field}}|^2}{2\eta_0}\vec{a}_r \tag{5.34}$$

$$= \frac{E_0^2}{2\eta_0 r^2}\vec{a}_r$$

where E_0 depends on θ, the antenna type and the antenna current. To obtain a power pattern relationship that is independent of distance from the antenna, we multiply (5.34) by r^2 and define the resulting quantity to be the *radiation intensity*, that is,

$$U(\theta, \phi) = r^2 S_{av} \tag{5.35}$$

The radiation intensity will be a function of θ and ϕ but will be independent of distance from the antenna. The total average power radiated will be

$$\begin{aligned} P_{rad} &= \oint \vec{S}_{av} \cdot d\vec{s} \\ &= \int_{\theta=0}^{\pi} \int_{\phi=0}^{2\pi} U(\theta, \phi) \sin\theta \, d\phi \, d\theta \\ &= \oint_S U(\theta, \phi) \, d\Omega \end{aligned} \tag{5.36}$$

The quantity $d\Omega = \sin\theta \, d\phi \, d\theta$ is an element of *solid angle* Ω, and the unit of solid angle is the steradian (sr). The units of U are therefore watts per steradian (W/sr). Note that for $U = 1$, (5.36) integrates to 4π. *The total radiated power is therefore the integral of the radiation intensity over a solid angle of 4π sr.* Note also that the average radiation intensity is the total radiated power divided by 4π sr:

$$U_{av} = \frac{P_{rad}}{4\pi} \tag{5.37}$$

The radiation intensity for more complicated antennas is similarly defined. *The directive gain of an antenna in a particular direction, $D(\theta, \phi)$, is the ratio of the radiation intensity in that direction to the average radiation intensity*:

$$\begin{aligned} D(\theta, \phi) &= \frac{U(\theta, \phi)}{U_{av}} \\ &= \frac{4\pi U(\theta, \phi)}{P_{rad}} \end{aligned} \tag{5.38}$$

The directivity of the antenna is the directive gain in the direction that yields a maximum:

$$D_{max} = \frac{U_{max}}{U_{av}} \tag{5.39}$$

As an example, let us compute the directive gain and directivity of a Hertzia[n] dipole. The radiation intensity is found from (5.5) and (5.35),

$$U(\theta, \phi) = r^2 S_{av} = 15\pi\left(\frac{dl}{\lambda_0}\right)^2 |\hat{I}|^2 \sin^2\theta$$

and the radiated power is given by (5.6),

$$P_{rad} = 40\pi^2\left(\frac{dl}{\lambda_0}\right)^2 |\hat{I}|^2$$

Thus the directive gain is

$$D(\theta, \phi) = \frac{4\pi U(\theta, \phi)}{P_{rad}} = 1.5 \sin^2\theta$$

The directivity is therefore the directive gain at $\theta = \frac{1}{2}\pi$:

$$D_{max} = 1.5$$

For the half-wave dipole we obtain

$$D(\theta, \phi) = \frac{\eta_0}{\pi R_{rad}} F^2(\theta) = 1.64\, F^2(\theta)$$

where $F(\theta)$ is given by (5.21) and $R_{rad} = 73\ \Omega$ and

$$D_{max} = 1.64$$

which occurs for $\theta = \frac{1}{2}\pi$.

The directive gain $D(\theta, \phi)$ of an antenna is simply a function of the shap[e] of the antenna pattern. The *power gain* $G(\theta, \phi)$, on the other hand, takes int[o] account the losses of the antenna. Suppose that a total power P_{app} is applie[d] to the antenna and only P_{rad} is radiated. The difference is consumed in ohmi[c] losses of the antenna as well as in other inherent losses such as those in a[n] imperfect ground for monopole antennas. If we define an *efficiency factor* e a[s]

$$e = \frac{P_{rad}}{P_{app}} \tag{5.40}$$

then the power gain is related to the directive gain as

$$G(\theta, \phi) = eD(\theta, \phi) \tag{5.41}$$

where we have defined the power gain as

$$G(\theta, \phi) = \frac{4\pi U(\theta, \phi)}{P_{\text{app}}} \tag{5.42}$$

For most antennas, the efficiency is nearly 100%, and thus the power gain and directive gain are nearly equal.

We also need to discuss the concept of an isotropic point source. An *isotropic point source* is a fictitious lossless antenna that radiates power equally in all directions. Since this antenna is lossless, its directive and power gains are equal. For an isotropic point source radiating or transmitting a total power P_T, the power density at some distance d away is the total radiated power divided by the area of a sphere of radius d:

$$\vec{S}_{\text{av}} = \frac{P_T}{4\pi d^2}\vec{a}_r \tag{5.43}$$

We can calculate the electric and magnetic fields for the isotropic point source from the realization that the waves resemble (locally) uniform plane waves so that

$$\vec{S}_{\text{av}} = \frac{|\vec{\hat{E}}|^2}{2\eta_0} \quad \text{(in W/m}^2\text{)} \tag{5.44}$$

Combining (5.43) and (5.44) gives

$$|\vec{\hat{E}}| = \frac{\sqrt{60P_T}}{d}\vec{a}_\theta \quad \text{(in V/m)} \tag{5.45}$$

where we have substituted $\eta_0 = 120\pi\ \Omega$.

The isotropic point source, although quite idealistic, is useful as a standard or reference antenna to which we refer many of our calculations. For example, since the isotropic point source is lossless, the directive and power gains are equal, and both will be designated by G_0. The gain becomes

$$\begin{aligned} G_0(\theta, \phi) &= \frac{4\pi U_0(\theta, \phi)}{P_T} \\ &= 1 \end{aligned} \tag{5.46}$$

Therefore the directive gain and power gain of other antennas may be though of as being determined with respect to an isotropic point source. In certai other cases, the gain of an antenna may be referred to the gain of a half-wav dipole. When discussing gain, one must be careful to determine the referenc antenna.

Quite often the gain (directive or power) of an antenna is given in decibel where

$$G_{\mathrm{dB}} = 10 \log_{10} G \tag{5.47}$$

For example, the Hertzian dipole has a directivity of 1.76 dB and the isotropi point source has a directivity of 0 dB. The half-wave dipole has a maximur gain of 2.15 dB. Equivalently, we say that the gain of an antenna is the gai over (or with respect to) an isotropic antenna:

$$G_{\mathrm{dB}} = 10 \log_{10}\left(\frac{G}{G_0}\right) \tag{5.48}$$

We are frequently interested in the coupling between two antennas, one o which is used as a transmitter and the other as a receiver. An important princip in this problem is that of *reciprocity* [1, 3–6]. Reciprocity provides that th source and receiver can be interchanged without affecting the results so lon as the impedances of the source and receiver are the same. Several additiona properties can be proven. The impedance seen looking into an antenna terminal when it is used for transmission is the same as the Thévenin source impedanc seen looking back into its terminals when it is used for reception. In additio the transmission pattern of the antenna is the same as its reception pattern.

5.4.2 Effective Aperture

An additional useful concept is that of an antenna's effective aperture. Th *effective aperture* of an antenna is related to the ability of the antenna to extra energy from a passing wave. *The effective aperture of an antenna, A_e, is the rati of the power received (in its load impedance), P_R, to the power density of th incident wave, S_{av}, when the polarization of the incident wave and the polarizatio of the receiving antenna are matched*:

$$A_e = \frac{P_R}{S_{\mathrm{av}}} \quad (\text{in m}^2) \tag{5.49}$$

The *maximum effective aperture* A_{em} is the ratio in (5.49) when the load impedanc is the conjugate of the antenna impedance, which means that maximum powe transfer to the load takes place. For a linearly polarized incident wave and receiving antenna such as a dipole or monopole that produces linearly polarize waves when it is used for transmission, the requirement for matched polarization essentially means that the antenna is oriented with respect to the incident wave to produce the maximum response; that is, the electric field vector of the incident wave is parallel to the electric field vector that would be produced by this antenna when it is used for transmission.

For example, let us compute the maximum effective aperture of a Hertzian dipole antenna. If the dipole is terminated in an impedanze $\hat{Z}_L$, we assume that $\hat{Z}_L = R_{\text{rad}} - jX$ where the input impedance to the dipole is $\hat{Z}_{\text{in}} = R_{\text{rad}} + jX$ and the dipole is assumed to be lossless. Suppose that the incident wave is arriving at an angle θ, with the electric field vector in the θ direction as shown in Fig. 5.9. The open-circuit voltage produced at the terminals of the antenna is

$$|\hat{V}_{OC}| = |\hat{E}_\theta|\, dl \sin\theta \tag{5.50}$$

The power density in the incident wave is

$$S_{\text{av}} = \frac{1}{2}\frac{|\hat{E}_\theta|^2}{\eta_0} \tag{5.51}$$

Since the load is matched for maximum power transfer, the power received is

$$\begin{aligned} P_R &= \frac{|\hat{V}_{OC}|^2}{8R_{\text{rad}}} \\ &= \frac{|\hat{E}_\theta|^2\, dl^2 \sin^2\theta}{8R_{\text{rad}}} \end{aligned} \tag{5.52}$$

Substituting the value for R_{rad} given in (5.7) gives

$$P_R = \frac{|\hat{E}_\theta|^2 \lambda_0^2}{640\pi^2}\sin^2\theta \tag{5.53}$$

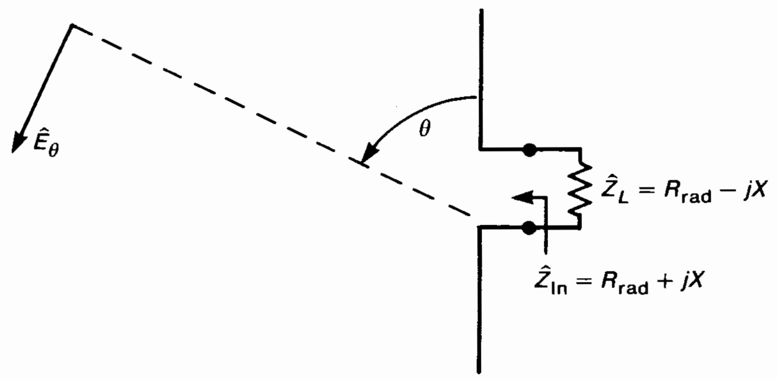

FIGURE 5.9 Illustration of the computation of the maximum effective aperture A_{em} of a linear antenna.

Thus the maximum effective aperture is

$$A_{em}(\theta, \phi) = \frac{P_R}{S_{av}} = 1.5 \sin^2 \theta \frac{\lambda_0^2}{4\pi} = \frac{\lambda_0^2}{4\pi} D(\theta, \phi) \quad (5.5$$

where we have substituted the directive gain of the Hertzian dipole

$$D(\theta, \phi) = 1.5 \sin^2 \theta \quad (5.5$$

and θ is the direction of the incoming wave. Observe that the maximum effecti aperature of an antenna is not necessarily related to its "physical aperture."

It can be shown that the result in (5.54) is a general result valid for mo general antennas; that is, the maximum effective aperture of an antenna us for reception is related to the directive gain *in the direction of the incoming wa* of that antenna when it is used for transmission as [1, 3–6]:

$$G(\theta, \phi) = \frac{4\pi}{\lambda_0^2} A_{em}(\theta, \phi) \quad (5.5$$

The direction for A_{em} (the direction of the incoming incident wave with respe to the receiving antenna) is the direction of the gain G (the gain of the anten in this direction when it is used for transmission). We have interchanged directi gain D and power gain G on the assumption that the antennas are lossless.

5.4.3 Antenna Factor

The above properties of antennas are more commonly used in the area of t use of antennas for communication such as signal transmission and rad applications. In the area of their use in EMC a more common way characterizing the reception properties of an antenna is with the notion of *antenna factor*. Consider a dipole antenna that is used to measure the elect field of an incident, linearly polarized uniform plane wave as shown in F 5.10(a). A receiver such as a spectrum analyzer is attached to the terminals this measurement antenna. The voltage measured by this instrument is denot as $\hat{V}_{rec}$. It is desired to relate this received voltage to the incident electric fie This is done with the antenna's *antenna factor*, which is defined as *the ratio the incident electric field at the surface of the measurement antenna to the receiv*

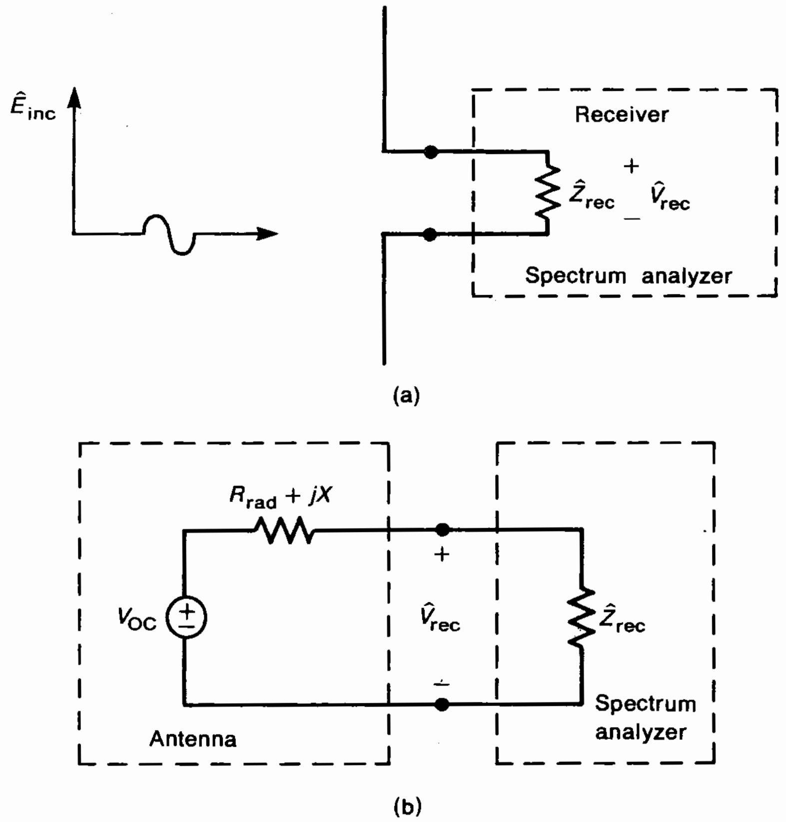

FIGURE 5.10 The antenna factor AF: (a) general circuit; (b) equivalent circuit.

voltage at the antenna terminals:

$$\mathrm{AF} = \frac{\mathrm{V/m\ in\ incident\ wave}}{\mathrm{V\ received}} \quad (\mathrm{in}\ 1/\mathrm{m}) \tag{5.57}$$

$$= \frac{|\hat{E}_{\mathrm{inc}}|}{|\hat{V}_{\mathrm{rec}}|}$$

This is frequently expressed in dB as

$$\mathrm{AF_{dB}} = \mathrm{dB}\mu\mathrm{V/m\ (incident\ field)} - \mathrm{dB}\mu\mathrm{V\ (received\ voltage)} \tag{5.58a}$$

or

$$\mathrm{dB}\mu\mathrm{V/m\ (incident\ field)} = \mathrm{dB}\mu\mathrm{V\ (received\ voltage)} + \mathrm{AF_{dB}} \tag{5.58b}$$

Note that the units of the antenna factor are 1/m. The units are frequently ignored, and the antenna factor is stated in dB. The antenna factor is usually furnished by the manufacturer of the antenna as measured data at various frequencies in the range of intended use of the measurement antenna. A typical such plot provided by the manufacturer of a biconical measurement antenna

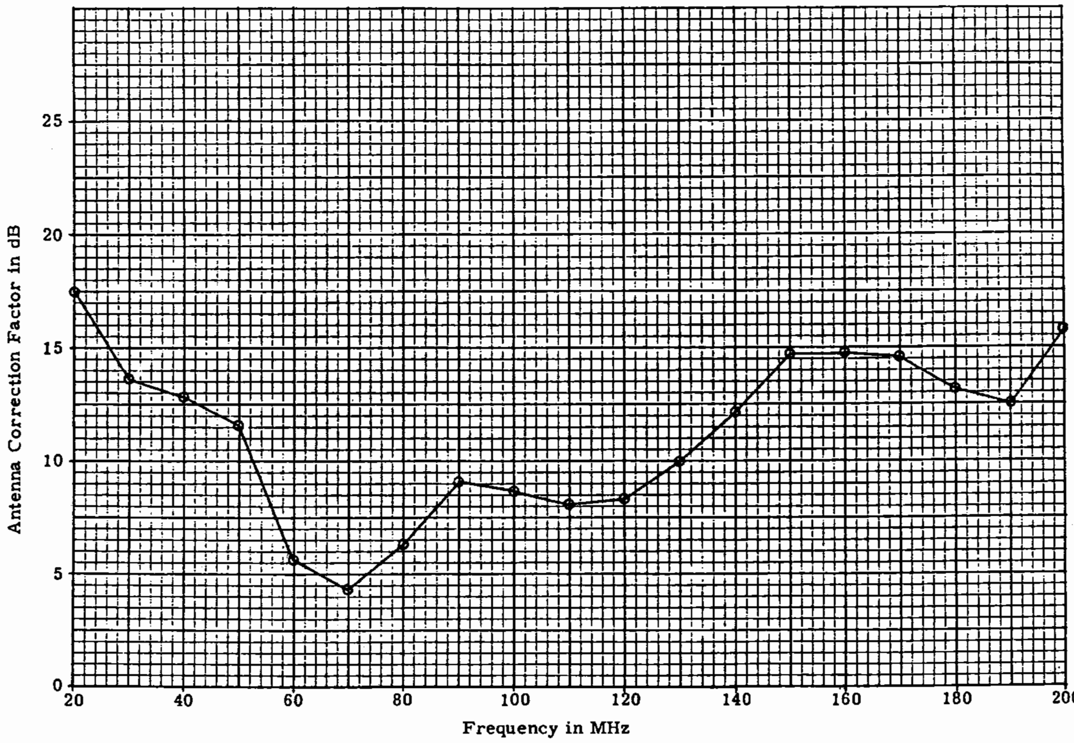

FIGURE 5.11 The antenna factor versus frequency for a typical biconical EMC measurement antenna (courtesy the Eaton Corporation).

is shown in Fig. 5.11. A known field is provided by some standard antenna at a calibrated test site such as the National Institute of Standards and Technology (NIST) in Boulder, Colorado in the US (formerly known as the National Bureau of Standards or NBS). The ratio of the known value of the incident field to the measured voltage at the terminals of the antenna in dB according to (5.57) or (5.58) is plotted for the antenna versus frequency. The reciprocal of the antenna factor is referred to as the *antenna effective height* h_e [3–6].

There are several important implicit assumptions in these measured antenna factor data. *If any of these implicit assumptions are not adhered to in the course of using this antenna for measurement then the measured data are invalid.* The first important assumption is that *the incident field is polarized for maximum response of the antenna.* For a dipole or other wire-type antenna this means that *the response will be the component of the incident field that is parallel to the antenna axis.* Ordinarily this is what is desired, since the antenna will be typically used to measure vertical and horizontal fields in testing for compliance to the radiated emission regulatory limits. The second important implicit assumption has to do with *the input impedance of the receiver that is used not only to make the measurement but also to calibrate the antenna.* The most common impedance is the typical input impedance to virtually all spectrum analyzers, and that is 50 Ω. Nevertheless, the antenna manufacturer should explicitly state what termination impedance was used in the calibration. Note that *this does not assume that the receiver is matched to the antenna,* and usually it will not be.

However, from the standpoint of using the antenna factor calibration chart for that antenna it is only important to use a termination impedance that is the same as was used to calibrate the antenna.

On the other hand, suppose we wish to *calculate* the antenna factor of an ideal antenna such as a dipole from the field equations, maximum effective aperture, etc. for that antenna. Since the spectrum analyzer input impedance is $\hat{Z}_{rec} = (50 + j0)\,\Omega$ and is therefore not matched, we must use the equivalent circuit of Fig. 5.10(b) to obtain this. First compute $\hat{V}_{rec,matched}$ *assuming a matched load*, $\hat{Z}_{rec} = R_{rad} - jX$, using the results in the previous sections. Then use this result to obtain the open-circuit voltage $\hat{V}_{OC} = 2\hat{V}_{rec,matched}$. Then use the equivalent circuit in Fig. 5.10(b) to compute the actual received voltage $\hat{V}_{rec}$, and from that the antenna factor.

As an example of the use of measured data to determine the antenna factor, consider the calibration of a measurement antenna shown in Fig. 5.12. A known, incident, linearly polarized, uniform plane wave is incident on the antenna, and the electric field *at the position of the antenna in the absence of the antenna* is 60 dBμV/m. A 30 foot length of RG-58U coaxial cable is used to connect the antenna to a 50 Ω spectrum analyzer. The spectrum analyzer measures 40 dBμV. Since the antenna factor relates the incident electric field to the *voltage at the base of the antenna*, we must relate the spectrum analyzer reading to the voltage at the base of the antenna. The coaxial cable has 4.5 dB/100 feet loss at the frequency of the incident wave, 100 MHz. Thus the cable loss of 1.35 dB must be *added* to the spectrum analyzer reading to give the voltage at the antenna terminals of 41.35 dBμV. Therefore the antenna factor is

$$\begin{aligned} AF_{dB} &= 60\ \text{dB}\mu\text{V/m} - 41.35\ \text{dB}\mu\text{V} \\ &= 18.65\ \text{dB} \end{aligned}$$

It is a simple matter to convert the spectrum analyzer readings to the value of incident field; *add the antenna factor in dB to the spectrum analyzer reading*

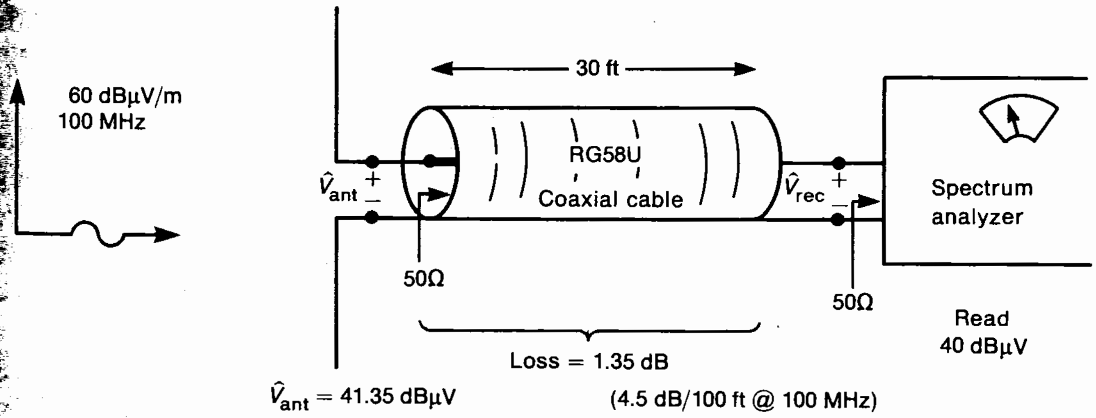

FIGURE 5.12 An example illustrating the use of the antenna factor to compute the received voltage.

in dBμV and add the connection loss in dB to give the incident electric field in dBμV/m:

$$E\,(\mathrm{dB}\mu\mathrm{V/m}) = \mathrm{AF}\,(\mathrm{dB}) + V_{SA}\,(\mathrm{dB}\mu\mathrm{V}) + \text{Cable Loss}\,(\mathrm{dB}) \tag{5.59}$$

Observe that the connection cable loss must be added and not subtracted, since the antenna factor is with respect to the base of the antenna and does not include any connection cable loss (unless explicitly stated by the antenna manufacturer).

5.4.4 Effects of Balancing and Baluns

The ideal antennas that we are considering are inherently *balanced structures*. There are numerous definitions of this concept of *balanced structure*. Generally, but not always, these seemingly different definitions lead to the same result. For example, consider the long dipole antenna shown in Fig. 5.3(a). In the analysis of this antenna we assumed that the current $\hat{I}(z_1)$ at a point z_1 on the upper arm is the same in magnitude as the current at the corresponding position on the lower arm, $-z_1$ (a point that is the same distance from the feed point as the point on the upper arm). From this standpoint of *symmetry of the antenna currents*, the antenna is *inherently* a balanced structure. This also inherently assumes that the current entering one terminal of the antenna is equal but opposite to the current entering the other terminal. Nearby metallic obstacles such as ground planes can upset this balance, causing the pattern to deviate substantially from the ideal pattern that was obtained from the assumption of balanced currents on the arms of the antenna [7].

Other factors can upset the balance of the currents on the antenna structure. The most common type of feedline that is used to supply signals to antennas is the coaxial cable. Under ideal conditions, the current returns to its source on the *interior* of the overall shield. If this type of cable is attached to an inherently balanced structure such as a dipole antenna, some of the current may flow *on the outside of the shield*. This current will radiate, whereas the current going down the interior wire and returning on the inside of the shield will not. The amount of current that flows on the outside of the shield depends on "the impedance to ground" between the shield exterior and the ground, $\hat{Z}_G$, along with the excitation of the shield exterior (unintentional excitation).

The common way of preventing unbalance due to a coaxial feed cable is the use of a *balun*, which is an acronym for BALanced to UNbalanced, referring to the transition from an unbalanced coaxial cable to a balanced antenna. The balun is inserted at the input to the antenna, as shown in Fig. 5.13(a). In the case of the coaxial feed cable, the intent of the balun is to increase the impedance between the outside of the shield and ground. A common form is the "bazooka balun" shown in Fig. 5.13(b). A quarter-wavelength section of shield is inserted over the shield of the original cable, and these are shorted together a quarter-wavelength from the feed point. A quarter-wavelength, short-circuited transmission line is formed between the outer coax and the inner coax. We found in the previous chapter that a short-circuited, quarter-wavelength