CHAPTER 3

Mobile Radio Propagation: Large-Scale Path Loss

The mobile radio channel places fundamental limitations on the performance of wireless communication systems. The transmission path between the transmitter and the receiver can vary from simple line-of-sight to one that is severely obstructed by buildings, mountains, and foliage. Unlike wired channels that are stationary and predictable, radio channels are extremely random and do not offer easy analysis. Even the speed of motion impacts how rapidly the signal level fades as a mobile terminal moves in space. Modeling the radio channel has historically been one of the most difficult parts of mobile radio system design, and is typically done in a statistical fashion, based on measurements made specifically for an intended communication system or spectrum allocation.

3.1 Introduction to Radio Wave Propagation

The mechanisms behind electromagnetic wave propagation are diverse, but can generally be attributed to reflection, diffraction, and scattering. Most cellular radio systems operate in urban areas where there is no direct line-of-sight path between the transmitter and the receiver, and where the presence of high-rise buildings causes severe diffraction loss. Due to multiple reflections from various objects, the electromagnetic waves travel along different paths of varying lengths. The interaction between these waves causes multipath fading at a specific location, and the strengths of the waves decrease as the distance between the transmitter and receiver increases.

Propagation models have traditionally focused on predicting the average received signal strength at a given distance from the transmitter, as well as the variability of the signal strength in close spatial proximity to a particular location. Propagation models that predict the mean signal strength for an arbitrary transmitter-receiver (T-R) separation distance are useful in estimating the radio coverage area of a transmitter and are called *large-scale* propagation models, since they characterize signal strength over large T-R separation distances (several hundreds or thousands of meters). On the other hand, propagation models that characterize the rapid fluctuations of the received signal strength over very short travel distances (a few wavelengths) or short time durations (on the order of seconds) are called *small-scale* or *fading* models.

As a mobile moves over very small distances, the instantaneous received signal strength may fluctuate rapidly giving rise to small-scale fading. The reason for this is that the received signal is a sum of many contributions coming from different directions, as described in Chapter 4. Since the phases are random, the sum of the contributions varies widely; for example, obeys a Rayleigh fading distribution. In small-scale fading, the received signal power may vary by as much as three or four orders of magnitude (30 or 40 dB) when the receiver is moved by only a fraction of a wavelength. As the mobile moves away from the transmitter over much larger distances, the local average received signal will gradually decrease, and it is this local average signal level that is predicted by large-scale propagation models. Typically, the local average received power is computed by averaging signal measurements over a measurement track of 5λ to 40λ. For cellular and PCS frequencies in the 1 GHz to 2 GHz band, this corresponds to measuring the local average received power over movements of 1 m to 10 m.

Figure 3.1 illustrates small-scale fading and the slower large-scale variations for an indoor radio communication system. Notice in the figure that the signal fades rapidly as the receiver moves, but the local average signal changes much more slowly with distance. This chapter covers large-scale propagation and presents a number of common methods used to predict received power in mobile communication systems. Chapter 4 treats small-scale fading models and describes methods to measure and model multipath in the mobile radio environment.

3.2 Free Space Propagation Model

The free space propagation model is used to predict received signal strength when the transmitter and receiver have a clear, unobstructed line-of-sight path between them. Satellite communication systems and microwave line-of-sight radio links typically undergo free space propagation. As with most large-scale radio wave propagation models, the free space model predicts that received power decays as a function of the T-R separation distance raised to some power (i.e. a power law function). The free space power received by a receiver antenna which is separated from a radiating transmitter antenna by a distance d, is given by the Friis free space equation,

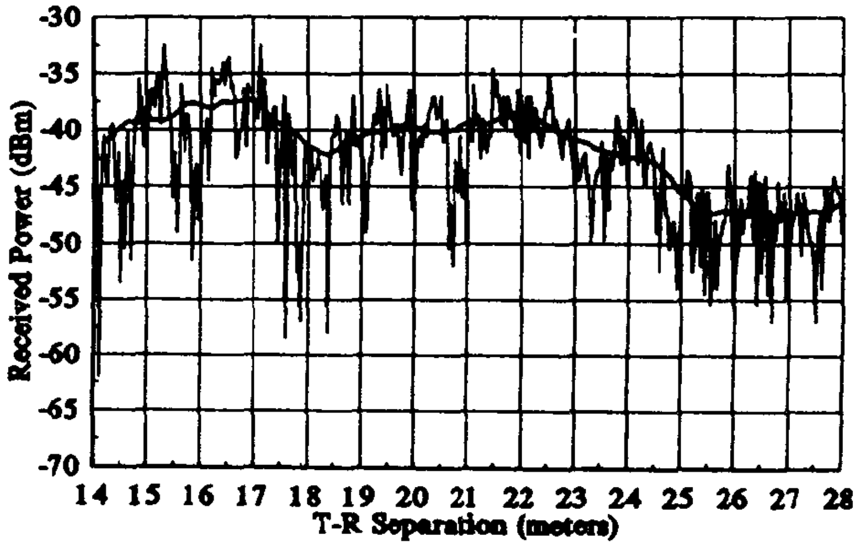

Figure 3.1
Small-scale and large-scale fading.

$$P_r(d) = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2 L} \tag{3.1}$$

where P_t is the transmitted power, $P_r(d)$ is the received power which is a function of the T-R separation, G_t is the transmitter antenna gain, G_r is the receiver antenna gain, d is the T-R separation distance in meters, L is the system loss factor not related to propagation ($L \geq 1$), and λ is the wavelength in meters. The gain of an antenna is related to its effective aperture, A_e, by

$$G = \frac{4\pi A_e}{\lambda^2} \tag{3.2}$$

The effective aperture A_e is related to the physical size of the antenna, and λ is related to the carrier frequency by

$$\lambda = \frac{c}{f} = \frac{2\pi c}{\omega_c} \tag{3.3}$$

where f is the carrier frequency in Hertz, ω_c is the carrier frequency in radians per second, and c is the speed of light given in meters/s. The values for P_t and P_r must be expressed in the same units, and G_t and G_r are dimensionless quantities. The miscellaneous losses L ($L \geq 1$) are usually due to transmission line attenuation, filter losses, and antenna losses in the communication system. A value of $L = 1$ indicates no loss in the system hardware.

The Friis free space equation of (3.1) shows that the received power falls off as the square of the T-R separation distance. This implies that the received power decays with distance at a rate of 20 dB/decade.

An *isotropic* radiator is an ideal antenna which radiates power with unit gain uniformly in all directions, and is often used to reference antenna gains in wireless systems. The *effective isotropic radiated power (EIRP)* is defined as

$$EIRP = P_t G_t \tag{3.4}$$

and represents the maximum radiated power available from a transmitter in the direction of maximum antenna gain, as compared to an isotropic radiator.

In practice, *effective radiated power (ERP)* is used instead of EIRP to denote the maximum radiated power as compared to a half-wave dipole antenna (instead of an isotropic antenna). Since a dipole antenna has a gain of 1.64 (2.15 dB above an isotrope), the ERP will be 2.15 dB smaller than the EIRP for the same transmission system. In practice, antenna gains are given in units of dBi (dB gain with respect to an isotropic source) or dBd (dB gain with respect to a half-wave dipole) [Stu81].

The *path loss*, which represents signal attenuation as a positive quantity measured in dB, is defined as the difference (in dB) between the effective transmitted power and the received power, and may or may not include the effect of the antenna gains. The path loss for the free space model when antenna gains are included is given by

$$PL\,(\mathrm{dB}) = 10\log\frac{P_t}{P_r} = -10\log\left[\frac{G_t G_r \lambda^2}{(4\pi)^2 d^2}\right] \tag{3.5}$$

When antenna gains are excluded, the antennas are assumed to have unity gain, and path loss is given by

$$PL\,(\mathrm{dB}) = 10\log\frac{P_t}{P_r} = -10\log\left[\frac{\lambda^2}{(4\pi)^2 d^2}\right] \tag{3.6}$$

The Friis free space model is only a valid predictor for P_r for values of d which are in the far-field of the transmitting antenna. The far-field, or *Fraunhofer region*, of a transmitting antenna is defined as the region beyond the far-field distance d_f, which is related to the largest linear dimension of the transmitter antenna aperture and the carrier wavelength. The Fraunhofer distance is given by

$$d_f = \frac{2D^2}{\lambda} \tag{3.7.a}$$

where D is the largest physical linear dimension of the antenna. Additionally, to be in the far-field region, d_f must satisfy

$$d_f \gg D \tag{3.7.b}$$

and

$$d_f \gg \lambda \tag{3.7.c}$$

Furthermore, it is clear that equation (3.1) does not hold for $d = 0$. For this reason, large-scale propagation models use a close-in distance, d_0, as a known received power reference point. The received power, $P_r(d)$, at any distance $d > d_0$, may be related to P_r at d_0. The value $P_r(d_0)$ may be predicted from equation (3.1), or may be measured in the radio environment by taking the average received power at many points located at a close-in radial distance d_0 from the transmitter. The reference distance must be chosen such that it lies in the far-field region, that is, $d_0 \geq d_f$, and d_0 is chosen to be smaller than any practical distance used in the mobile communication system. Thus, using equation (3.1), the received power in free space at a distance greater than d_0 is given by

$$P_r(d) = P_r(d_0)\left(\frac{d_0}{d}\right)^2 \qquad d \geq d_0 \geq d_f \tag{3.8}$$

In mobile radio systems, it is not uncommon to find that P_r may change by many orders of magnitude over a typical coverage area of several square kilometers. Because of the large dynamic range of received power levels, often dBm or dBW units are used to express received power levels. Equation (3.8) may be expressed in units of dBm or dBW by simply taking the logarithm of both sides and multiplying by 10. For example, if P_r is in units of dBm, the received power is given by

$$P_r(d)\ \text{dBm} = 10\log\left[\frac{P_r(d_0)}{0.001\ \text{W}}\right] + 20\log\left(\frac{d_0}{d}\right) \qquad d \geq d_0 \geq d_f \tag{3.9}$$

where $P_r(d_0)$ is in units of watts.

The reference distance d_0 for practical systems using low-gain antennas in the 1-2 GHz region is typically chosen to be 1 m in indoor environments and 100 m or 1 km in outdoor environments, so that the numerator in equations (3.8) and (3.9) is a multiple of 10. This makes path loss computations easy in dB units.

Example 3.1

Find the far-field distance for an antenna with maximum dimension of 1 m and operating frequency of 900 MHz.

Solution to Example 3.1

Given:

Largest dimension of antenna, $D = 1$ m

Operating frequency $f = 900$ MHz, $\lambda = c/f = \dfrac{3 \times 10^8\ \text{m/s}}{900 \times 10^6\ \text{Hz}}$ m

Using equation (3.7.a), far-field distance is obtained as

$$d_f = \frac{2(1)^2}{0.33} = 6 \text{ m}$$

Example 3.2

If a transmitter produces 50 watts of power, express the transmit power in units of (a) dBm, and (b) dBW. If 50 watts is applied to a unity gain antenna with a 900 MHz carrier frequency, find the received power in dBm at a free space distance of 100 m from the antenna. What is $P_r(10 \text{ km})$? Assume unity gain for the receiver antenna.

Solution to Example 3.2

Given:
Transmitter power, $P_t = 50$ W.
Carrier frequency, $f_c = 900$ MHz

Using equation (3.9),
(a) Transmitter power,

$$\begin{aligned} P_t(\text{dBm}) &= 10\log[P_t(\text{mW})/(1 \text{ mW})] \\ &= 10\log[50 \times 10^3] = 47.0 \text{ dBm}. \end{aligned}$$

(b) Transmitter power,

$$\begin{aligned} P_t(\text{dBW}) &= 10\log[P_t(\text{W})/(1 \text{ W})] \\ &= 10\log[50] = 17.0 \text{ dBW}. \end{aligned}$$

The received power can be determined using equation (3.1).

$$P_r = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2 L} = \frac{50(1)(1)(1/3)^2}{(4\pi)^2(100)^2(1)} = 3.5 \times 10^{-6} \text{ W} = 3.5 \times 10^{-3} \text{ mW}$$

$$P_r(\text{dBm}) = 10\log P_r(\text{mW}) = 10\log\left(3.5 \times 10^{-3} \text{ mW}\right) = -24.5 \text{ dBm}.$$

The received power at 10 km can be expressed in terms of dBm using equation (3.9), where $d_0 = 100$ m and $d = 10$ km

$$\begin{aligned} P_r(10 \text{ km}) &= P_r(100) + 20\log\left[\frac{100}{10000}\right] = -24.5 \text{ dBm} - 40 \text{ dB} \\ &= -64.5 \text{ dBm}. \end{aligned}$$

3.3 Relating Power to Electric Field

The free space path loss model of Section 3.2 is readily derived from first principles. It can be proven that any radiating structure produces electric and magnetic fields [Gri87], [Kra50]. Consider a small linear radiator of length L, that is placed coincident with the z-axis and has its center at the origin, as shown in Figure 3.2.

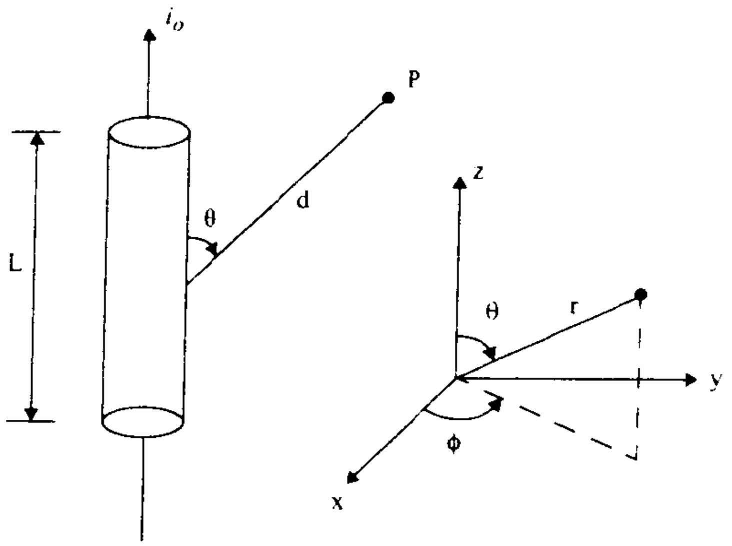

Figure 3.2
Illustration of a linear radiator of length L $(L \ll \lambda)$, carrying a current of amplitude i_0 and making an angle θ with a point, at distance d.

If a current flows through such an antenna, it launches electric and magnetic fields that can be expressed as

$$E_r = \frac{i_0 L \cos\theta}{2\pi\varepsilon_0 c}\left\{\frac{1}{d^2} + \frac{c}{j\omega_c d^3}\right\} e^{j\omega_c(t-d/c)} \tag{3.10}$$

$$E_\theta = \frac{i_0 L \sin\theta}{4\pi\varepsilon_0 c^2}\left\{\frac{j\omega_c}{d} + \frac{c}{d^2} + \frac{c^2}{j\omega_c d^3}\right\} e^{-j\omega_c(t-d/c)} \tag{3.11}$$

$$H_\phi = \frac{i_0 L \sin\theta}{4\pi c}\left\{\frac{j\omega_c}{d} + \frac{c}{d^2}\right\} e^{j\omega_c(t-d/c)} \tag{3.12}$$

with $E_\phi = H_r = H_\theta = 0$. In the above equations, all $1/d$ terms represent the radiation field component, all $1/d^2$ terms represent the induction field component, and all $1/d^3$ terms represent the electrostatic field component. As seen from equations (3.10) to (3.12), the electrostatic and inductive fields decay much faster with distance than the radiation field. At regions far away from the transmitter (far-field region), the electrostatic and inductive fields become negligible and only the radiated field components of E_θ and H_ϕ need be considered.

In free space, the *power flux density* P_d (expressed in W/m^2) is given by

$$P_d = \frac{EIRP}{4\pi d^2} = \frac{P_t G_t}{4\pi d^2} = \frac{E^2}{R_{fs}} = \frac{E^2}{\eta} \text{ W/m}^2 \tag{3.13}$$

where R_{fs} is the intrinsic impedance of free space given by $\eta = 120\pi\ \Omega$ (377Ω). Thus, the power flux density is

$$P_d = \frac{|E|^2}{377\Omega} \text{ W/m}^2 \tag{3.14}$$

where $|E|$ represents the magnitude of the radiating portion of the electric field in the far field. Figure 3.3a illustrates how the power flux density disperses in free space from an isotropic point source. P_d may be thought of as the *EIRP* divided by the surface area of a sphere with radius d. The power received at distance d, $P_r(d)$, is given by the power flux density times the effective aperture of the receiver antenna, and can be related to the electric field using equations (3.1), (3.2), (3.13), and (3.14).

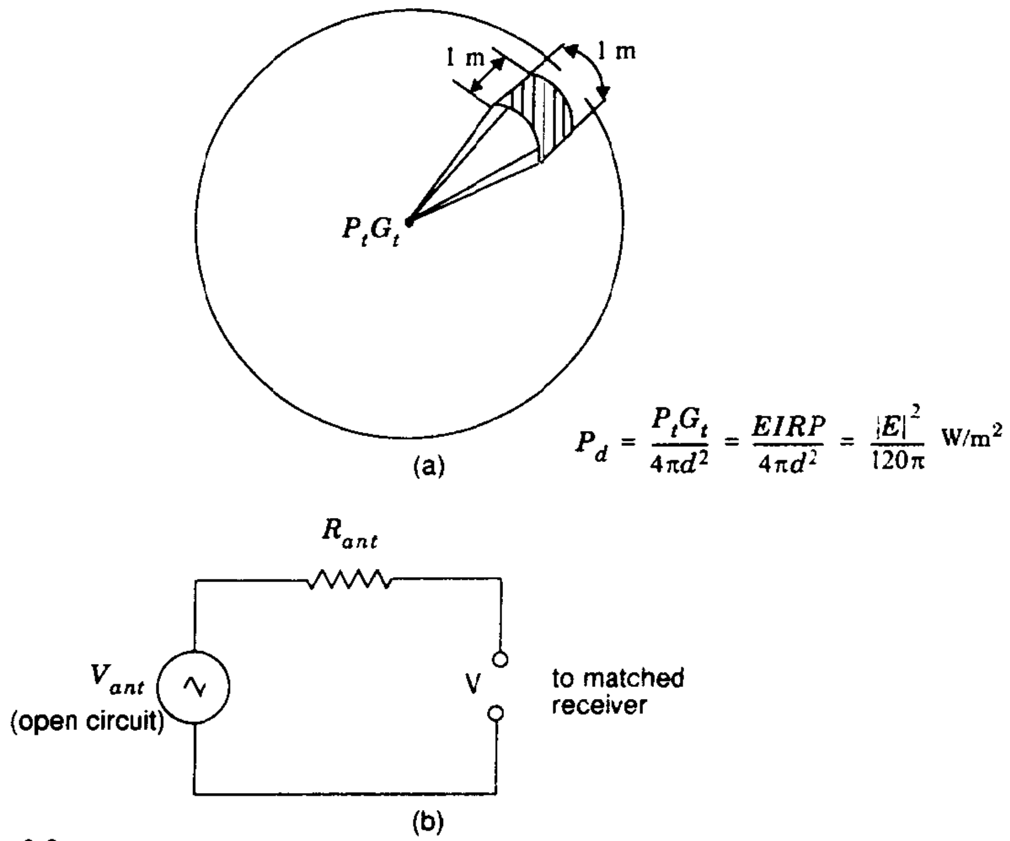

Figure 3.3
(a) Power flux density at a distance d from a point source.
(b) Model for voltage applied to the input of a receiver.

$$P_r(d) = P_dA_e = \frac{|E|^2}{120\pi}A_e = \frac{P_tG_tG_r\lambda^2}{(4\pi)^2d^2} \text{ Watts} \tag{3.15}$$

Equation (3.15) relates electric field (with units of V/m) to received power (with units of watts), and is identical to equation (3.1) with $L = 1$.

Often it is useful to relate the received power level to a receiver input voltage, as well as to an induced E-field at the receiver antenna. If the receiver antenna is modeled as a matched resistive load to the receiver, then the receiver antenna will induce an rms voltage into the receiver which is half of the open circuit voltage at the antenna. Thus, if V is the rms voltage at the input of a receiver (measured by a high impedance voltmeter), and R_{ant} is the resistance of the matched receiver, the received power is given by

$$P_r(d) = \frac{V^2}{R_{ant}} = \frac{[V_{ant}/2]^2}{R_{ant}} = \frac{V_{ant}^2}{4R_{ant}} \tag{3.16}$$

Through equations (3.14) to (3.16), it is possible to relate the received power to the received E-field or the open circuit rms voltage at the receiver antenna terminals. Figure 3.3b illustrates an equivalent circuit model. Note $V_{ant} = V$ when there is no load.

Example 3.3

Assume a receiver is located 10 km from a 50 W transmitter. The carrier frequency is 900 MHz, free space propagation is assumed, $G_t = 1$, and $G_r = 2$, find (a) the power at the receiver, (b) the magnitude of the E-field at the receiver antenna (c) the rms voltage applied to the receiver input assuming that the receiver antenna has a purely real impedance of 50 Ω and is matched to the receiver.

Solution to Example 3.3

Given:

Transmitter power, $P_t = 50$ W
Carrier frequency, $f_c = 900$ MHz
Transmitter antenna gain, $G_t = 1$
Receiver antenna gain, $G_r = 2$
Receiver antenna resistance = 50 Ω

(a) Using equation (3.5), the power received at a distance $d = 10$ km is

$$P_r(d) = 10\log\left(\frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2}\right) = 10\log\left(\frac{50 \times 1 \times 2 \times (1/3)^2}{(4\pi)^2 10000^2}\right)$$

$$= -91.5 \text{ dBW} = -61.5 \text{ dBm}$$

(b) Using equation (3.15), the magnitude of the received E-field is

$$|E| = \sqrt{\frac{P_r(d)120\pi}{A_e}} = \sqrt{\frac{P_r(d)120\pi}{G_r\lambda^2/4\pi}} = \sqrt{\frac{7 \times 10^{-10} \times 120\pi}{2 \times 0.33^2/4\pi}} = 0.0039 \text{ V/m}$$

(c) Using equation (3.16), the open circuit rms voltage at the receiver input is

$$V_{ant} = \sqrt{P_r(d) \times 4R_{ant}} = \sqrt{7 \times 10^{-10} \times 4 \times 50} = 0.374 \text{ mV}$$