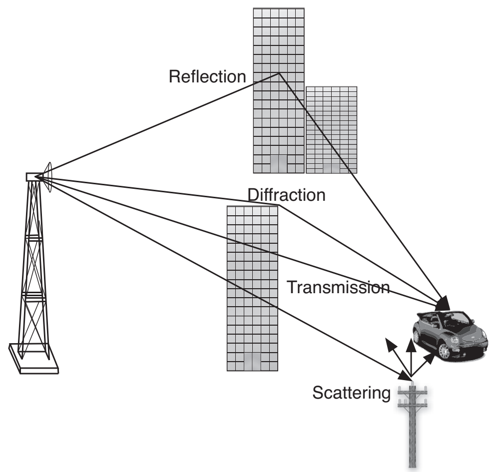


Figure 13.5 Mechanisms of radio propagation.

elements follows some geometric pattern (example, linearly spaced elements, elements on a rectangular grid or elements placed on a circle) to achieve the requirements. The spacing of antenna elements is once again measured in fractions or multiples of λ. In recent years, it has been possible to electronically shift the phases and amplitudes of signals entering the antenna elements to *steer* the beams created by arrays dynamically without mechanically moving the antenna itself. The interested reader is referred to [1] and [2] for a detailed mathematical treatment of antennas and antenna arrays.

13.3 RADIO PROPAGATION

In electrical wires and fiber optic lines we have well defined models for the attenuation of a signal as it propagates along the media. This is not true for the wireless environment, it is fairly difficult to understand the nature of electromagnetic wave propagation in complex environments. Maxwell's equations, discussed in Appendix B can only be solved for simple geometries and material with homogeneous properties. Hence approximations, empirical characterization and statistical models are necessary to predict radio wave propagation in all cases. In Section subsec:radpropmechanisms, we describe elementary mechanisms of propagation qualitatively. Different types of fading are discussed in the other two subsections.

13.3.1 Mechanisms

Radio waves suffer attenuation and dispersion like light waves. At frequencies greater than 500 MHz, radio wave propagation can be approximated as ray propagation in a manner similar to optics (see Chapter 12). There are three basic mechanisms by which a radio signal can propagate from the transmitter to the receiver as shown in Figure 13.5—namely reflection/transmission, diffraction and scattering. A combination of these mechanisms leads to *multipath propagation*.

Reflection and transmission of radio "rays" are the most common forms of propagation and dominant in indoor areas. The signal can go through "transparent" or "translucent" objects or bounce off objects (with some attenuation in either case). Recall from Chapter 12 that all the power in the incident ray is not transferred to the reflected ray because of some transmission and some absorption.

Reflection and transmission occur when the intervening objects are much larger than the wavelength of the carrier. Examples of reflectors are the ground, walls of buildings and varying atmospheric layers. For example, ionospheric waves[2] can be reflected between the upper atmosphere and the ground and propagate for thousands of miles. In general, radio waves are transmitted through the atmosphere or through walls and other objects. Attenuation of signals due to reflection or transmission depends upon the angle of incidence (whether the wave impinges on the object directly or grazes it), the frequency of the carrier, the nature of the surface (rough or smooth) and the properties of the material making up the object. Usually, transmission through an object leads to larger losses due to absorption than due to reflection off an object. However multiple reflections can result in weak signals. As an example, a signal at 60 GHz is severely attenuated if it travels through oxygen, but this attenuation can be worse if it is also raining.

Diffraction is the phenomenon that occurs when a radio signal is incident upon the edge of a sharp object (such as the roof of a building, wall edge, or door). the edge of such an object becomes a *secondary source of transmission* (almost like an antenna). However, the resulting loss in power can be significant. In micro-cellular areas, where mobile stations can be in the shadow of buildings, diffraction at roof edges is an important mechanism for radio propagation, but diffraction is not a significant means of propagation in indoor areas (except perhaps across multiple floors where the signal can diffract at a window edge and travel upwards or downwards rather than through floors of a building).

Scattering occurs when a radio signal impinges upon an irregular object comparable in size to the wavelength. Foliage, furniture, lamp posts, and vehicles are examples of scatterers. Such objects scatter rays in all directions (in essence each scatterer again acts as a secondary source). Scattering results in signal propagation in all directions albeit with large losses in signal strength. Scattering is thus not a dominant mechanism for radio propagation except in very cluttered environments.

Multipath Propagation and Fading. In general, a transmitted radio signal can reach the receiver through a multiplicity of paths and this is called multipath propagation. The signal arriving along one path may have been diffracted once, reflected off two objects and travelled through another. A signal arriving along a second path may have only been reflected off several objects. A third signal may be the result of yet an entirely different path. The power in these signals will be different, as will the length of the paths and thus the time taken by these signals to arrive at receiver. Unlike wired channels where signal strength is relatively stationary and predictable, in wireless channels the signal strength

2 The ionosphere is a region in the atmosphere (around 50 to 300 miles above the Earth's surface) with many layers of charged particles. EM waves can propagate through the ionosphere much like how light travels through a fiber.

varies with time and location because of such multipath propagation and a dynamically changing environment. The result is that the receiver sees a "combined" signal comprised of individual signals arriving along different paths with different delays and amplitudes. The received signal depends on the location and the speed of the transmitter, the receiver and all objects in the intervening environment.

As we will see next, this results in a complex time varying signal with varying amplitudes and phases. Since the amplitude of the combined signal fluctuates, we refer to the signal as "fading." Two main issues that arise because of fading are (a) how far can the signal propagate before its reliable reception is not possible and (b) how is the bit error rate (BER) of the signal impacted and how we can reduce it by proper receiver design. The former relates to network design and deployment. In order to predict signal strength for cell planning and deployment various analytical models have been proposed and are still being developed. These "large scale" (LS) models are so-called propagation path-loss models which focus on the power loss over the communications path from transmitter to receiver. They predict the *mean* signal strength as a function of the transmitter-receiver separation distance d. The latter "small-scale" (SS) models predict random fluctuations in signal amplitude over a short distance or short time duration around a fixed transmitter-receiver separation distance d. They also provide some idea of the degradation in signal quality and possible methods to mitigate the degradation.

13.3.2 Large-Scale Fading and Path Loss Models

Large-scale (LS) fading considers the *average* variation of received signal strength (RSS) or power as a function of the distance d between the transmitter and the receiver. The average is computed over short periods of time or over short distances around d. A plot of such average RSS values as a function of the logarithm of d is used to predict how much the signal strength drops with distance. If possible, a linear regression fit to the measured average values of the RSS is obtained. The linear fit is called the *path-loss model* and the variation around the fit is called the *large-scale or shadow fading*. These aspects are illustrated in Figure 13.6 (a).

In the LS cases one focuses on the "path-loss" L_p which represents the signal attenuation as a positive quantity measured in decibels. The path-loss and the associated LS models are used to compute the link power budget described later in Section 13.5. The path loss is defined as the difference between the effective transmitted power and the average received power and may or may not include the effect of antenna gains. Thus the path loss in dB is defined as:

$$\text{Path Loss} = L_p(dB) = 10\log_{10}\left[\frac{P_t}{P_r}\right] \tag{13.4}$$

where P_t is the power of the transmitter and P_r is the power of the received signal. Note that P_t and P_r are normally expressed in watts (W), but due to the large range the signals may take, they maybe expressed in decibel watts (dBW) or decibel milliwatts (dBm).

Example. For P_r, $P_r(\text{dBW}) = 10\log\left[\frac{P_r(\text{W})}{1(\text{W})}\right]$ or $P_r(\text{dBm}) = 10\log\left[\frac{P_r(\text{W})}{.001(\text{W})}\right]$.

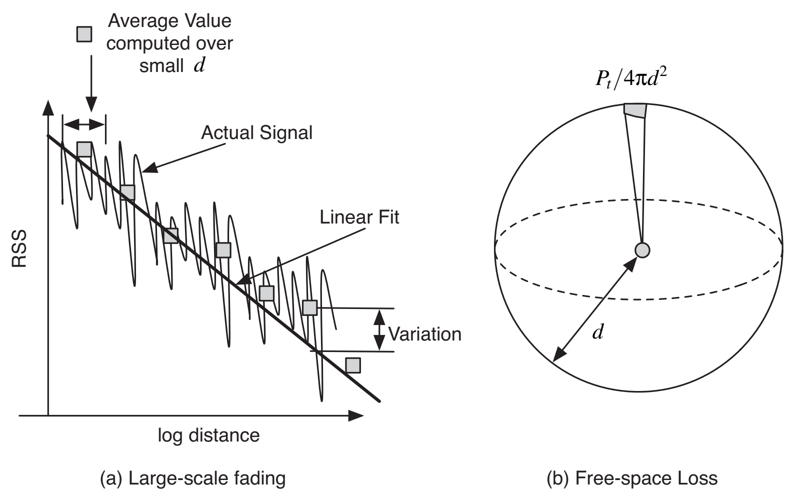


Figure 13.6 Large-scale fading and free-space loss.

Example. The transmitted power is 100 mW and the received signal is 1 microwatt. $L_p = 10 \times \log\left(\frac{100}{0.001}\right) = 50$ dB.

Free-space propagation. Free space loss models a best-case scenario where the transmitter and reciever have a clear unobstructed *line of sight* path between them and both antennas are surrounded by empty space with no other object close enough to interact. In this situation the only propagation effect is the natural spreading out of the radiowave at greater and greater distances from the transmitter (think of an expanding sphere). Consider an isotropic antenna that radiates equally in all directions. If it is a point source, we can represent it as shown in Figure 13.6 (b). Let the transmit power be P_t. At a distance d from the source, the area of a sphere enclosing the source is $4\pi d^2$. The power is uniformly distributed on this sphere and so the *power density* is $\frac{P_t}{(4\pi d^2)}$. Consider an isotropic receive antenna that is placed at a distance d from the source. Its effective area is $\frac{\lambda^2}{4\pi}$. The power it can capture is the product of the effective area and the power density. Thus the received power is:

$$P_r = \frac{P_t}{4\pi d^2} \times \frac{\lambda^2}{4\pi} = \frac{P_t \lambda^2}{(4\pi d)^2} \tag{13.5}$$

In dB, the relationship between P_t and P_r can be written as:

$$P_r(dBm) = P_t(dBm) - 21.98 + 20\log_{10}\lambda - 20\log_{10} d \tag{13.6}$$

We can generalize this to the case where there are real antennas. If G_t is the gain of the transmitter antenna and G_r is the gain of the receiver antenna,

$$P_r = P_t G_t G_r \frac{\lambda^2}{(4\pi d)^2} \tag{13.7}$$

From (13.7) we have the path loss for free space $L_{p,fs}$ in dB as

$$L_{p,fs} = 10\log_{10}\frac{P_t}{P_r} = 10\log_{10}\left[\frac{G_t G_r \lambda^2}{(4\pi d)^2}\right] \tag{13.8}$$

From this formula one can see that in the best case scenario the signal decays 20 dB per decade increase in d. Also, one can see that the path loss increases with λ at a rate of 20 dB per decade decrease in λ. If the carrier frequency is fixed, the path-loss is some constant minus a term comprising of $20\log_{10} d$. As $\log_{10} d$ increases, the path-loss also increases, but by a factor of 20. In general, the path-loss fit has a term of the form $10n\log_{10} d$. The quantity n is called the *path-loss exponent or gradient*. Note $n = 2$ in (13.5), (13.6) and (13.7).

Example. A cell site operating at 900 MHz produces 50 W of power and the system uses unity gain antennas at transmitter and reciever. Assuming free space path loss what is the P_r in dBm at 100 m and 10 km.

$$\begin{aligned}
P_r &= \frac{P_t G_t G_r \lambda^2}{(4\pi d)^2} = \frac{(50)(1)(1)\left(\frac{3\times10^8}{900\times10^6}\right)^2}{(4\pi)^2(100)^2} = 3.5\times10^{-6} \\
P_r(dBm) &= 10\log_{10}\left[\frac{P_r}{0.001}\right] = -24.5 \text{ dBm} \\
P_r(10\,\text{km}) &= P_r(100) + 20\log_{10}\left[\frac{100}{10,000}\right] \\
&= -24.5\,\text{dBm} - 40\,\text{dB} = -64.5\,\text{dBm}
\end{aligned}$$

Example. Compare the free space path loss at 1 km and 5 km of two signals one in the lower cellular band at 880 MHz and the other in the upper cellular band at 1960 MHz. Assume unity gain antennas are used.

Distance	880 MHz	1960 MHz
1 km	91.29	98.25
5 km	105.27	112.23

Thus the upper cellular band has at least a 7 dB greater path loss in the free space case.

Other Simple Propagation Path Loss Models. In reality, the path loss depends on a complex variety of factors, such as the antenna heights, frequency used and immediate environment (signs, cars, etc). A simple model that is often used as a first approximation has the form:

$$P_r = P_t C d^{-n} \tag{13.9}$$

where C is a constant that depends on the frequency used and n is the path loss exponent. For example, for a free space environment $C = \frac{\lambda^2}{(4\pi)}$ and $n = 2$. The path loss exponent is ofen estimated from measurement data or determined based on the environment (e.g., open area $n = 2.2$, dense urban area $n = 5$), with typical values ranging from 2.2 to 5.5. Note, that in terms of dB the path loss L_p will have the form:

$$L_p = L_0 + 10n \log_{10}(d) \tag{13.10}$$

One can see that the path loss will increase at a rate of $10n$ per decade increase in d. The constant L_0 is often a function of the carrier frequency. It is large for higher frequencies implying a greater path-loss as the carrier frequency increases.

Another class of path loss models that are widely used in practice are empirical models which involve curve fitting to measurement data collected in typical environments along specific paths in that environment. In the case of cell-phone services, it is common to drive a vehicle equipped with a variety of radio signal measurement devices and a global-positioning system to collect such data. This process is called a "measurement test-drive". Two of the most popular measurement based models for determining path loss are the Okumura-Hata model and COST-231-Hata model which are given by equations (13.11) and (13.12) below. These two models estimate the path loss in dB. The Okumura-Hata model is tailored to the cellular band around 800 MHz, whereas the Cost-231-Hata model is specified for the personal communications system (PCS) band around 1900 MHz.

Okumura-Hata

$$\begin{aligned} L_{p,oh} = {} & 69.55 + 26.16 \log(f) - 13.82 \log(h) \\ & + [44.9 - 6.55 \log(h)] \times \log(d) + C \end{aligned} \tag{13.11}$$

Cost-231-Hata

$$\begin{aligned} L_{p,c2h} = {} & 46.3 + 33.9 \log(f) - 13.82 \log(h) \\ & + [44.9 - 6.55 \log(h)] \times \log(d) + C \end{aligned} \tag{13.12}$$

where f is the frequency in MHz, d the transmitter-receiver seperation in km, h is the effective height of transmitter antenna in meters (i.e., the difference between height of transmitter antenna and height of receiver antenna) and C is an environmental correction factor in dB as shown below. All logarithms are to the base 10.

Environment	Okumura-Hata	Cost-231-Hata
Dense Urban	−2	0
Urban	−5	−5
Suburban	−10	−10
Rural	−26	−17

The environmental factor roughly adjusts for the differences in environment due to obstructions, ground reflections, etc. All things being equal a PCS band base station will have a smaller coverage area than a cellular band base station due to the propagation effects at higher frequency.

Example. You are designing a cellular system for greater Pittsburgh for a carrier with a set of 416 radio channels in the A block frequency band (824–835, 845–846.5 MHz for the uplink, 869–880, 890–891.5 MHz for forward or downlink). The base stations to be deployed produce 40 W of power and use 10-dB gain omnidirectional antennas. The mobile handsets require a signal level of −90 dBm and use antennas that produce a 3-dB gain. The base stations will be located such that the average effective height difference between the cell site antenna and the user is 100 m. Using the Okumura-Hata propagation model determine the maximum radius for a cell in an urban environment.

The worst case propagation is at the maximum frequency in the band, namely 891.5 MHz. Determining the path loss from (13.11) we get

$$\begin{aligned} L_{p,oh} &= 69.55 + 26.16\log(891.5) - 13.82\log(100) \\ &\qquad + [44.9 - 6.55\log(100)] \times \log(d) - 5 \\ &= 114.0852 + 31.8\log(d) \end{aligned}$$

The received power at the mobile P_r must be greater than −90 dBm and is given by

$$P_r = P_t + G_t + G_r - PL_{oh}$$

and results in

$$\begin{aligned} -90 &\leq 10\log\left(\frac{40}{.001}\right) + 10 + 3 - (114.0852 + 31.8\log(d)) \\ -34.935 &\leq -31.8\log(d) \\ 12.549 &\geq d \end{aligned}$$

Thus the maximum cell size is about 12.5 km.

In indoor areas, a *partition-dependent* path-loss model is commonly used. Given the transmitter and receiver locations, a straight line is drawn connecting the two points. The straight line intersects walls, floors and other obstructions that lie in between the transmitter and receiver. For each such obstruction, a loss is added (for example, 3 dB for a cubicle soft wall, 10 dB for a brick wall). Finally, the free-space loss is added to determine the overall path-loss. In the general case where there are N_u walls of type u with loss L_u and N_v floors of type v with loss L_v between a transmitter and receiver separated by a distance d, the path-loss will be given by:

$$L_p = L_0 + 20\log d + \sum_u N_u L_u + \sum_v N_v L_v \qquad (13.13)$$

In our discussions, we have considered specific path-loss models for cellular systems at 800 MHz, cellular (PCS) systems at 1900 MHz, and indoor wireless systems. As the propagation characteristics are very different in different areas, it is common to use models that are site or environment specific. Similarly, there are path-loss models for hilly areas, urban areas, suburban areas, microcellular systems that use transmitters mounted on lampposts and roofs of buildings, open areas, rural areas, and so on. Path-loss models that take into account detailed terrain data are also available and used in many cases.

Shadowing. As noted above large scale propagation models predict the mean signal strength at a transmitter-receiver separation d. The local mean received power $P_r(d)$ at a transmitter-receiver separation d is determined in part by the local environment of buildings, trees, cars, etc. One approach to modeling these shadowing effects is a statistical model where a random component is added to the mean received power predicted by a path loss model. This results in:

$$P_r(d) = P_t - L_p(d) + X_\sigma \tag{13.14}$$

where P_t is the transmitter power, $L_p(d)$ is the path loss—which is determined from one of the path loss models presented previously, and X_σ is the shadowing loss. Measurement studies have shown that in cellular systems the shadowing has a lognormal distribution in terms of variations in the signal strength, which results in a Normal distribution with mean zero and standard deviation σ for the term X_σ in dB. The value of σ at a particular location is usually determined by measurements, with typical values being: rural 3 dB, suburban 6 dB, urban, 8 dB, dense urban 10 dB. Since X_σ has a normal distribution and the other terms on the right hand sided of (13.14) are deterministic the overall received power $P_r(d)$ has a normal distribution with mean equal to $P_t - L_p(d)$ and standard deviation σ. This fact is often used to determine the % or probability of radio coverage at the edge of a cell. This amounts to determining the probability of the received signal power is above a given threshold T at distance d, which is given by $P\{P_r(d) \geq T\}$. The threshold value T is usually the minimum signal strength requirements of the mobile terminals. Since $P_r(d)$ has a normal distribution with mean of $P_t - L_p(d)$ the probability of being above a certain level can be found from the standard normal (0,1) random variable Z.

$$P\left\{P_r(d) \geq T\right\} = P\left\{Z \geq \frac{T - (P_t - L_p(d))}{\sigma}\right\} \tag{13.15}$$

In practice one is more interested in determining a path loss requirement which must be met to ensure coverage at a threshold value T for a given distance d, this additional path loss requirement is termed the shadow margin (SM). From (13.14) a particular quantile value such at .9 or .95 can be found, in fact these values depend only on σ and are determined from the standard normal random variable as .9 quantile $\rightarrow$ $SM = 1.282\sigma$ dB and .95 quantile $\rightarrow$ $SM = 1.645\sigma$ dB.

Example. For the cellular system in the previous example, measurements show that lognormal shadowing occurs in each cell with a standard deviation of 8 dB. If we require that 90% coverage area in each cell at the edge meet the −90 dBm signal level what must the mean recieved signal level (i.e., what is the shadow margin required)? Using the SM results resize the maximum cell radius using the Okumura-Hata propagation model.

From above for 90% coverage, we want the 0.9 quantile level of the normal distribution which results in a $SM = 1.282\sigma = 10.26$ dB. Adding in the SM requirements results in

$$\begin{aligned}P_r &= P_t + G_T + G_r - PL_{oh} - SM\\ -90 &\leq 10\log\left(\frac{40}{.001}\right) + 10 + 3 - (114.0852 + 31.8\log(d)) - 10.24\\ -24.69 &\leq -31.8\log(d)\\ 5.98 &\geq d\end{aligned}$$

Thus the maximum cell size is about 6 km.

Prediction/Planning Tools. Experts often make the analogy between radio propagation prediction and weather forcasting in that they have about the same accuracy. This analogy also holds true in being location dependent since just as there are places where it is possible to accurately predict the weather (e.g., southern California) and places where it is difficult (Pittsburgh), the same holds true for predicting radio propagation. In general before deploying a cell, service providers usually put up a test base station and take extensive measurements to actually determine the coverage and interference. Even after deployment, measurement data is continually gathered from operational systems in order to tune the system performance. In practice, companies also use a variety of computer aided design (CAD) tools for prediction and planning for both wireless LAN and cellular systems. For cellular networks the tools incorporate a terrain/geographic database of the area along with a traffic density overlay interms of population and vehicle traffic. The output of these tools is a map (in various colors) which shows radio coverage at specific levels (or probabilities) and interference values. In general these tools use a variety of prediction models similar to the ones discussed above (i.e., Free Space, Okumura-Hata, etc.). Another approach that is often used is ray tracing—where a series of rays are drawn from the base-station antenna to a particular point in the cell and from the geographic information the appropriate propagation model is used to predict the signal strength. Note some models allow for the determination of penetration into buildings and cars. Many of the CAD tools allow for the incorporation of measurement data to improve the accuracy of the models. Furthermore, most cellular tools allow for link budget computations to ensure signal strength at various distances.

13.3.3 Small-Scale Fading

As we will see in Chapter 14, most transmissions make use of sinusoidal carriers (of the form $A\cos(2\pi f_c t + \phi)$) of amplitude A, frequency f_c and phase ϕ. The transmit power in such signals is proportional to A^2. We can use the path-loss models from the previous section to determine how the average power drops as a function of distance. This works only as far as the average power is concerned (which, you must recall, is computed over a short distance or over a short time span). If we consider the signal characteristics over a smaller scale, we will see that the received sinusoidal carrier will NOT have a constant amplitude or phase. The reason for this is multipath propagation discussed below.

Consider two paths taken by a signal $s(t) = A\cos(2\pi f_c t + \phi)$. The first path incurs a delay of τ_1 and an attenuation of A_1. the second path incurs a delay of τ_2 and an attenuation

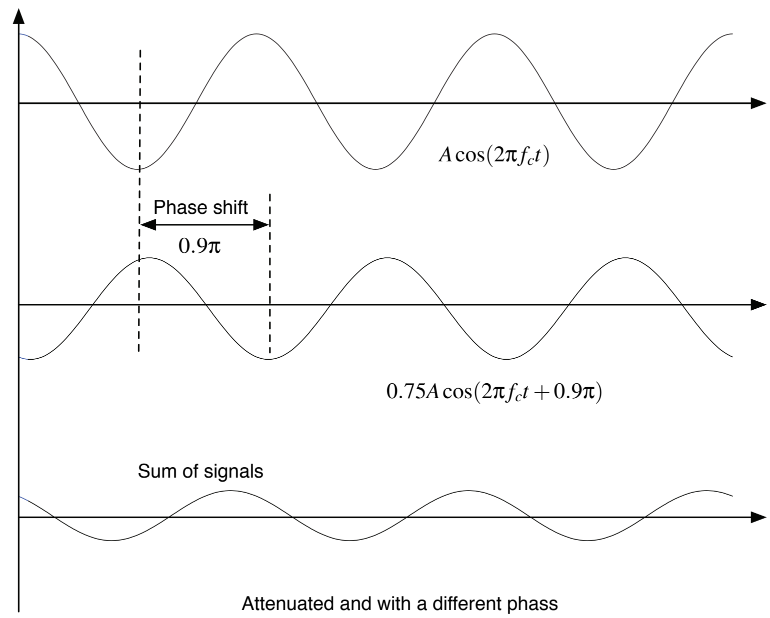


Figure 13.7 Illustration of fading.

of A_2. The received signal will be (ignoring noise):

$$\begin{aligned} r(t) &= AA_1 \cos(2\pi f_c(t - \tau_1) + \phi) + AA_2 \cos(2\pi f_c(t - \tau_2) + \phi) \\ &= B_1 \cos(2\pi f_c t + \theta_1) + B_2 \cos(2\pi f_c t + \theta_1) \end{aligned} \tag{13.16}$$

where $\theta_i = 2\pi f_c \tau_i + \phi$ and $B_i = AA_i$ for $i = 1, 2$. Recall from the discussion of phasors in Chapter 2 that the sum of two sinusoids can be written as a third sinusoid. The third sinusoid will have an amplitude and phase depending on the relative amplitudes and phases of the two sinusoids that are summing up (we can once again use phasor addition to compute the resulting amplitude and phase). The amplitude of the third sinusoid can be higher than the individual amplitudes or lower. It is also possible for two sinusoids to *cancel each other completely* (for example if $B_1 = B_2$ and $\theta_1 = \pi + \theta_2$). This phenomenon is shown in Figure 13.7 and is called *fading*.

In the general case, there will be several paths (say L) taken by the carrier to reach the receiver, so that the received signal will be:

$$\begin{aligned} r(t) &= \sum_{i=1}^{L} B_i \cos(2\pi f_c t + \theta_i) \\ &= \cos(2\pi f_c t) \sum_{i=0}^{L} B_i \cos(\theta_i) - \sin(2\pi f_c t) \sum_{i=0}^{L} B_i \sin(\theta_i) \\ &= X \cos(2\pi f_c t) + Y \sin(2\pi f_c t) \end{aligned} \tag{13.17}$$

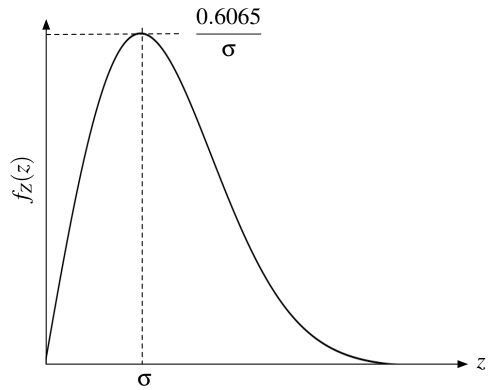


Figure 13.8 The Rayleigh probability density function.

The quantities B_i and θ_i can be assumed to be random, independent and identically distributed. The quantities X and Y are then the sum of several identically distributed and random quantities. From the central limit theorem, we can say that X and Y are normally distributed random variables. The envelope of $r(t)$ (which we are interested in) is given by $Z = \sqrt{X^2 + Y^2}$ and it is possible to show that it has a Rayleigh distribution. The probability density function (PDF) of Z is given by:

$$f_Z(z) = \frac{z}{\sigma^2} e^{\left(-z^2/2\sigma^2\right)}, \; z \geq 0 \tag{13.18}$$

Figure 13.8 shows the PDF of the Rayleigh distribution. If there is a dominant line-of-sight (LOS) path in addition to the L paths, the envelope of the signal takes on what is called a *Ricean* distribution. What this means is that the amplitude of the received signal can take on values that are random, but with probability distributions that can be evaluated. Measurements of signal amplitudes in LOS and non-LOS environments have shown that there is a close fit to the Ricean and Rayleigh distributions. Other distributions (such as log-normal and Nakagami) have also been found to have good fit with empirical measurements under certain conditions.

What is the impact of small-scale fading on wireless systems? Since the amplitude of the received signal is random, it is possible that there are periods of time when the signal has almost no power even though the average signal power over a larger period of time is sufficient for reliable reception. In such periods of "deep fade," almost no information can be recovered. If the transmission is analog, the signal can be barely heard (in the case of audio) or seen (in the case of video). Bursts of errors (bits being flipped) are seen with digital signals leading to high bit error rates. To reduce the bit error rate, transmit powers may have to be increased drastically.

The common method of overcoming the problem of bursts of errors is to use *diversity* where the same data is sent on multiple copies of signals that fade *independently*. There is a probability $p < 1$ that any one signal is in a deep fade. The probability that two of the signals are in deep fade will be $p^2 < p$. Thus, the bit error rate can be reduced through diversity. A common method used in cellular telephony is to use *receive antenna diversity*. Multiple (usually two or three) antennas are used at the base station to receive the signal from a cell-phone. The antennas are separated in space such that the signals they "see" are fading independently. Antenna diversity provides some gains in terms of reducing the required transmit power for reliable communication as we will see in Section 13.5. Another method of overcoming performance degradation due to bursts of errors without significantly increasing the transmit power is to use error control coding (see Chapter 8). Recall that error control coding introduces redundancies in the transmitted data for reliable transmission. You can think of such redundancies as a form of diversity. Interleaving, where the bits or symbols of a codeword are spread in time (to prevent multiple errors in the same codeword), is used to ensure that there is some independent fading of related bits.

Also of importance in understanding the impact of small-scale fading on performance are answers to the following questions: (a) Suppose the signal is in a deep fade—the amplitude is very low. How long can we expect the signal to be in fade? (b) Does the symbol rate have any relationship with the performance of the system in terms of BER? The answer to (a) can provide designers information about what sort of error control coding and interleaving depths are necessary. It can be shown that this *time variation* of the signal is dependent upon the speed at which the environment is changing and the carrier frequency of the signal. In the case of (b), it so happens that there is inter-symbol interference if the symbol durations are small compared to the delays between multipath components. This phenomenon called *time dispersion* can lead to unrecoverable errors at the receiver. We discuss the time variation and time dispersion of the channel below.

Time Variation. As discussed above, system designers are often interested in determining how quickly the signal envelope changes. For a time invariant channel, the signal envelope does not change with time. However, wireless communications occur through time-varying channels and signal envelopes can rapidly change as shown in Figure 13.9. The changes are random and as is the case with all random quantities, they are described through their statistics. Change is also relative (we have to define with respect to what)—so there is a need to specify a baseline. This baseline is usually the mean or RMS value of the signal envelope and/or certain levels above or below the mean/RMS value.

The rate of change of the signal is determined by what is known as the *Doppler Spectrum* of the signal. Let us suppose that the transmitter is fixed and the mobile receiver is moving at a speed of υ m/s. The maximum Doppler shift associated with a signal (carrier frequency f_c) is given by:

$$f_m = \frac{f_c \upsilon}{c} \tag{13.19}$$

where c is the speed of light. The average time for which the channel (and hence the signal characteristics) can be assumed to be constant is given by the *coherence time* T_c that is a function of f_m. A good approximation for T_c is:

$$T_c = \frac{9}{16\pi f_m} \tag{13.20}$$

Note that if a symbol carried by the signal is of duration T seconds, we can assume that it is not distorted if $T < T_c$. If however $T > T_c$, the amplitude of a symbol may have

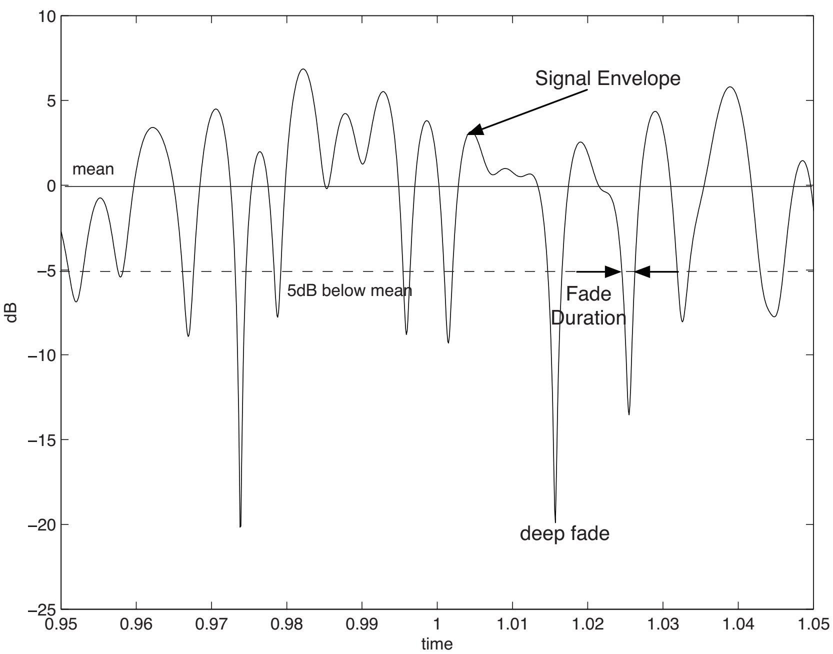


Figure 13.9 Time variation of signal envelope.

changed drastically over the symbol duration resulting in distortion that can lead to errors. A channel where $T < T_c$ is called *slow fading* because the channel is effectively constant over a symbol. If $T > T_c$, the channel is called *fast fading*.

Example. A vehicle is moving at 100 km/hour and communicating with a base station at a carrier frequency of 900 MHz. What is the maximum Doppler shift? What is the coherence time of the channel? What should the symbol rate be for the channel to be considered "slow fading"? Clearly $f_m = \frac{v f_c}{c} = \frac{100 \times 10^3 \times 900 \times 10^6}{(3 \times 10^8 \times 3600)} = 83.3$ Hz. From this, the coherence time is $T_c = \frac{9}{(16\pi f_m)} = 2.1$ ms. The symbol rate when $T = T_c$ is $R = \frac{1}{T_c} = \frac{1}{(2.1 \times 10^{-3})} \approx 500$ symbols/s. If the symbol rate is smaller than 500 symbols/s, the channel can be considered to be slow fading.

It is also possible to determine what is the average amount of time the signal envelope will be *below* a certain value smaller than the mean or RMS signal amplitude. This time is called the *fade duration*. A common assumption to make is that all symbols within the fade duration will be in error. This can help system designers decide how much of coding and interleaving is necessary. The *fade rate* is another quantity that tells us how many times (on average) that the signal will cross a certain level below the mean/RMS value. The

Table 13.1
Example power-delay profile.

Relative power in dB	*Relative delay in μs*
-10	0
0	1
-10	2
0	3
-20	4

expressions for the fade duration and fade rate are as follows:

$$\text{Fade Rate:} \quad N_R = \sqrt{2\pi}\, f_m\, \bar{r}\, \mathrm{e}^{-\bar{r}^2} \tag{13.21}$$

$$\text{Fade Duration:} \quad \tau = \frac{\mathrm{e}^{-\bar{r}^2} - 1}{\bar{r}\ f_m \sqrt{2\pi}} \tag{13.22}$$

In the above equations, $\bar{r} = \frac{r}{r_{rms}}$ is the ratio of the signal level at which we are considering the fade rate or duration (for example, in Figure 13.9, a level that is 5 dB below the mean is marked).

Time Dispersion. In all of the above discussions, we have assumed that multipath components arrive very closely spaced in time and that their relative delays are small compared to the symbol duration. This is true for signals where the symbol duration is large or the data rate and hence the bandwidth is small. Consequently, the phenomenon is sometimes called *narrowband fading*. The impact of multipath becomes an obstruction to the amount of data being sent when the symbol durations get smaller in comparison to the delays between the multipath components. We discuss this below.

Consider a system where two symbols are transmitted each lasting for T seconds. The signal arrives via two paths—the first path takes x seconds to reach the receiver and the second takes $x + 1.2T$ to reach the receiver. The *relative delay* between the two paths is $1.2T > T$. Let us suppose the receiver is trying to decode the symbols that have arrived from the first path. The first symbol is received at x seconds. It is processed till $x + T$ seconds. The second symbol arrives at $x + T$ seconds. While it is being processed, *the first symbol arrives again* at $x + 1.2T$ and *interferes* with the second symbol. This is called time dispersion of the channel—it describes how much a symbol is spread in time causing interference to subsequent symbols. The interference between symbols is called multipath-induced inter-symbol interference (ISI).

To characterize such channels, *wideband channel models* are often used. They specify the channel impulse response as consisting of several delta functions, each weighted by a Rayleigh distributed random variable α and delayed in time by τ. For example, at baseband, the channel impulse response can be written as:

$$h(t) = \sum_{l=1}^{L} \alpha_l \delta(t - \tau_l) \tag{13.23}$$

In this equation, it is assumed that there are L distinct multipath components that can be resolved. Usually, the τ_l values are fixed, but there are models that allow for randomness in L and τ_l as well. Another form of characterizing the wideband channel is to use models for the *power-delay* profile of the channel. Essentially, this is a table of values corresponding to $\beta_l = E\{\alpha_l^2\}$ and τ_l in (13.23). For example, consider the power-delay profile in Table 13.1. This profile says that the earliest multipath component has an average relative power of -10 dB (usually with respect to the strongest component). The next multipath component arrives 1 μs later and has an average relative power 0 dB (it is the strongest component) and so on.

Recall that the delayed multipath components are those that cause ISI. Different measures of the *spread of the delay* are possible. The excess delay of a path is the relative delay of a path compared to the first arriving path ($\tau_l - \tau_1$). The total excess delay is the excess delay of the last arriving path ($\tau_L - \tau_1$). The *mean delay spread* τ_M is given by:

$$\tau_M = \frac{\sum_{l=1}^{L} \beta_l \tau_l}{\sum_{l=1}^{L} \beta_l} \tag{13.24}$$

The RMS delay spread is given by:

$$\tau_{rms} = \sqrt{\frac{\sum_{l=1}^{L} \beta_l \tau_l^2}{\sum_{l=1}^{L} \beta_l} - \left(\frac{\sum_{l=1}^{L} \beta_l \tau_l}{\sum_{l=1}^{L} \beta_l}\right)^2} = \sqrt{\bar{\tau^2} - \tau_M^2} \tag{13.25}$$

The impact of time dispersion is to result in irreducible error rates in the case of digital transmission (or echos and ghosts in the case of analog audio and video). The RMS delay spread provides an idea of the limitations on the achievable data rates on a given wireless channel. If the symbol duration is much larger than the RMS delay spread (about 5 times), the effects of ISI are not very significant. Otherwise, ISI may create irreducible errors in detecting the data.

Example. Compute the RMS delay spread of the channel specified in Table 13.1. Compute the maximum symbol rate that can be supported on this channel. From (13.25), we can compute the RMS delay spread as follows. We first convert the dB values to absolute values ($10^{\text{dBvalue}/10}$). The (β_l, τ_l) values are (0.1, 0), (1, 1), (0.1, 2), (1, 3), (0.01, 4). $\tau_M = \frac{(0.1\times 0+1\times 1+0.1\times 2+1\times 3+0.01\times 4)}{(0.1+1+0.1+1+0.01)} = 1.918\,\mu\text{s}$. $\bar{\tau^2} = \frac{(0.1\times 0^2+1\times 1^2+0.1\times 2^2+1\times 3^2+0.01\times 4^2)}{(0.1+1+0.1+1+0.01)} = 4.7783\,\mu\text{s}^2$. Thus $\tau_{rms} = \sqrt{4.7783 - 1.9186^2} = 1.0486\,\mu\text{s}$. The maximum symbol rate possible in this channel is $\frac{1}{(5\times 1.0486\times 10^{-6})} = 190$ kbps.

Another way of viewing time dispersion is to consider its effects in the frequency domain. If there is effectively only a single fading path between the transmitter and the receiver (or the relative delays are so small that we can assume that signals are all arriving at the same time), the impulse response of the channel can be written as $h(t) = \alpha\delta(t)$. Here α is a Rayleigh distributed random variable. The channel transfer function is $H(f) = \alpha$ which is independent of the frequency of the signal (flat on the frequency axis). We sometimes call this a *flat fading* channel for that reason. In this case, if the spectrum of the transmitted signal occupies a bandwidth of B Hz, each spectral component suffers the same attenuation.

If the channel impulse response however needs to be expressed as $h(t) = \sum_{l=1}^{L} \alpha_l \delta(t - \tau_l)$, the channel transfer function becomes $H(f) = \sum_{l=1}^{L} \alpha_l e^{-j2\pi f \tau_l}$ which is now dependent upon the frequency f. Such channels are called *frequency-selective channels*. Different components of the spectrum of a transmitted signal get attenuated by different values resulting in a distorted signal at the receiver. A measure of the frequency selectivity of a channel is the *coherence bandwidth* B_c. In simple terms, the coherence bandwidth of the channel is the average range of frequencies over which the transfer function is relatively constant. The relationship between the coherence bandwidth and the RMS multipath delay spread is simply given by $B_c = \frac{1}{5\tau_{rms}}$. If the bandwidth of the signal B is smaller than the coherence bandwidth, we can use the "flat-fading" model in the previous paragraph. Otherwise, a frequency selective channel model has to be considered for evaluating the system performance. Once again, frequency selectivity results in irreducible error rates if proper mitigating steps are not taken.

Techniques that have been used to overcome the problem of *multipath induced inter-symbol interference* (frequency selectivity) are equalization (that cancels outs the ISI created due to multipath), spread spectrum (that resolves individual multipath components and enables combining them for better performance) and orthogonal frequency division multiplexing (that uses multiple carriers with longer symbol durations). Some of these topics are briefly considered in Chapter 14.

13.4 INFRARED TRANSMISSION

Unlike RF transmissions, infrared (IR) transmissions are not very popular except for simple applications like remote control of entertainment appliances. IR is also a form of wireless transmission because the medium of transmission is still air. As technology advances, IR could prove to be a viable option for high speed wireless communications especially because of the huge available bandwidth and absence of interference. In this section we briefly consider some aspects of IR transmission. The interested reader is referred to [3] for more details.

IR Systems. In general, there are three types of IR communication systems—directed beam, diffused and hybrid. Directed beam IR systems employ highly directional transmissions and receptions. The transmitter and receiver have to be aligned properly for reliable communication. The benefit of this approach is that most of the power is used for communications making it very efficient in terms of power consumption. Diffused transmissions are more flexible. The transmissions are aimed at a wide angle at the receiver. Such transmissions can also reflect off smooth surfaces and reach the receiver. Standards such as IrDA (Infrared Data Association) and 802.11 for short-range IR transmissions have been implemented in real devices. Longer range point-to-point IR transmissions using powerful lasers have also been prototyped and to a limited extent commercialized.

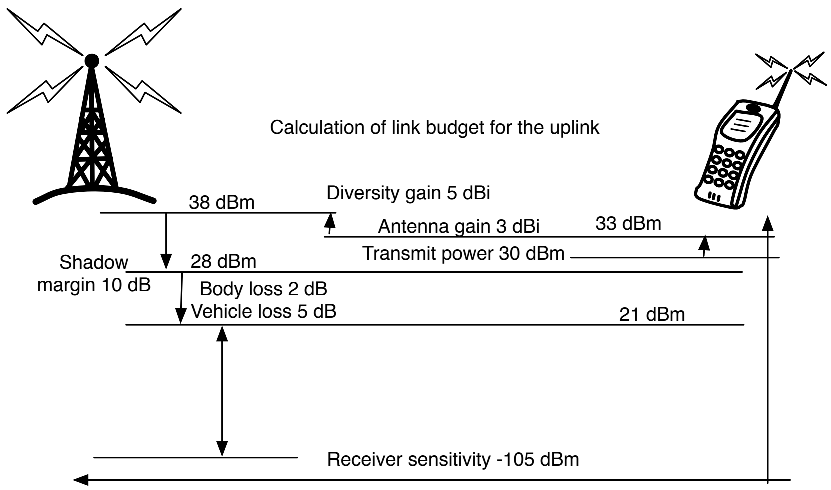
IR Transceivers. IR transmitters use LEDs and laser diodes, typically in the wavelength band of 780–950 nm. At these frequencies, plaster walls and ceiling tiles have good reflection properties enabling diffused communications. LEDs have lower modulation bandwidth and electrical to optical conversion efficiencies, but are generally considered safer than laser diodes to the human eye. Moreover, LEDs are lower in cost compared to laser diodes. Laser diodes are more efficient and can also provide very low spectral widths (better for rejecting ambient noise). IR receivers need to collect signals at a given frequency. Unlike RF receivers, the filters in IR can be either bandpass or lowpass. Lowpass filters (also called as longpass filters) allow all light beyond a cutoff wavelength. Low cost photodiodes for detecting IR signals in the 750–980 nm bands are available. The light collection area impacts the efficiency of the detector in the receiver. Concentrators like lenses can be used to increase the light collection area in the receiver.

IR Channels. IR, like visible light can penetrate transparent surfaces but not opaque objects and it can also be reflected off lighter colored and smooth surfaces. In indoor areas, transmissions are mostly confined to a single room. In outdoor areas, IR transmissions can be affected by rain, snow and other atmospheric obstructions. Ambient light (sunlight, fluorescent lighting, etc.) act as sources of noise for IR transmissions. Shot noise (see Chapter 9) is also an important noise source for IR transmissions. IR signals have the same types of degradation as RF signals such as large scale path loss and shadowing and small-scale multipath effects. Fading of amplitude is not of great concern, but delay spread effects can be important for diffused IR transmissions. In certain indoor environments with diffused IR transmissions, RMS delay spreads of up to 12 ns have been observed. Other measurements seem to indicate that the delay spread and path loss are correlated—the larger the path-loss, the more the delay spread as well.

13.5 LINK-POWER BUDGET

As in the wired (including optical fiber) transmission media discussed in the previous chapters, link power budgets are used to determine the effective range of a wireless link. A wireless link budget traces the transmit power through the communication path to the receiver. The link budget accumulates the various gains and losses along the path and the received power must exceed a required receiver sensitivity threshold level for the transmitted information to be successfully recovered. Typical factors in the link budget are the transmit power, antenna gain, antenna diversity gain, path-loss, receiver sensitivity, shadow margin, building (or vehicle) penetration loss and body loss. The link budget is often used to find the maximum allowable path-loss on a link. Note that the link-budget primarily uses the LS fading models. SS fading models appear in the link budget indirectly. SS fading impacts the bit error rate and thus the transceiver design. If the transmitter and receiver use error control coding, better modulation schemes and the like, the receiver sensitivity can be reduced. However, it will appear only as a single number in the link budget calculations.

Note there are actually two link budgets that need consideration since wireless links are typically used for two way communications with the two directions having different characteristics (see duplexing in Chapter 15). In wireless systems the link from the network to the user is termed the downlink or forward link, the other from the user to the network is called the uplink or reverse link. As an example consider the link budget for a 900 MHz cellular network shown in Figure 13.10. On the downlink from the cellular base station to the user in a moving vehicle, the transmit power is 30 dBm, an antenna with a 5-dB



Link	Uplink	Downlink
Transmit power	30 dBm	30 dBm
Antenna gain	3 dBi	5 dBi
Diversity gain	5 dB	0 dB
Shadow margin	10 dB	10 dB
Body penetration	2 dB	2 dB
Vehicle penetration	5 dB	5 dB
Receiver sensitivity	-105 dBm	-90 dBm
Path loss budget	126 dB	108 dB

Figure 13.10 Illustration of link power budget.

gain is used, a 10-dB shadow margin is assumed appropriate for the environment, 2 dB is allowed for attentuation due the user handset againist the body of the user, similarly 5 dB is allowed for attenuation by the vehicle in which the user is located, the receiver sensititivity threshold is −90 dBm. Adding the gains and subtracting the losses results a maximum allow path loss budget of 108 dB. Performing a similar calculation for the uplink we get a maximum allowable path-loss of 126 dB (see the top part of Figure 13.10). Note that the difference is due to the use of a more sensitive receiver at the base station and addition of antenna diversity gain of 5 dB at the base station.

CONCLUSION

The reader has now seen the three major channels for telecommunications, copper (wire), optical fiber, and wireless. Although fundamentally each of the channels is a carrier of