where $G_p = W_{ss}/R$ represents the *processing gain*, and $(J/S)_{reqd}$ can be written

$$\left(\frac{J}{S}\right)_{reqd} = \frac{G_p}{(E_b/J_0)_{reqd}} \tag{12.39}$$

The ratio $(J/S)_{reqd}$ is a figure of merit that provides a measure of how *invulnerable* a system is to interference. Which system has better jammer-rejection capability: one with a larger $(J/S)_{reqd}$ or a smaller $(J/S)_{reqd}$? The *larger* the $(J/S)_{reqd}$, the *greater* is the system's noise rejection capability, since this figure of merit describes how much noise power relative to signal power is *required* in order to degrade the system's specified error performance. Of course, the communicator would like the communication system *not* to degrade at all.

Another way of describing the relationship in Equation (12.39) is as follows. An adversary would like to employ a jamming strategy that forces the effective $(E_b/J_0)_{reqd}$ to be as large as possible. The adversary may employ pulse, tone, or partial-band jamming rather than wideband noise jamming. A large $(E_b/J_0)_{reqd}$ implies a small $(J/S)_{reqd}$ ratio for a fixed processing again. This may force the communicator to employ a larger processing gain to increase the $(J/S)_{reqd}$. The system designer strives to choose a signaling waveform such that the jammer can gain no special advantage by using a jamming strategy other than wideband Gaussian noise.

12.6.1.4 Anti-Jam Margin

Sometimes the $(J/S)_{reqd}$ ratio is referred to as the *anti-jam* (AJ) *margin,* since it characterizes the system jammer-rejection capability. But this is not really a good use of the phrase since AJ margin usually means the safety margin against a *particular threat.* Using the same approach as in Chapter 5 (for calculating the margin against thermal noise), we can define the AJ margin as

$$M_{AJ}(\text{dB}) = \left(\frac{E_b}{J_0}\right)_r (\text{dB}) - \left(\frac{E_b}{J_0}\right)_{reqd} (\text{dB}) \tag{12.40}$$

where $(E_b/J_0)_r$ is the E_b/J_0 *actually received.* Following the same format as Equation (12.38), we can express $(E_b/J_0)_r$ as

$$\left(\frac{E_b}{J_0}\right)_r = \frac{G_p}{(J/S)_r} \tag{12.41}$$

where $(J/S)_r$, or simply J/S, is the ratio of the actually received jammer power to signal power. Later, we develop an expression for received E_b/I_0, similar to Equation (12.41), where I_0 is the interference power spectral density due to other users in a CDMA cellular system. The concept of computing such a bit-energy to interference ratio is the same, whether the interference stems from a jammer, an accidental interferer, or other users who are authorized to share the same spectral region.

We now combine Equation (12.41) with Equations (12.38) and (12.40), as follows:

$$M_{AJ}(\text{dB}) = \frac{G_p}{(J/S)_r}(\text{dB}) - \frac{G_p}{(J/S)_{reqd}}(\text{dB}) \tag{12.42}$$

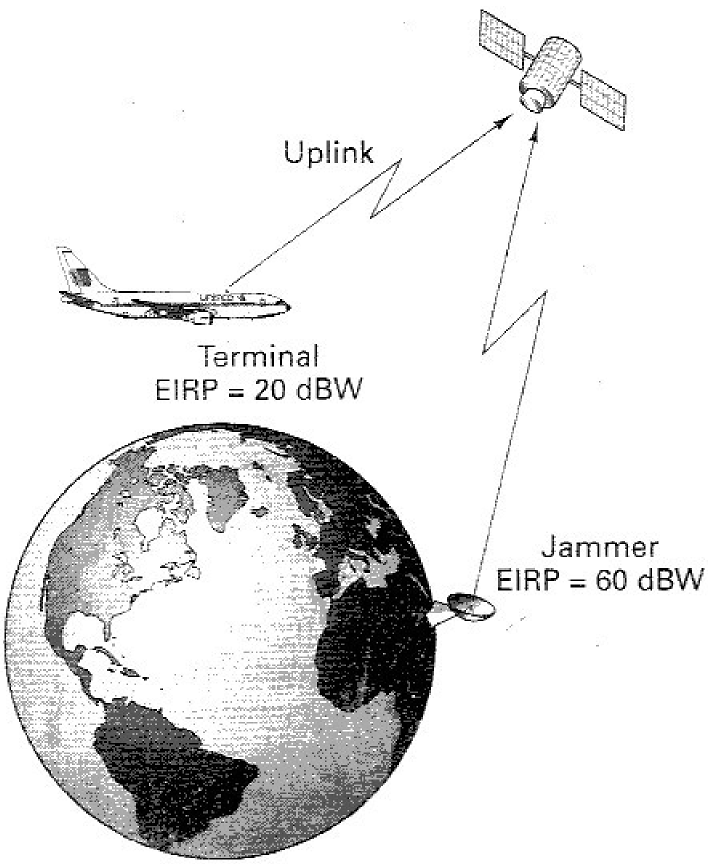


Figure 12.26 Satellite jamming scenario.

$$= \left(\frac{J}{S}\right)_{\text{reqd}} (\text{dB}) - \left(\frac{J}{S}\right)_r (\text{dB}) \tag{12.43}$$

Example 12.2 Satellite Jamming

Figure 12.26 illustrates a satellite jamming scenario. The airplane terminal is equipped with a frequency hopping (FH) spread-spectrum system transmitting with an EIRP_T = 20 dBW. The data rate is R = 100 bits/s. The jammer is transmitting wideband Gaussian noise, continually, with an EIRP_J = 60 dBW. Assume that $(E_b/J_0)_{\text{reqd}}$ = 10 dB and that the path loss is identical for both the airplane terminal and the jammer.

(a) Should the communicators be concerned more with the jamming of the uplink or with that of the downlink?

(b) If it is desired to have an AJ margin of 20 dB, what should be the value of the hopping bandwidth W_{ss}?

Solution

(a) Jamming the uplink is of much greater concern, since such single-point interference could degrade the communications of a multitude of terminals that are simultaneously using the satellite transponder. To achieve an equivalent degradation by jamming the downlink, the jammer would have to jam each of the receiving terminals. Downlink jamming is of some concern for critical military missions, but of less concern than uplink jamming.

(b) With the assumption that the path loss is the same for both the communicator and the jammer, we can replace $(J/S)_r$ in Equation (12.43) with the ratio of *transmitted* jammer-to-signal power, $\text{EIRP}_J/\text{EIRP}_T$. Therefore, we can write

$$M_{AJ}\,(\text{dB}) = (J/S)_{\text{reqd}}\,(\text{dB}) + \text{EIRP}_T\,(\text{dBW}) - \text{EIRP}_J\,(\text{dBW})$$

$$= G_p\,(\text{dB}) - \left(\frac{E_b}{J_0}\right)_{\text{reqd}}(\text{dB}) + \text{EIRP}_T\,(\text{dBW}) - \text{EIRP}_J\,(\text{dBW})$$

$$G_p = 20\text{ dB} + 10\text{ dB} - 20\text{ dBW} + 60\text{ dBW} = 70\text{ dB}$$

$$W_{ss} = G_p\,(\text{dB}) + R\,(\text{dB-Hz}) = 70\text{ dB} + 20\text{ dB-Hz}$$

$$= 90\text{ dB-Hz} = 1\text{ GHz}$$

Example 12.3 Satellite Downlink Jamming

In Example 12.2 the distance from the transmitting airplane to the receiving satellite and the distance from the jammer to the satellite were assumed identical. Certainly, the closer the jammer gets to the receiver, the greater will be the jamming interference. Consider a downlink jamming scenario where the satellite $\text{EIRP}_s = 35$ dBW, the jammer $\text{EIRP}_J = 60$ dBW, the space loss from the satellite to the receiving terminal is $L_s = 200$ dB, and the space loss from the jammer to the receiving terminal is $L_s' = 160$ dB. How much processing gain is needed to close the link with an AJ margin of 0 dB? Assume that $(E_b/J_0)_{\text{reqd}} = 10$ dB.

Solution

For the downlink jamming scenario the proximity of the jammer to the receiving airplane is much closer than that of the satellite to the airplane. These distances show up as the space losses in the $(J/S)_r$ term of Equation (12.43), as follows:

$$M_{AJ}\,(\text{dB}) = \left(\frac{J}{S}\right)_{\text{reqd}}(\text{dB}) - \left(\frac{J}{S}\right)_r(\text{dB})$$

where

$$\left(\frac{J}{S}\right)_r(\text{dB}) = \text{EIRP}_J\,(\text{dBW}) - L_s'\,(\text{dB}) - \text{EIRP}_s\,(\text{dBW}) + L_s\,(\text{dB})$$

and

$$\left(\frac{J}{S}\right)_{\text{reqd}}(\text{dB}) = \frac{W_{ss}}{R}(\text{dB}) - \left(\frac{E_b}{J_0}\right)_{\text{reqd}}(\text{dB})$$

Combining the above equations, and solving for processing gain, $G_p = W_{ss}/R$, yields

$$G_p\,(\text{dB}) = 75\text{ dB}$$

12.6.2 Broadband Noise Jamming

If the jamming signal is modeled as a zero-mean wide-sense stationary Gaussian noise process with a flat power spectral density over the frequency range of interest, then for a fixed jammer received power, J, the jammer power spectral density J'_0 is equal to J/W, where W is the bandwidth that the jammer chooses to occupy. If the jammer strategy is to jam the entire spread-spectrum bandwidth, W_{ss}, with its fixed power, the jammer is referred to as a wideband or *broadband jammer,* and the jammer power spectral density is

$$J_0 = \frac{J}{W_{ss}} \tag{12.44}$$

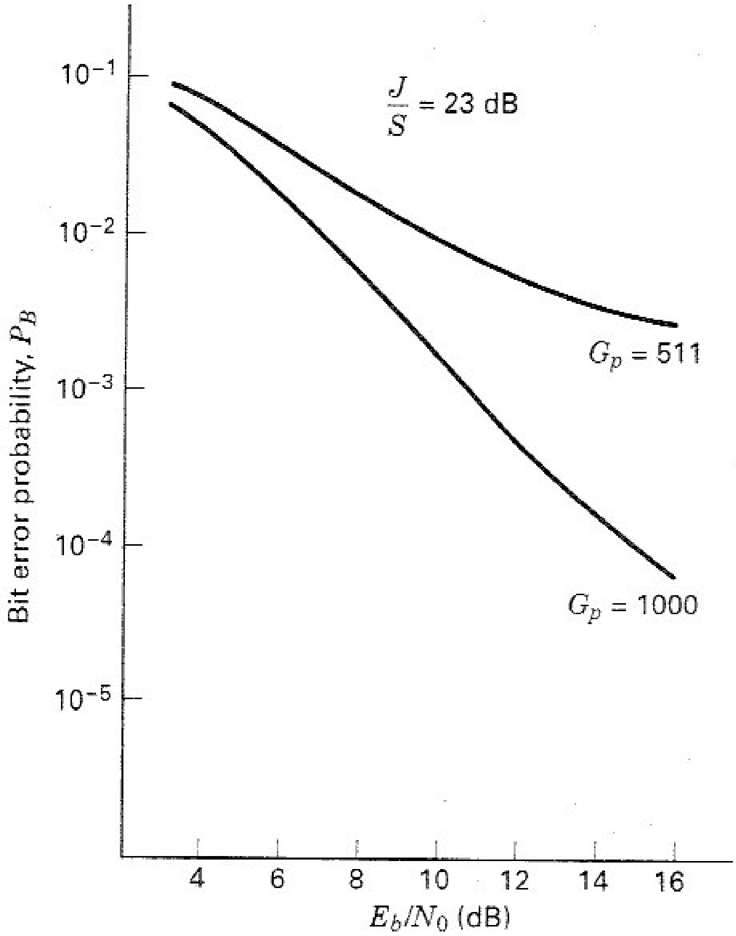
In Chapter 4 it was shown that the bit error probability P_B for a coherently demodulated BPSK system (without channel coding) is

$$P_B = Q\left(\sqrt{\frac{2E_b}{N_0}}\right) \tag{12.45}$$

where $Q(x)$ is defined in Equations (3.43) and (3.44) and tabulated in Table B.1. The single-sided noise power spectral density N_0 represents thermal noise at the front end of the receiver. The presence of the jammer increases this noise power spectral density from N_0 to $(N_0 + J_0)$. Thus the average bit error probability for a coherent BPSK system in the presence of broadband jamming is

$$P_B = Q\left(\sqrt{\frac{2E_b}{N_0 + J_0}}\right) = Q\left[\sqrt{\frac{2E_b/N_0}{1 + (E_b/N_0)(J/S)/G_p}}\right] \tag{12.46}$$

When P_B is plotted versus E_b/N_0 for a given J/S ratio, the resulting curves are such as those in Figure 12.27 [6, 21]. The curves in Figure 12.27, shown for two different



12.27 Bit-error probability versus E_b/N_0 for a given J/S ratio. (Reprinted with permission from R. L. Pickholtz, D. L. Schilling, and L. B. Milstein, "Theory of Spread-Spectrum Communications—A Tutorial," *IEEE Trans. Commun.*, vol. COM30, no. 5, May 1982, Fig. 11, p. 866 © 1982 IEEE.)

values of processing gain, *tend to flatten out* as E_b/N_0 increases, indicating that for a given ratio of jammer power to signal power, the jammer will cause some irreducible error probability. The only way to reduce this error probability is to increase the processing again.

12.6.3 Partial-Band Noise Jamming

A jammer can often increase the degradation to a FH system by employing *partial-band* jamming. Assuming that the frequency hopped modulation format is noncoherently detected binary FSK, the probability of a bit error, from Equation (4.96), is

$$P_B = \frac{1}{2}\exp\left(-\frac{E_b}{2N_0}\right) \tag{12.47}$$

Let us define a parameter, ρ, where $0 < \rho \leq 1$, representing the fraction of the band being jammed. The jammer can trade bandwidth jammed for in-band jammer power, such that by jamming a band $W = \rho W_{ss}$, the jammer noise power spectral density can be concentrated to a level J_0/ρ, thus maintaining a constant average jamming received power J where $J = J_0 W_{ss}$.

In the case of partial-band jamming, a specific transmitted symbol will be received unjammed, with probability $(1 - \rho)$, and will be perturbed by jammer power with spectral density J_0/ρ, with probability ρ. Therefore, the average bit error probability can be written from Equation (12.47), as follows:

$$P_B = \frac{1-\rho}{2}\exp\left(-\frac{E_b}{2N_0}\right) + \frac{\rho}{2}\exp\left[-\frac{E_b}{2(N_0 + J_0/\rho)}\right] \tag{12.48}$$

Since, in a jamming environment, it is often the case that $J_0 >> N_0$, we can simplify Equation (12.48) to the form

$$P_B \approx \frac{\rho}{2}\exp\left(-\frac{\rho E_b}{2J_0}\right) \tag{12.49}$$

Figure 12.28 illustrates the probability of bit error versus E_b/J_0 for various values of the fraction, ρ. Clearly, the jammer would choose the fraction $\rho = \rho_0$ that maximizes P_B. Notice that ρ_0 decreases with increasing values of E_b/J_0 (see the ρ_0 locus in Figure 12.28). An expression for ρ_0 is easily found by differentiation (setting $dP_B/d\rho = 0$ and solving for ρ). This yields

$$\rho_0 = \begin{cases} \dfrac{2}{E_b/J_0} & \text{for } \dfrac{E_b}{J_0} > 2 \\ 1 & \text{for } \dfrac{E_b}{J_0} \leq 2 \end{cases} \tag{12.50}$$

In this case, $(P_B)_{max}$ is given by

$$(P_B)_{max} = \begin{cases} \dfrac{e^{-1}}{E_b/J_0} & \text{for } \dfrac{E_b}{J_0} > 2 \\ \dfrac{1}{2}\exp\left(-\dfrac{E_b}{2J_0}\right) & \text{for } \dfrac{E_b}{J_0} \leq 2 \end{cases} \tag{12.51}$$

where e is the base of the natural logarithm ($e = 2.7183$). This result is dramatic; the effect of a worst-case partial-band jammer on a system with spread spectrum *but without coding* changes the exponential relationship of Equation (12.49) into the inverse linear one of Equation (12.51). The ρ_0 locus in Figure 12.28 illustrates the

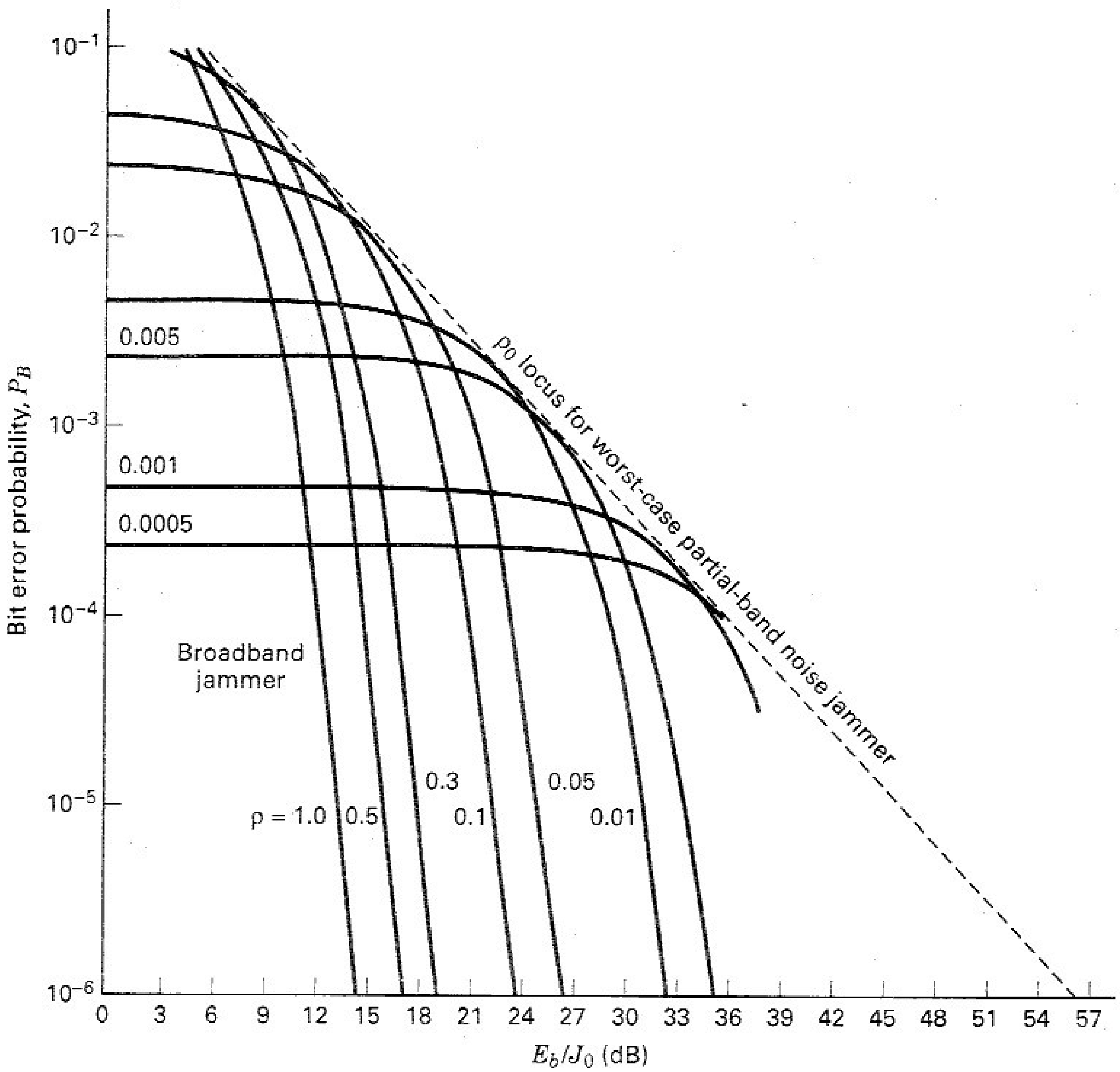


Figure 12.28 Partial-band noise jammer (FH/BFSK signaling). (Reprinted from M. K. Simon, J. K. Omura, R. A. Scholtz, and B. K. Levitt, *Spread Spectrum Communications,* Vol. 1, Fig. 3.24, p. 173. © 1985, with permission of the publisher, Computer Science Press, Inc., 1803 Research Blvd., Rockville, MD., 20850, USA.)

P_B versus E_b/J_0 performance for the worst-case partial-band jammer. Here at 10^{-6} bit-error probability there is over 40-dB difference between broadband noise jamming and the worst-case partial-band jamming for the same jamming power [4, 22]. Hence, an intelligent jammer, with fixed finite power, can produce significantly greater degradation with partial-band jamming than is possible with broadband jamming. Forward error correction (FEC) coding with appropriate interleaving can mitigate this degradation [9]. In fact, for codes with low enough rates, FEC can *force* a partial-band jammer to be a worst-case jammer only when operating as a broadband jammer [23, 24].

12.6.4 Multiple-Tone Jamming

In the case of *multiple-tone jamming,* the jammer divides its total received power, J, into distinct, equal-power, random-phase CW tones. These are distributed over the spread-spectrum bandwidth, W_{ss}, according to some strategy [9]. The analysis of the effects of tone jamming is more complicated than that of noise jamming, especially for DS systems. Therefore, the effect of a despread tone is often approximated as Gaussian noise. Reference [25] provides analysis of the performance of DS systems in the presence of multiple-tone interference. For a noncoherent FH/FSK system operating in the presence of partial-band tone jamming, the performance is often assumed the same as that of partial-band noise jamming [26]. However, multiple-CW-tone jamming can be more effective than partial-band noise against FH/MFSK signals because CW tones are the most efficient way for a jammer to inject energy into noncoherent detectors [8]. References, [8, 9, 26, 27] provide extensive treatment and analysis of the performance of various communications systems in the presence of various types of jammers.

In the FFH/MFSK demodulator of Figure 12.16, a chip-clipping circuit is shown between each envelope detector and accumulator. The function of such a circuit in a tone-jamming environment can best be understood with the aid of the example shown in Figure 12.29. An 8-ary FSK frequency-hopping system with no diversity, indicated in Figure 12.29a, is compared with a *fast* frequency-hopping system that combines chip repeating ($N = 4$ in this example) with the clipping of each chip, indicated in Figure 12.29b. Each row in the figures represents one of the $M = 8$ accumulators shown in Figure 12.16. The presence of a signal in the accumulator is indicated by a vector. In Figure 12.29a we see that, for a particular frequency hop, the data band is occupied by a received message symbol with received signal power S. If, by chance, a jamming tone with received power J, where $J \geq S$, falls on a different tone within this data band during the same hop, the detector would not be able to decide reliably on the correct symbol.

In Figure 12.29b, the communicator's four chips (the length of each vector is a measure of the clipped signal power, S') sum to the maximum capacity of the accumulator. If the jammer tones, by chance, fall in the same spectral region as that of the signal, they will not confuse the detector, since the jamming tones are also clipped to the same level, $J' = S'$, as the signal chips. In Figure 12.29b, two of the jamming tones fall in the data band, but because they are clipped, there is no confusion about the correct symbol decision.

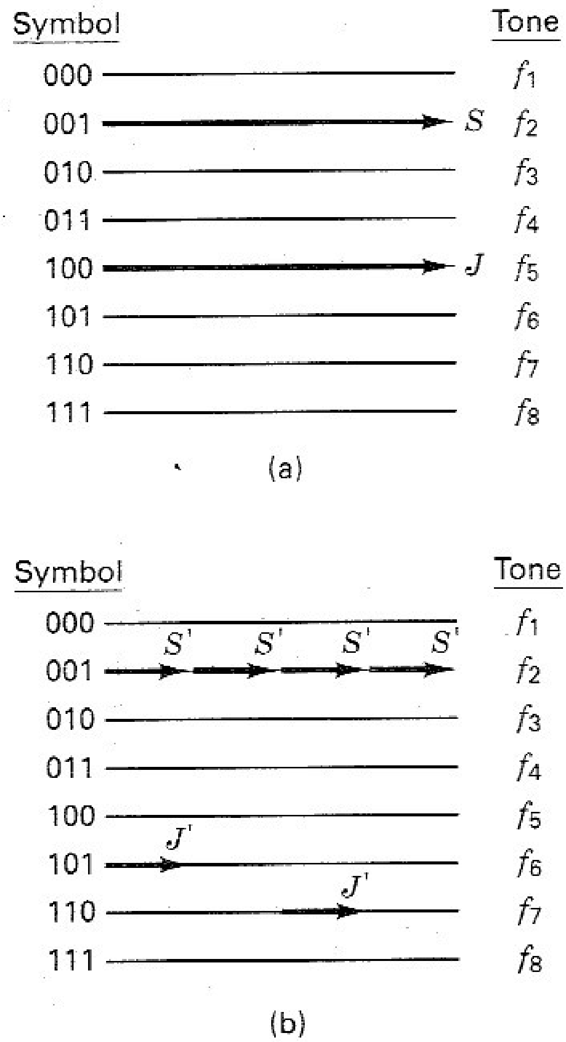


Figure 12.29 Fast-hopping symbol repeat versus tone jamming. (a) One frequency hop. (b) Four frequency hops.

12.6.5 Pulse Jamming

Consider a spread-spectrum DS/BPSK communication system in the presence of a pulse-noise jammer. A pulse-noise jammer transmits pulses of bandlimited white Gaussian noise having a time-averaged received power J, although the actual power during a jamming pulse duration is larger. Assume that the jammer can choose the center frequency and bandwidth of the noise to be the same as the receiver's center frequency and bandwidth. Assume also that the jammer can trade duty cycle for increased (concentrated) jammer power, such that if the jamming is present for a fraction $0 < \rho < 1$ of the time, then during this time, the jammer power spectral density is increased to a level J_0/ρ, thus maintaining a constant time-averaged power J (where $J = J_0 W_{ss}$ and W_{ss} is the system spread-spectrum bandwidth).

The bit error probability P_B for a coherently demodulated BPSK system (without channel coding) was given in Equation (12.45):

$$P_B = Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$$

The single-sided noise power spectral density N_0 represents thermal noise at the front end of the receiver. The presence of the jammer increases this noise power spectral density from N_0 to $(N_0 + J_0/\rho)$. Since the jammer transmits with duty cycle ρ, the average bit-error probability is

$$P_B = (1 - \rho)Q\left(\sqrt{\frac{2E_b}{N_0}}\right) + \rho Q\left(\sqrt{\frac{2E_b}{N_0 + J_0/\rho}}\right) \tag{12.52}$$

We can generally assume that in a jamming environment, N_0 can be neglected. Therefore, we can write

$$P_B \approx \rho Q\left(\sqrt{\frac{2E_b \rho}{J_0}}\right) \tag{12.53}$$

The jammer will, of course, attempt to choose the duty cycle ρ that maximizes P_B. Figure 12.30 illustrates P_B for various values of ρ. The value of $\rho = \rho_0$ that maximizes P_B decreases with increasing values of E_b/J_0, as was the case with partial-band jamming. This is seen by differentiating Equation (12.53) to obtain [4]

$$\rho_0 = \begin{cases} \dfrac{0.709}{E_b/J_0} & \text{for } \dfrac{E_b}{J_0} > 0.709 \\ 1 & \text{for } \dfrac{E_b}{J_0} \le 0.709 \end{cases} \tag{12.54}$$

which results in the maximum bit error probability

$$(P_B)_{\max} = \begin{cases} \dfrac{0.083}{E_b/J_0} & \text{for } \dfrac{E_b}{J_0} > 0.709 \\ Q\left(\sqrt{\dfrac{2E_b}{J_0}}\right) & \text{for } \dfrac{E_b}{J_0} \le 0.709 \end{cases} \tag{12.55}$$

The effect of a worst-case pulse jammer upon a system with spread spectrum *but without coding* changes the complementary error function relationship of Equation (12.53) into the inverse linear one of Equation (12.55). As a result, at an error probability of 10^{-6}, there is about a 40-dB difference in E_b/J_0 between the broadband jammer and the worst-case pulse jammer (see Figure 12.30). For the same jammer power, the jammer can do considerably more harm to an uncoded DS/BPSK system with pulse jamming than with constant power jamming. The effect of a pulse-noise jammer on uncoded DS/BPSK is similar to the effect of a partial-band noise jammer on uncoded FH/BFSK, treated in Section 12.6.3. In both cases considerable degradation is brought about by concentrating more jammer power on a fraction of the transmitted uncoded symbols. Forward error correction coding with appropriate interleaving can almost fully restore this degraded performance [8, 23–25, 28].

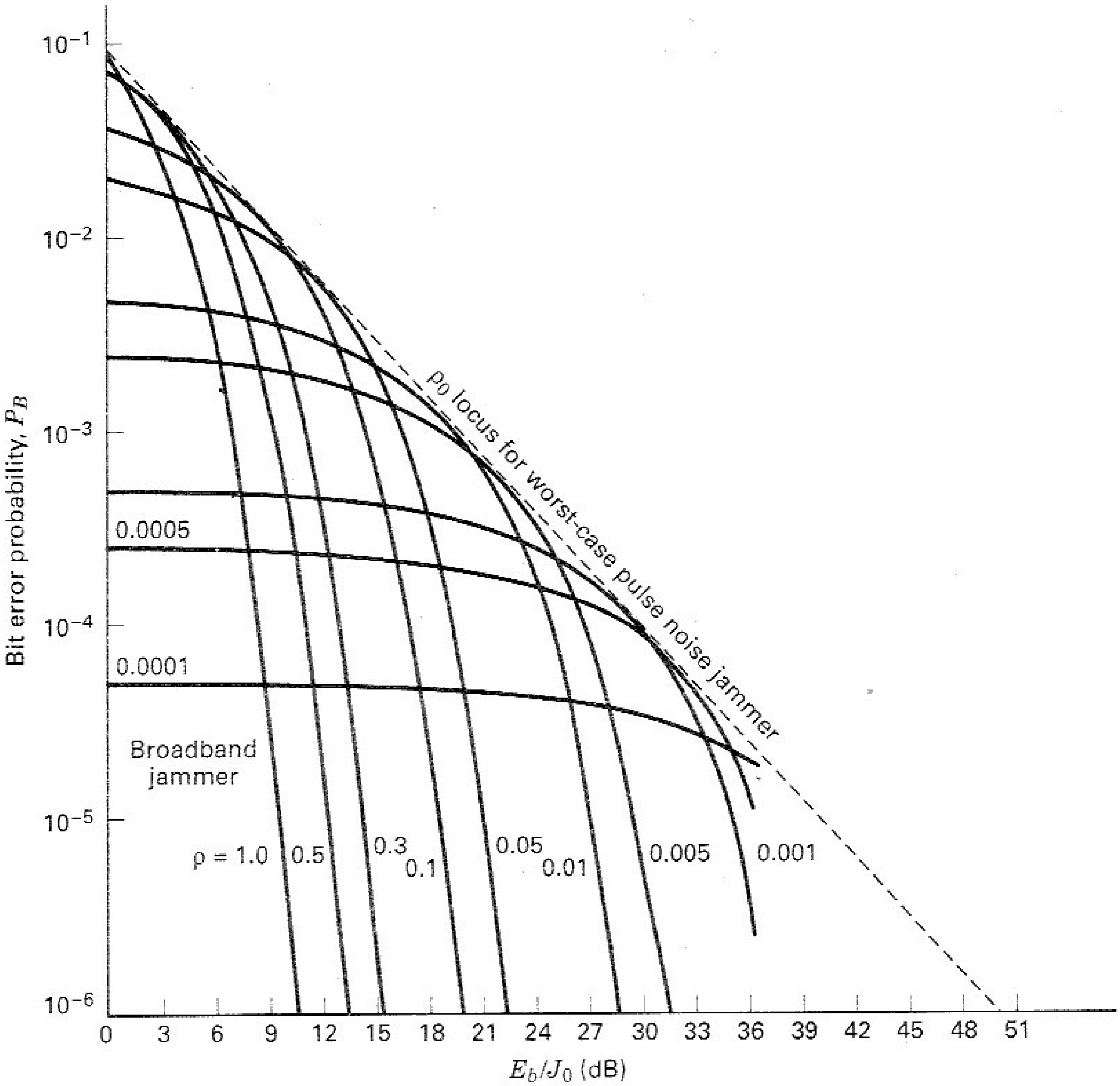


Figure 12.30 Pulse noise jammer (DS/BPSK signaling). (Reprinted from M. K. Simon, J. K. Omura, R. A. Scholtz, and B. K. Levitt, *Spread Spectrum Communications,* Vol. 1, Fig. 3.7, p. 150 © 1985, with permission of the publisher, Computer Science Press, Inc., 1803 Research Blvd., Rockville, Md. 20850 USA.)

12.6.6 Repeat-Back Jamming

In Examples 12.2 and 12.3 we considered an FH spread-spectrum system performance against a broadband Gaussian noise jammer. Notice that the frequency hopping rate did not enter into the margin computations. Isn't this disturbing? Intuitively, it would seem that the faster the frequency hops, the easier it is to "hide" the signal from the jammer. If the hopping rate truly does not enter into the computations, why not hop only once a day or once a week? The answer is that the measure of jammer-rejection capability, namely processing gain, G_p, is based on the assumption that the jammer is a "dumb" jammer; that is, the jammer knows the extent of the spread-spectrum bandwidth, W_{ss}, but does *not* know the exact spectral location of the signal at any moment in time. We assume that the hopping rate is *fast enough* to preclude the jammer from monitoring the transmitted signal so as to usefully change this jamming strategy. Under what condition is this assumption questionable? There are "smart" jammers that are known as *repeat-back jammers* or *frequency-follower (FF) jammers.* These jammers monitor a communicator's signal (usually via a sidelobe beam from the transmitting antenna). They possess wideband receivers and high-speed signal processing capability that enable them to rapidly concentrate their jamming signal power in the spectral vicinity of a communicator's FH/FSK signal. By so doing, the smart jammer can increase the jamming power in the communicator's instantaneous bandwidth, thereby gaining an advantage over a wideband jammer. Notice that this strategy is useful only against frequency-hopping spread-spectrum signals. In direct-sequence systems, there is no instantaneous narrowband signal for the jammer to detect.

What can be done to defeat the repeat-back jammer? One method is to simply hop so fast that by the time the jammer receives, detects, and transmits the jamming signal, the communicator is already transmitting at a *new* hop (which of course will be unaffected by jamming at the frequency of the prior hop). The following example should make this point clear.

Example 12.4 Fast Hopping to Evade the Repeat-Back Jammer

Assume that a repeat-back jammer is located $d = 30$ km away from the communicator. Assume further that the jammer can monitor any uplink transmission from the com-

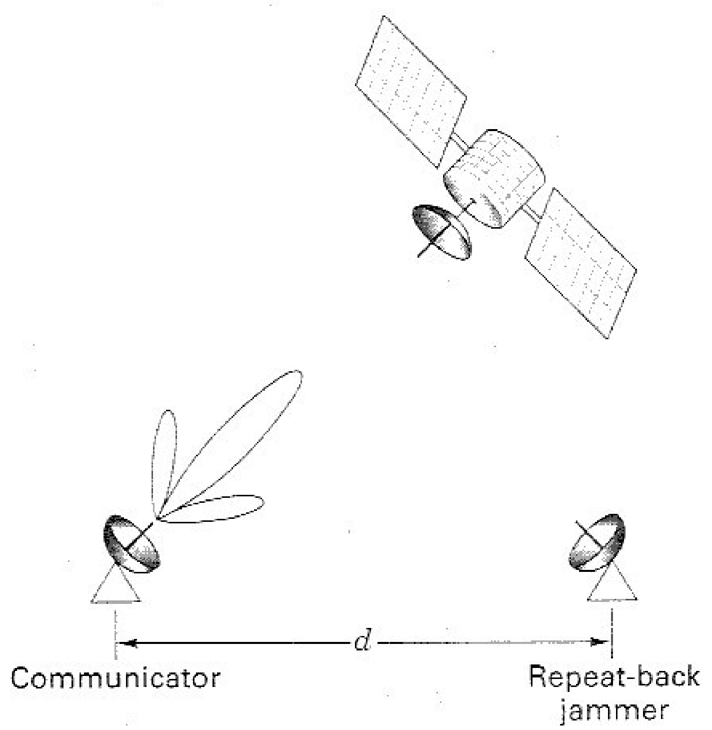


Figure 12.31 Example of fast hopping to evade the repeat-back jammer.

municator to a nearby satellite, as shown in Figure 12.31. How fast must the communicator hop his frequency to evade the repeat-back jammer? Assume that the jammer can change its jamming frequency in zero time, and that the only differential delay between the communicator's uplink signal and the jamming uplink signal is the propagation delay from the communicator to the jammer.

Solution

To ensure that the communicator's tone transmission and the jammer's attempt to disrupt that tone do not overlap in time, it is necessary that the duration of each hop have the value

$$T_{\text{hop}} \leq \frac{d}{c} = \frac{3 \times 10^4 \text{m}}{3 \times 10^8 \text{ m/s}} = 10^{-4} \text{ s}$$

where c is the speed of light. Then $R_{\text{hop}} \geq 10{,}000$ hops/s.

12.6.7 BLADES System

Another technique capable of defeating the repeat-back jammer dates back to the mid-1950s, when Sylvania engineers developed a system named the *Buffalo Laboratories Application of Digitally Exact Spectra,* or BLADES. The system used its code generator to independently select two new frequencies for each bit; the *final choice* of the frequency tone actually transmitted was dictated by the data bit about to be transmitted. Figure 12.32 illustrates a typical data stream of binary ones and zeros, called *marks* and *spaces,* respectively, and a sequence of frequency pairs f_1 and f'_1, f_2 and $f'_2, \ldots$. The appearance of a mark dictates the choice of frequency f_i, while the appearance of a space dictates the choice of frequency f'_i. As shown in the figure, the data stream in this example gives rise to the sequence of transmitted tones, $f'_1, f_2, f'_3, f'_4, f_5$, and so on. How can such a system defeat a repeat-back jammer? The jammer monitors the transmissions and sends up energy in the neighborhood of the frequencies it perceives. The modulation of the BLADES system has no structure in the usual sense; either there *is* energy present or there is *no* energy

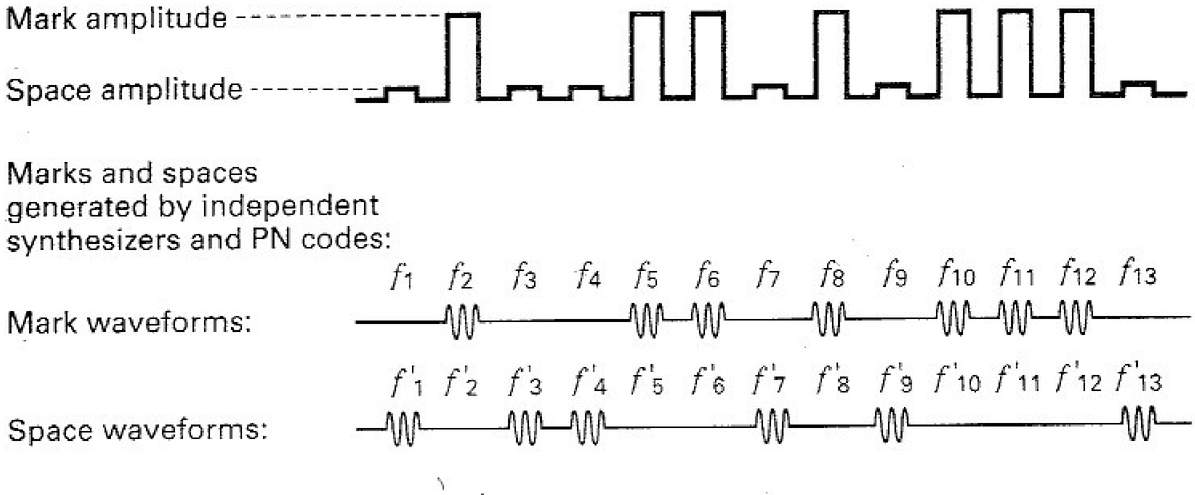


Figure 12.32 BLADES system.

present at a given frequency. The jammer sending narrowband energy in the same spectral neighborhood as the signal, is not destroying any modulation structure. For a noncoherent system, the jammer is only enhancing the communicator's signal. The only recourse for the repeat-back jammer is to change strategy by becoming a broadband jammer, and to jam the entire spread-spectrum bandwidth.

Notice that it is not really necessary to have a *pair* of frequencies for each bit. A *single* frequency will do. The communicator then transmits the pseudo-random frequency for a binary one and sends nothing for a binary zero. The receiver has the same code generator and therefore monitors the same pseudo-random frequencies. A binary one is detected by virtue of energy at the monitored frequency, and a binary zero is known by a lack of energy at the monitored frequency. Of course, the system is not as robust as when the marks and spaces are each transmitted on independently selected frequencies.

12.7 COMMERCIAL APPLICATIONS

12.7.1 Code-Division Multiple Access

Spread-spectrum multiple access techniques allow multiple signals occupying the same RF bandwidth to be transmitted simultaneously without interfering with one another. The application of spread-spectrum techniques to the problem of multiple access was discussed in Chapter 11 for a frequency hopped code-division multiple access (FH/CDMA) scheme. Here we consider CDMA using direct sequence (DS/CDMA). In these schemes, each of N user groups is given its own code, $g_i(t)$, where $i = 1, 2, \ldots, N$. The user codes are approximately orthogonal, so that the cross-correlation of two different codes is near zero. The main advantage of a CDMA system is that all the participants can share the full spectrum of the resource asynchronously; that is, the transition times of the different users' symbols do not have to coincide.

A typical DS/CDMA block diagram is shown in Figure 12.33. The first block illustrates the data modulation of a carrier, $A \cos \omega_0 t$. The output of the data modulator belonging to a user from group 1 is

$$s_1(t) = A_1(t) \cos [\omega_0 t + \phi_1(t)] \tag{12.56}$$

The waveform is very general in form; no restriction has been placed on the type of modulation that can be used.

Next, the data-modulated signal is multiplied by the spreading signal $g_1(t)$ belonging to user group 1, and the resulting signal $g_1(t)s_1(t)$ is transmitted over the channel. Simultaneously, users from group 2 through N multiply their signals by their own code functions. Frequently, each code function is kept secret, and its use is restricted to the community of authorized users. The signal present at the receiver is the linear combination of the emanations from each of the users. Neglecting signal delays, we show this linear combination as

$$g_1(t)s_1(t) + g_2(t)s_2(t) + \cdots + g_N(t)s_N(t) \tag{12.57}$$

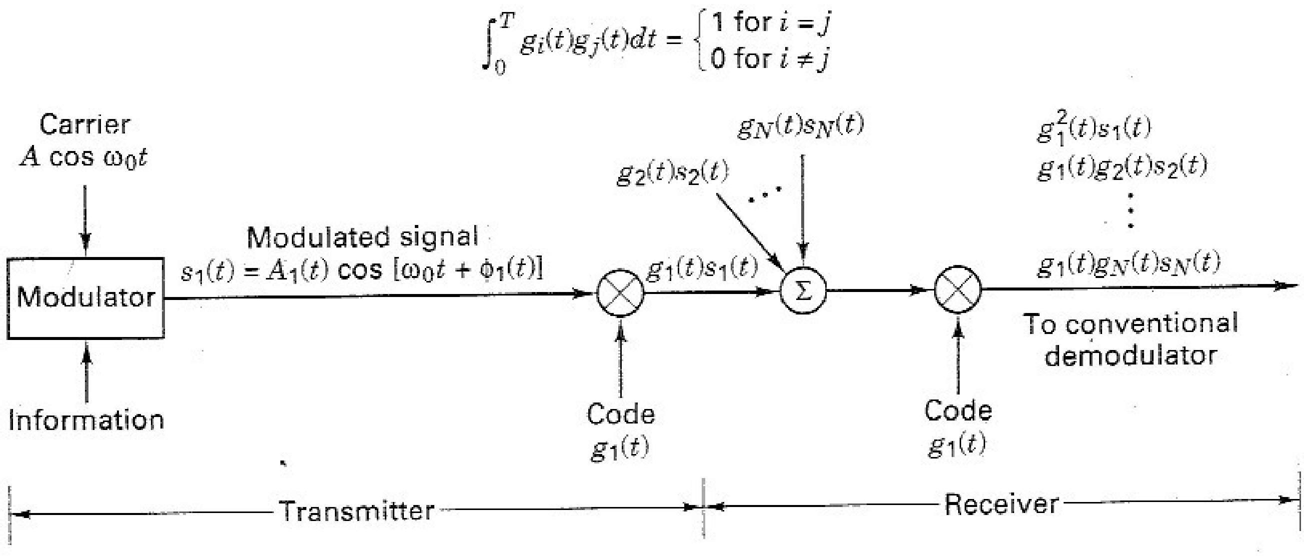


Figure 12.33 Code-division multiple access.

As mentioned earlier, multiplication of $s_1(t)$ by $g_1(t)$ produces a signal whose spectrum is the convolution of the spectrum of $s_1(t)$ with the spectrum of $g_1(t)$. Thus, assuming that the signal $s_1(t)$ is relatively narrowband compared with the code or spreading signal $g_1(t)$, the product signal $g_1(t)s_1(t)$ will have approximately the bandwidth of $g_1(t)$. Assume that the receiver is configured to receive messages from user group 1. Assume, too, that the $g_1(t)$ code, generated at the receiver, is perfectly synchronized with the received signal from a group 1 user. The first stage of the receiver multiplies the incoming signal of Equation (12.57) by $g_1(t)$. The output of the multiplier will yield the desired signal,

$$g_1^2(t)s_1(t)$$

plus a composite of undesired signals,

$$g_1(t)g_2(t)s_2(t) + g_1(t)g_3(t)s_3(t) + \cdots + g_1(t)g_N(t)s_N(t) \tag{12.58}$$

If the code functions $\{g_i(t)\}$ are chosen with orthogonal properties, similar to Equation (12.14), the desired signal can be extracted perfectly in the absence of noise since $\int_0^T g_i^2(t) = 1$, and the undesired signals are easily rejected, since $\int_0^T g_i(t)g_j(t)\, dt = 0$ for $i \neq j$. In practice, the codes are not perfectly orthogonal; hence, the cross-correlation between user codes introduces performance degradation, which limits the maximum number of simultaneous users.

Consider the frequency-domain view of the DS/CDMA receiver. Figure 12.34a illustrates the wideband input to the receiver; it consists of wanted and unwanted signals, each spread by its own code with code rate R_{ch}, and each having a power spectral density of the form $\text{sinc}^2(f/R_{ch})$. Receiver thermal noise is also shown as having a flat spectrum across the band. The combined waveform of Equa-

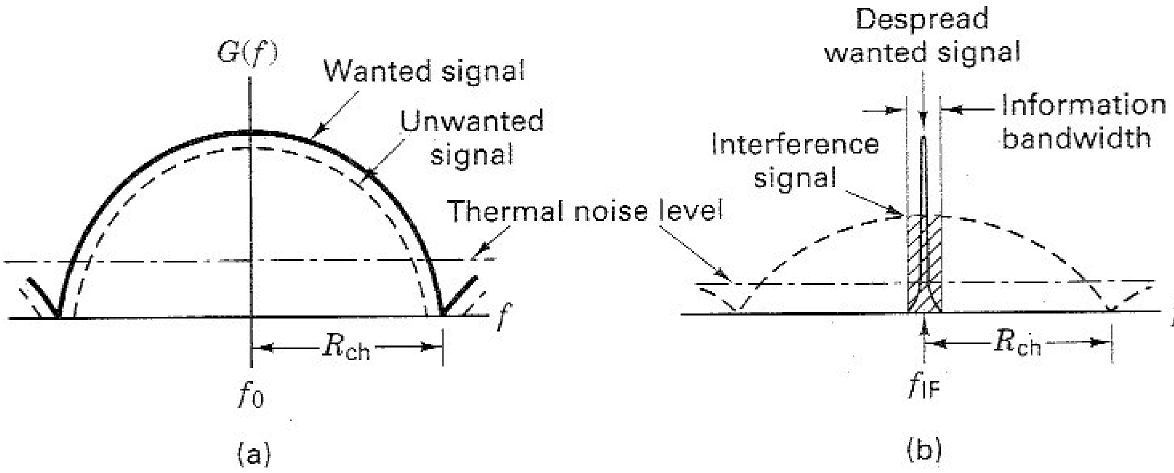


Figure 12.34 Spread-spectrum signal detection. (a) Spectrum at the input to receiver. (b) Spectrum after correlation with the correct and synchronized PN code.

tion (12.58) (desired plus undesired signals) is applied to the input of the receiver correlator driven by a synchronous replica of $g_1(t)$. Figure 12.34b illustrates the spectrum after correlation with the code $g_1(t)$ (despreading). The desired signal, occupying the information bandwidth centered at an intermediate frequency (IF), is then applied to a conventional demodulator, with bandwidth just wide enough to accommodate the despread signal. The undesired signals of Equation (12.58) remain effectively spread by $g_1(t)g_i(t)$. Only that portion of the spectrum of the unwanted signals falling in the information bandwidth of the receiver will cause interference with the desired signal.

Pursley [17] presents an excellent treatment on the performance of SSMA using DS, taking correlation properties of the code sequences into account. Also, Geraniotis [18] and Geraniotis and Pursley [19, 20] evaluate the performance of FH and DS multiple access systems subject to interference.

12.7.2 Multipath Channels

Consider a DS binary PSK communication system operating over a multipath channel that has more than one path from the transmitter to the receiver. Such multiple paths may be due to atmospheric reflection or refraction, or reflections from buildings or other objects, and may result in fluctuations in the received signal level. The different paths may consist of several discrete paths each with a different attenuation and time delay, or they might consist of a continuum of paths. Figure 12.35 illustrates a communication link with two discrete paths. The multipath wave is delayed by some time τ, compared with the direct wave. In television receivers, signals such as these cause "ghosts," or under extreme conditions, complete loss of picture synchronization.

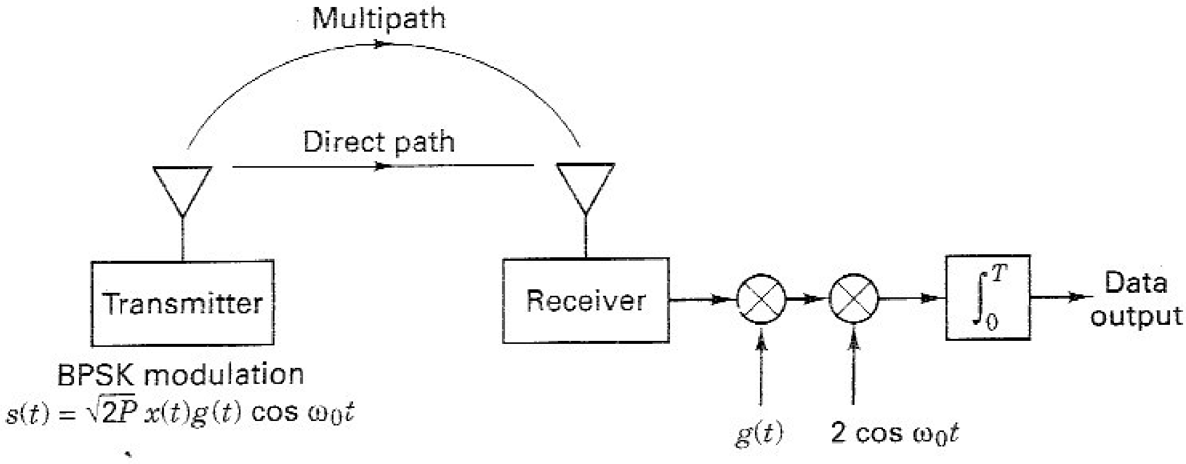


Figure 12.35 Direct-sequence BPSK system operating over a multipath channel.

In a direct-sequence spread-spectrum system, if we assume that the receiver is synchronized to the time delay and RF phase of the direct path, the received signal can be expressed as

$$r(t) = Ax(t)g(t) \cos \omega_0 t + \alpha Ax(t - \tau)g(t - \tau) \cos (\omega_0 t + \theta) + n(t) \quad (12.59)$$

where $x(t)$ is the data signal, $g(t)$ is the code signal, $n(t)$ is a zero-mean Gaussian noise process, and τ is the differential time delay between the two paths, assumed to be in the interval $\theta < \tau < T$. The angle θ is a random phase, assumed to be uniformly distributed in the range $(0, 2\pi)$, and α is the attenuation of the multipath signal relative to the direct path signal. For the receiver, synchronized to the direct path signal, the output of the correlator can be written as

$$z(t = T) = \int_0^T [Ax(t)g^2(t) \cos \omega_0 t \quad (12.60)$$

$$+ \alpha Ax(t - \tau)g(t)g(t - \tau) \cos (\omega_0 t + \theta) + n(t)g(t)]2 \cos \omega_0 t \, dt$$

where $g^2(t) = 1$. Also, for $\tau > T_c$, $g(t)g(t - \tau) \approx 0$ (for codes with long periods), where T_c is the chip duration. Therefore, if T_c is less than the differential time delay between the multipath and direct path signals, we can write

$$z(t = T) = \int_0^T 2Ax(t)\cos^2\omega_0 t + 2n(t)g(t)\cos \omega_0 t \, dt = Ax(T) + n_0(T) \quad (12.61)$$

where $n_0(T)$ is a zero-mean Gaussian random variable. We see that the spread-spectrum system, similar to the case of CDMA, effectively eliminates the multipath interference by virtue of its code-correlation receiver.

If frequency hopping (FH) is used against the multipath problem, improvement in system performance is also possible but through a different mechanism. FH receivers avoid multipath losses by rapid changes in the transmitter frequency band, thus avoiding the interference by changing the receiver band position before the arrival of the multipath signal.

12.7.3 The FCC Part 15 Rules for Spread-Spectrum Systems

In the United States, the Federal Communications Commission (FCC) allows the general unlicensed operation of very lower power (less than 1 mW) radio equipment freely, except in certain restricted frequency bands. In 1985, Dr. Michael Marcus of the FCC was responsible for allowing higher power (up to 1 W) spread-spectrum radios in some of the bands, referred to as *Industrial, Scientific,* and *Medical* (ISM). The rules of allowable electromagnetic radiation for unlicensed devices appear in the *Code of Federal Regulations* (CFR) Title 47, Part 15; they are known simply as the Part-15 rules (wherein Section 15.247 covers spread spectrum).

The ISM frequency bands are used for instruments (e.g., medical diathermy equipment) as well as for critical government systems (e.g., radio location equipment) that radiate strong electromagnetic fields which can cause interference to other users. The ISM bands are particularly noisy bands. An unlicensed radio can be thought of as an "unwelcome guest" in a licensee's band. An unlicensed radio must be able to suffer interference but is not permitted to cause any interference to a licensed user.

For frequency-hopping systems, the Part-15 rules require that the average time of occupancy on any frequency shall not be greater than 0.4 second (or a minimum hopping rate of 2.5 hops/s). For direct sequence systems, the minimum required processing gain is 10 dB. For hybrid systems employing both direct sequence and frequency hopping, the minimum required processing gain is 17 dB. Three ISM spectral regions were designated for the operation of unlicensed spread-spectrum radios. Some of the details regarding the operation in these bands are shown in Table 12.1.

As a result of allowing higher power limits and no FCC licensing as described previously, commercial companies have introduced a wide range of innovative spread-spectrum radios capable of communications over greater distances than earlier low-power narrowband unlicensed radios. Some of these products include radios that link office equipment (e.g., shared printer, wireless local area net-

TABLE 12.1 Spread-Spectrum Operation Under Part-15 Rules

ISM Band	Total Bandwidth	Max. Bandwidth per Channel for FH*	Min. Number of Hopping Frequencies per Channel	Min. Bandwidth per Channel for DS*
902–928 MHz	26 MHz	500 kHz	25–50**	500 kHz
2.4000–2.4835 Ghz	83.5 MHz	1 MHz	75	500 kHz
5.7250–5.8500 GHz	125 MHz	1 MHz	75	500 kHz

*Maximum bandwidth per channel for frequency hopping is defined as the 20 dB bandwidth; minimum bandwidth per channel for direct sequence is defined as the 6 dB bandwidth.

**FH channels with bandwidth less than 250 kHz require at least 50 hopping frequencies per channel; FH channels with bandwidth greater than 250 kHz require at least 25 hopping frequencies per channel.

works), cordless telephones, wireless point-of-sales equipment (e.g., cash registers, bar-code readers).

12.7.4 Direct Sequence versus Frequency Hopping

Without interference from other radios and in free space, both direct-sequence (DS) and frequency-hopping (FH) spread-spectrum radios can, in theory, give the same performance. For mobile applications with large multipath delays, DS represents a reliable mitigation method, because such signaling renders all multipath signal copies that are delayed by more than one chip time from the direct signal as "invisible" to the receiver. (See Section 12.7.2.) FH systems can provide the same mitigation, only if the hopping rate is faster than the symbol rate, and if the hopping bandwidth is large. (See Chapter 15.)

Implementing a fast frequency-hopping (FFH) radio can be costly due to the need for high-speed frequency synthesizers. Consequently, hopping rates of commercial FH radios are generally slow compared with the data rate, and hence such systems behave like narrowband radios. Slow frequency hopping (SFH) and DS signaling each experience somewhat different interference. SFH radios typically suffer occasional strong bursty errors, while DS radios encounter more randomly distributed errors that are continuous and lower level. For high data rates, the impact of multipath tends to degrade such SFH radios more than DS radios. To mitigate the effects of bursty errors in FH radios, interleaving would have to be performed over long time durations. (See Chapter 15.) SFH is used for providing diversity in fixed wireless access applications or slowly moving systems, or merely to meet the Part-15 Rules. For commercial applications, implementation of DS radios with large processing gain can also be costly due to the need for high-speed circuits; thus, the processing gain for such radios is usually limited to less than 20 dB to avoid having to use high-speed circuits [29].

Example 12.5 Detection of Signals Buried in the Noise

In Section 12.1.1.1 it was shown that spread-spectrum techniques offer no error-performance advantage against thermal noise. In this example, we show that whatever value of received E_b/N_0 might be available to a narrowband system remains the same after spreading. Although there is no error-performance advantage against thermal noise, there is *no disadvantage* either, which makes it an attractive option for building systems to meet the FCC Part-15 Rules, as described in Section 12.7.3, and for multiple access systems (e.g., CDMA systems meeting Interim Standard IS-95).

Direct-sequence spread-spectrum techniques allow for the detection of signals that have power-spectral density (psd) levels well below the noise. Consider Figure 12.36a illustrating the received psd of a communication signal with an ideal rectangular shape, having an intensity of $S_0(f) = 10^{-5}$ W/Hz over a bandwidth of 1 MHz. Assume that the transmitted data rate is $R = 10^6$ bits/s. Also, assume AWGN noise and a received psd level of $N_0(f) = 10^{-6}$ W/Hz (not drawn to scale) at all frequencies. Find the received E_b/N_0 for this narrowband case. Next, consider that the communication signal is spread over a spread-spectrum bandwidth of $W_{ss} = 10^8$ Hz, as shown in Figure 12.36b, keeping the total average signal power the same as in the narrowband case

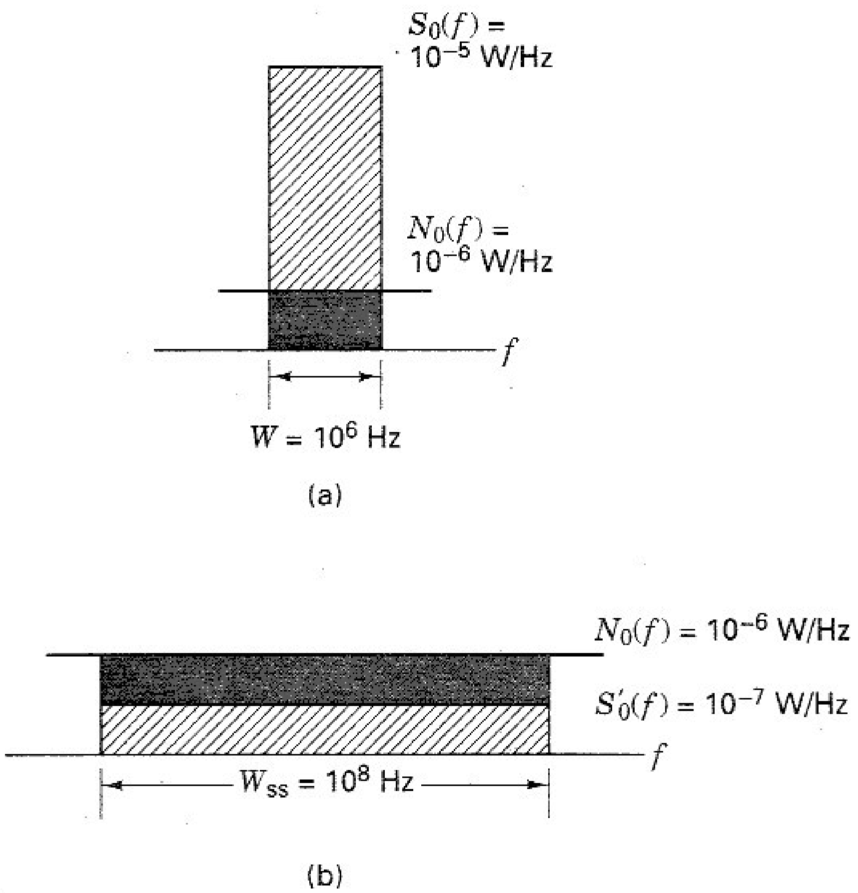


Figure 12.36 Signal and noise power spectral densities (a) Before spreading (b) After spreading.

(not drawn to scale). Show that with a spread-spectrum receiver, the received E_b/N_0 is the same as in the narrowband case, and hence the error performance is unchanged.

Solution

Before spreading, the total average power of the communication signal is $S = 10^{-5}$ W/Hz × 10^6 Hz = 10 W, and the total average noise power is $N = 10^{-6}$ W/Hz × 10^6 Hz = 1 Watt. The received E_b/N_0 is

$$\frac{E_b}{N_0} = \frac{S/R}{N_0} = \frac{10\ \text{W}/10^6\ \text{bits/s}}{10^{-6}\ \text{W/Hz}} = 10 \text{ or } 10\ \text{dB}$$

After spreading, the psd, $S'_0(f)$, of the communication signal is reduced by the same factor (two orders of magnitude) as the increase in bandwidth; hence, the total average power of the communication signal is still 10 W. However, since the noise was assumed to be AWGN, its psd is not reduced; hence, the total average noise power is now equal to $N' = 10^{-6}$ W/Hz × 10^8 Hz = 100 W. Therefore, after spreading, the received E_b/N_0 can be expressed as

$$\frac{E_b}{N_0} = \frac{S/R}{N'/W_{ss}} = \frac{S}{N'}\left(\frac{W_{ss}}{R}\right) = \frac{S}{N'}G_p = \frac{10\ \text{W}}{100\ \text{W}} \times 100 = 10 \text{ or } 10\ \text{dB}$$

where $G_p = W_{ss}/R = 100$ is the processing gain. The process of detecting direct-sequence spread-spectrum signals "buried in noise" does not lend itself to an intuitive illustration, as can be verified in Figure 12.36b. Similarly, in the expression for received E_b/N_0 after spreading, the communication signal power is only 10 W and the noise power is 100 W, leading to the same lack of intuition regarding the detectability of the signal. It is the processing gain (which is not very visual) that allows us to receive the same value of E_b/N_0 as in the narrowband case.

12.8 CELLULAR SYSTEMS

Wireless personal communication systems, particularly cellular systems are relatively young applications of the communications technology. The following list includes some of the events that illustrate the evolution of this ever-growing business:

- 1921 Radio dispatch service initiated for police cars in Detroit, Michigan.
- 1934 Amplitude modulation (AM) mobile communication systems used by hundreds of state and municipal police forces in the U.S.
- 1946 Radiotelephone connections made to the public-switched telephone network (PSTN).
- 1968 Development of the cellular telephony concept at Bell Laboratories.
- 1981 Ericsson Corporation's Nordic Mobile Telephone (NMT) in Scandinavian countries becomes the first cellular system fielded.
- 1983 Cellular service in the United States—called the Advanced Mobile Phone System (AMPS) and using frequency modulation (FM)—placed in service in Chicago by Ameritech Corporation.
- 1990s Second generation digital cellular deployed throughout the world. The Global system for Mobile (GSM) Communications becomes the pan-European standard. (Prior to GSM, many different cellular systems operating in Europe became operationally impractical.)
- 1990s Second generation digital systems known as IS-54 and its successor IS-136 (TDMA), and IS-95 (CDMA) become operational in the United States.
- 2000s Third-generation digital systems standardized at the network level to allow world-wide roaming start becoming operational. They offer enhanced services, such as connection to various PSTN systems with a single phone, and connecting to high data rate packet systems such as Internet Protocol (IP) networks.

12.8.1 Direct Sequence CDMA

Figures 11.3 and 11.7 depict sharing a communications resource via FDMA or TDMA. In the case of FDMA, different frequency bands are orthogonal to one another (assuming ideal filtering), and in the case of TDMA, different time slots are orthogonal to one another (assuming perfect timing). One can visualize a similar orthogonality among different channels in the case of frequency-hopping CDMA (as shown in Figure 11.14) if the codes that control the frequency hopping operate in