UNIT IV SYNCHRONIZATION OF SS SYSTEMS

Acquisition and tracking in DS SS – FH SS receivers – Sequential estimation – Matched filter techniques of acquisition and tracking –Delay locked loop – Tau Dither loop - Jamming considerations – Broad band – partial – multitone – repeat back jamming

4.1 SYNCHRONIZATION

For both DS and FH spread-spectrum systems, a receiver must employ a *synchronized* replica of the spreading or code signal to demodulate the received signal successfully. The process of synchronizing the locally generated spreading signal with the received spread-spectrum signal is usually accomplished in two steps. The first step, called *acquisition,* consists of bringing the two spreading signals into *coarse* alignment with one another. Once the received spread-spectrum signal has been acquired, the second step, called *tracking,* takes over and continuously maintains the best possible waveform *fine* alignment by means of a feedback loop.

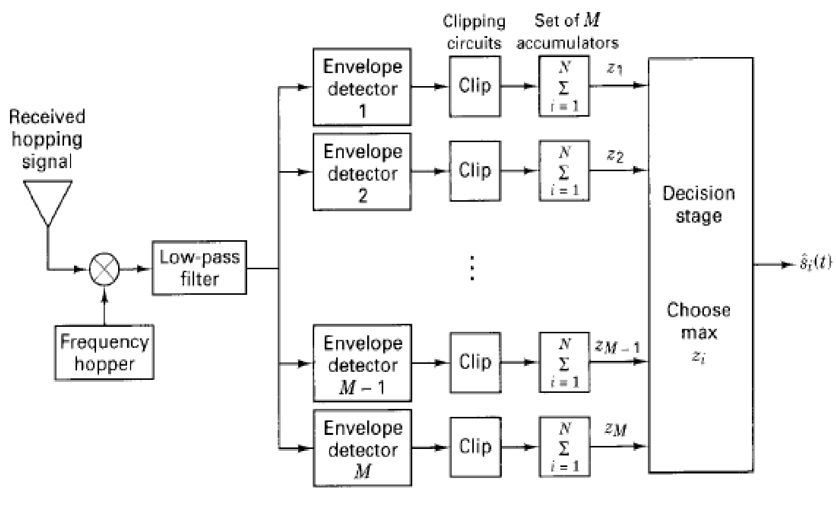

Figure 4.1 FFH/MFSK demodulator.

4.2 ACQUISITION (COARSE SYNCHRONIZATION)

The acquisition problem is one of searching throughout a region of time and frequency uncertainty in order to synchronize the received spread-spectrum signal with the locally generated spreading signal. Acquisition schemes can be classified as coherent or noncoherent. Since the despreading process typically takes place before carrier synchronization, and therefore the carrier phase is unknown at this point, most acquisition schemes utilize noncoherent detection. When determining the limits of the uncertainty in time and frequency, the following items must be considered:

1. Uncertainty in the distance between the transmitter and the receiver translates into uncertainty in the amount of propagation delay
2. Relative clock instabilities between the transmitter and the receiver result in phase differences between the transmitter and receiver spreading signals that will tend to grow as a function of elapsed time between synchronization
3. Uncertainty of the receiver's relative velocity with respect to the transmitter translates into uncertainty in the value of Doppler frequency offset of the incoming signal
4. Relative oscillator instabilities between the transmitter and the receiver result in frequency offsets between the two signals

4.3 CORRELATOR STRUCTURES

A common feature of all acquisition methods is that the received signal and the locally generated signal are first correlated to produce a measure of similarity between the two. This measure is then compared with a threshold to decide if the two signals are in synchronism. If they are, the tracking loop takes over. If they are not, the acquisition procedure provides for a phase or frequency change in the locally generated code as a part of a systematic search through the receiver's phase and frequency uncertainty region, and another correlation is attempted.

Consider the direct-sequence *parallel-search* acquisition system shown in Figure 4.2. The locally generated code *g(t)* is available with delays that are spaced one-half chip ($T_C/2$) apart. If the time uncertainty between the local code and the received code is N_C chips and a complete parallel search of the entire time uncertainty region is to be accomplished in a single search time, $2N_C$ correlators are used. Each correlator simultaneously examines a sequence of λ chips, after which the $2N_C$, correlator outputs are compared. The locally generated code, corresponding to the correlator with the largest output is chosen.

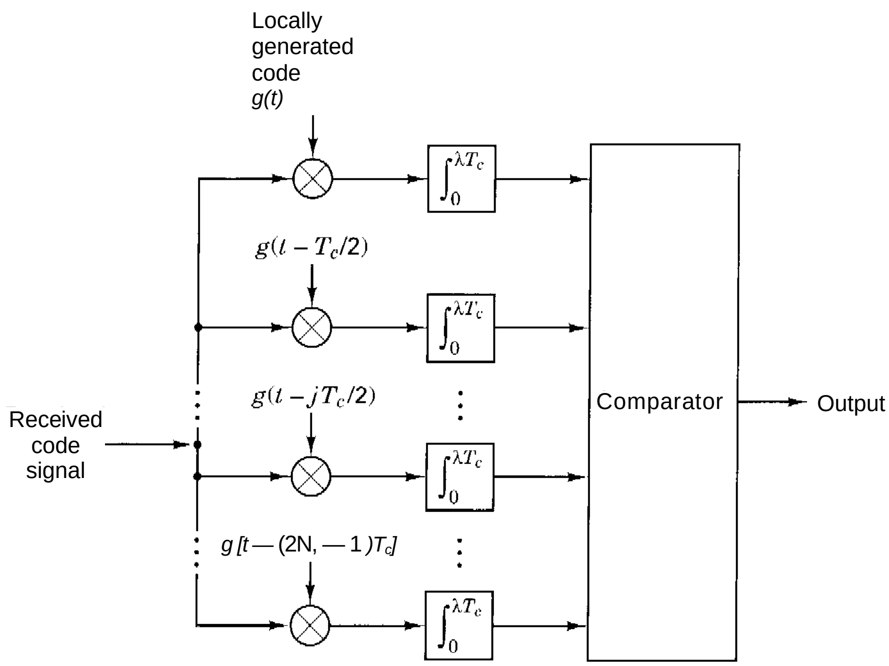

Figure 4.2 Direct-sequence parallel search acquisition.

Conceptually, this is the simplest of the search techniques; it considers all possible code positions (or fractional code positions) in parallel and uses a maximum likelihood algorithm for acquiring the code. Each detector output pertains to the identical observation of received signal plus noise. As λ increases, the synchronization error probability (i.e., the probability of choosing the incorrect code alignment) decreases. Thus, λ is chosen as a compromise between minimizing the probability of a synchronization error and minimizing the time to acquire.

Figure 4.3 illustrates a simple acquisition scheme for a frequency hopping system. Assume that a sequence of N consecutive frequencies from the hop sequence is chosen as a synchronization pattern. The N noncoherent matched filters each consists of a mixer followed by a bandpass filter (BPF) and a square-law envelope detector. If the frequency hopping sequence is $f_1,f_2, \ldots \ldots\ldots, f_N$, delays are inserted into the matched filters so that when the correct frequency hopping sequence appears, the system produces a large output, indicating detection of the synchronization sequence. Acquisition can be accomplished rapidly because all possible code offsets are examined simultaneously. The presence of band-pass filters (BPF) indicates that the local oscillator frequencies $f_1,f_2, \ldots \ldots\ldots, f_N$ are chosen to have offsets by some intermediate frequency (IF) from the expected received hop sequence. The same system can be implemented with local oscillator frequencies chosen (without offsets) so that the mixers yield baseband signals, and thus the filters would need to be low-pass filters (LPF). The mixers are typically complex, yielding in-phase and quadrature terms.

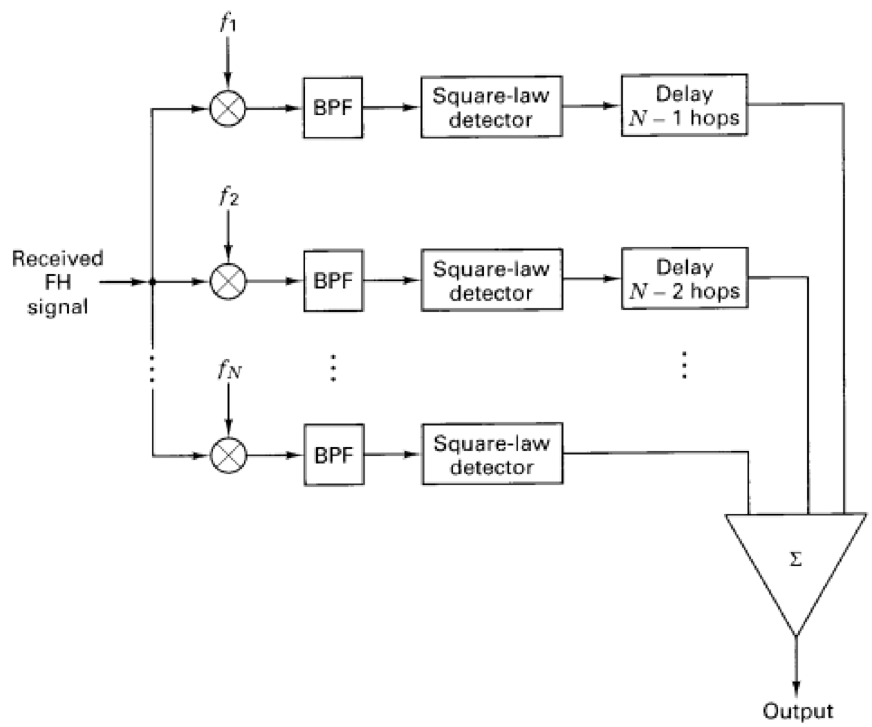

Figure 4.3 Frequency hopping acquisition scheme.

If, during each correlation, λ chips (each chip having a duration of T) are examined, the maximum time required for a fully parallel search is

$$(T_{\mathrm{acq}})_{\max} = \lambda T_c$$

The mean acquisition time of a parallel search system can be approximated by noting that after integrating over λ chips, a correct decision will be made with probability P_D, called the *probability of detection.* If an incorrect output is chosen, an additional λ chips are again examined to make a determination of the correct output. Therefore, on the average, the acquisition time is

$$\begin{aligned}\overline{T}_{\mathrm{acq}} &= \lambda T_c P_D + 2\lambda T_c P_D(1 - P_D) + 3\lambda T_c P_D(1 - P_D)^2 + \cdots \\ &= \frac{\lambda T_c}{P_D}\end{aligned}$$

Since the required number of correlators or matched filters can be prohibitively large, fully parallel acquisition techniques are not usually used. In place of Figures 4.2 and 4.3, a single correlator or matched filter can be implemented that will *serially search* until synchronization is achieved. Naturally, trade-offs between fully parallel, fully serial, and combinations of the two involve hardware complexity versus time to acquire for the same uncertainty and chip rate.

4.3.1 Serial Search

A popular strategy for the acquisition of spread-spectrum signals is to use a single correlator or matched filter to serially search for the correct phase of the DS code signal or the correct hopping pattern of the FH signal. A considerable reduction in complexity, size, and cost can be achieved by a serial implementation that repeats the correlation procedure for each possible sequence shift. Figures 4.4 and 4.5 illustrate the basic configuration for DS and FH spread-spectrum schemes, respectively. In a stepped serial acquisition scheme for a DS system, the timing epoch of the local PN code is set, and the locally generated PN signal is correlated with the incoming PN signal. At fixed examination intervals of λT_C, (search dwell time), where $\lambda >> 1$, the output signal is compared to a preset threshold. If the output is below the threshold, the phase of the locally generated code signal is incremented by a fraction (usually one-half) of a chip and the correlation is reexamined. When the threshold is exceeded, the PN code is assumed to have been acquired, the phaseincrementing process of the local code is inhibited, and the code tracking procedure will be initiated. In a similar scheme for FH systems, shown in Figure 4.5, the PN code generator controls the frequency hopper. Acquisition is accomplished when the local hopping is aligned with that of the received signal.

The maximum time required for a fully serial DS search, assuming that the search proceeds in half-chip increments, is

$$(T_{\text{acq}})_{\max} = 2N_c \lambda T_c$$

where the uncertainty region to be searched is Nc, chips long.

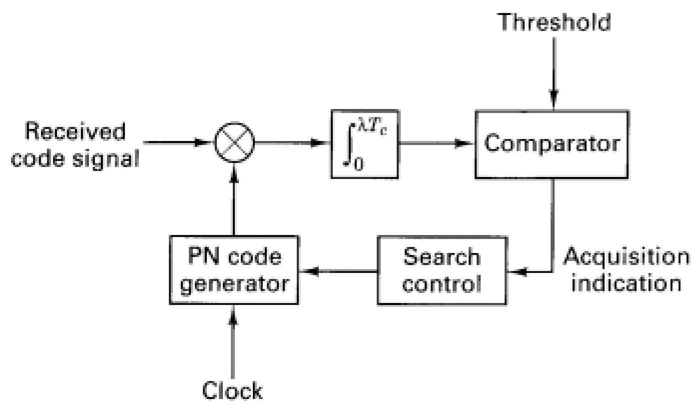

Figure 4.4 Direct-sequence serial search acquisition.

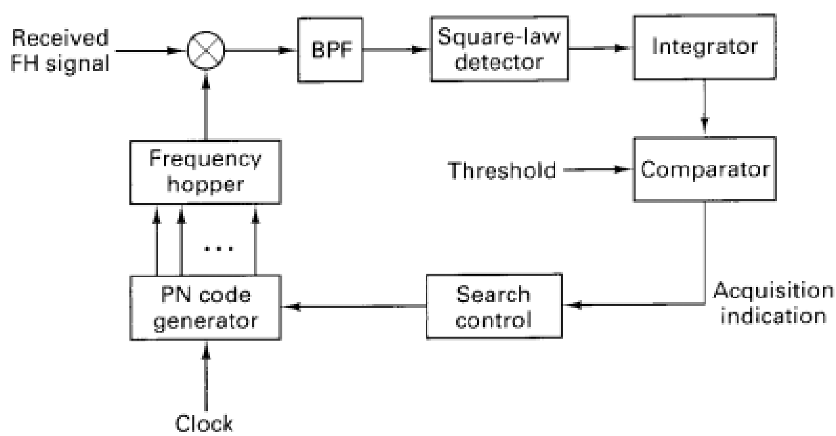

Figure 4.5 Frequency hopping serial search acquisition.

The mean acquisition lime of a serial DS search system can be shown, for *Nc»1/2* chip, to be

$$\overline{T}_{\text{acq}} = \frac{(2 - P_D)(1 + KP_{\text{FA}})}{P_D}(N_c \lambda T_c)$$

where λTc is the search dwell time, P_D is the probability of correct detection, and P_{FA} is the probability of false alarm. We can regard the time interval $K\lambda T_C$. where K » 1, as the time needed to verify a detection. Therefore, in the event of a false alarm, $K\lambda T_C$, seconds is the time penalty incurred. For *Nc»1/2* chip and *K «2* Nc, the variance of the acquisition time is

$$(\text{var})_{\text{acq}} = (2N_c \lambda T_c)^2(1 + KP_{\text{FA}})\left(\frac{1}{12} + \frac{1}{P_D^2} - \frac{1}{P_D}\right)$$

4.3.2 Sequential Estimation

Another search technique, called *rapid acquisition by sequential estimation* (RASE), proposed by Ward, is illustrated in Figure 4.6. The switch is initially in position 1. The RASE system enters its best estimate of the first *n* received code chips into the *n* stages of its local PN generator. The fully loaded register defines a starting state from which the generator begins its operation. A PN sequence has the property that the next combination of register states depends only on the present combination of states. Therefore, if the first *n* received chips are correctly estimated, all the following chips from the local PN generator will be correctly generated. The switch is next thrown to position 2. If the starting state had been correctly estimated, the local generator generates the same sequences as the incoming waveform, in the absence of noise. If the correlator output after λT_C exceeds a preset threshold level, it is assumed that synchronization has occurred. If the output is less than the threshold, the switch is returned to position 1, the register is reloaded with estimates of the next *n* received chips, and the procedure is repeated. Once synchronization has occurred, the system no longer needs estimates of the input code chips. We can calculate the *minimum* acquisition time for the case when no noise is present. The first *n* chips will be correctly loaded into the register, and therefore, the acquisition time is

$$T_{\text{acq}} = nT,$$

While the RASE system has a rapid acquisition capability it has the drawback of being highly vulnerable to noise and interference signals. The reason for this is that the estimation process consists of a simple chip-by-chip hard-decision demodulation, without using the interference rejection benefits of the PN code.

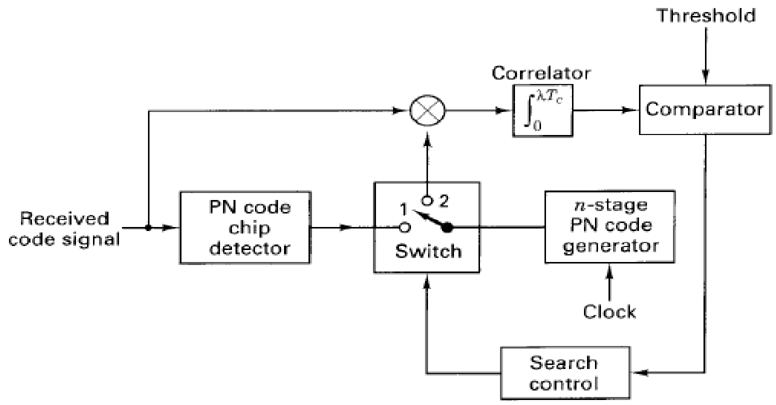

Figure 4. Rapid acquisition by sequential estimation.

4.4 TRACKING (FINE SYNCHRONIZATION)

Once acquisition or coarse synchronization is completed, tracking or fine synchronization takes place. Tracking code loops can be classified as coherent or noncoherent. A coherent loop is one in which the carrier frequency and phase are known exactly so that the loop can operate on a baseband signal. A noncoherent loop is one in which the carrier frequency is not known exactly (due to Doppler effects, for example), nor is the phase. In most instances, since the carrier frequency and phase are not known exactly, a priori, a noncoherent code loop is used to track the received PN code. Tracking loops are further classified as a *full-time* early-late tracking loop, often referred to as a *delay-locked loop* (DLL), or as a *time-shared* early-late tracking loop, frequently referred to as a *tau-dither loop* (TDL). A basic noncoherent DLL loop for a direct-sequence spread-spectrum system using binary phase shift keying (BPSK) is shown in Figure 4.7. The data *x(t)* and the code *g(t)* each modulate the carrier wave using BPSK, and as before in the absence of noise and interference, the received waveform can be expressed as

$$r(t) = A\sqrt{2P}\, x(t) g(t) \cos(\omega_0 t + \phi)$$

where the constant A is a system gain parameter and Φ is a random phase angle in the range $(0, 2\pi)$. The locally generated code of the tracking loop is offset in phase from the incoming *g(t)* by a time τ, where $\tau < T_c/2$. The loop provides *fine* synchronization by first generating two PN sequences $g(t + T_c/2 + \tau)$ and $g(t - T_c/2 + \tau)$ delayed from each other by one chip. The two bandpass filters are designed to pass the data and to average the product of *g(t)* and the two PN sequences $g(t \pm T_c/2 + \tau)$. The square-law envelope detector eliminates the data since $|x(t)| = 1$. The output of each envelope detector is given approximately by

$$E_D \approx \mathbf{E}\left\{\left| g(t)g\left(t \pm \frac{T_c}{2} + \tau\right)\right|\right\} = \left|R_g\left(\tau \pm \frac{T_c}{2}\right)\right|$$

where the operator E{•} means *expected value* and $R_g(x)$ is the autocorrelation function of the PN sequence. The feedback signal $Y(\tau)$ is shown in Figure 4.8. When τ is positive, the feedback signal $Y(\tau)$ instructs the voltage-controlled oscillator (VCO) to increase its frequency, thereby forcing τ to decrease, and when τ is negative, $Y(\tau)$ instructs the VCO to decrease, thereby forcing τ to increase. When τ is a suitably small number, $g(t)g(t + \tau) \approx 1$, yielding the despread signal *Z(t),* which is then applied to the input of a conventional data demodulator.

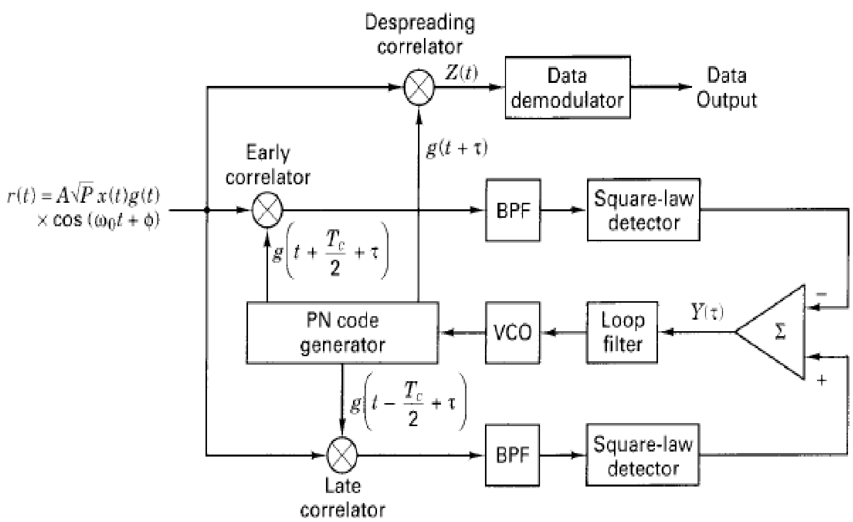

Figure 4.7 Delay-locked loop for tracking direct-sequence signals.

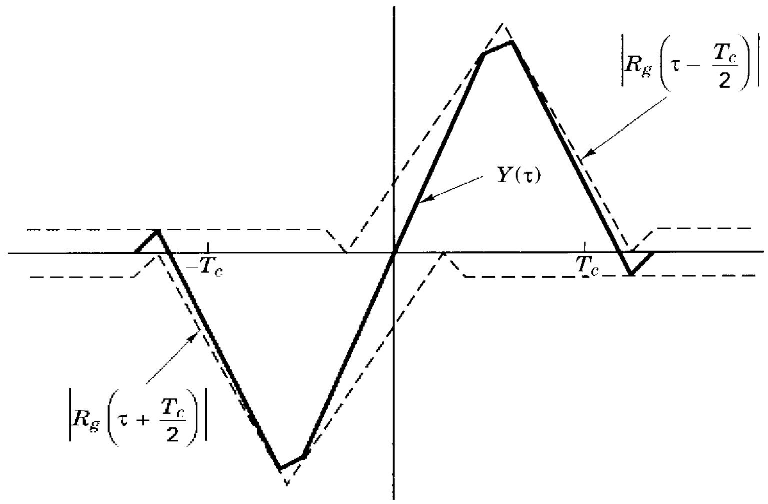

Figure 4.8 DLL feedback signal Y(τ)

A problem with the DLL is that the early and late arms must be precisely gain balanced or else the feedback signal Y(τ) will be offset and will not produce a zero signal when the error is zero. This problem is solved by using a time-shared tracking loop in place of the full-time delay-locked loop. The time-shared loop time shares the use of the early-late correlators. The main advantages are that only one correlator need be used in the design of the loop and further, that dc offset problems are reduced.

A problem with some control loops is that if things are going well and the loop is tracking accurately, the control signal is essentially zero. When the control signal is zero, the loop can get "confused" and do erratic things. This is especially the case in more sophisticated tracking loops that modify their own loop gain in response to the perceived environment. An offshoot of the time-shared tracking loop, called the *tau-dither loop* (TDL), shown in Figure 4.9, tends to deal with this potential problem by intentionally injecting a small error in the tracking correction, so that the loop kind of vibrates around the correct answer. This vibration is typically small, so that the loss in performance is minimal. This design has the advantage that only one correlator is needed to provide the code *tracking* function *and* the *despreading* function. Just as in the case of a DLL, the received signal is correlated with an early and a late version of the locally generated PN code. As shown in Figure 4.9, the PN code generator is driven by a clock signal whose phase is *dithered* back and forth with a square-wave switching function; this eliminates the necessity of ensuring identical transfer functions of the early and late paths. The signal-to-noise performance of the TDL is only about 1.1 dB worse than that of the DLL if the arm filters are designed properly.

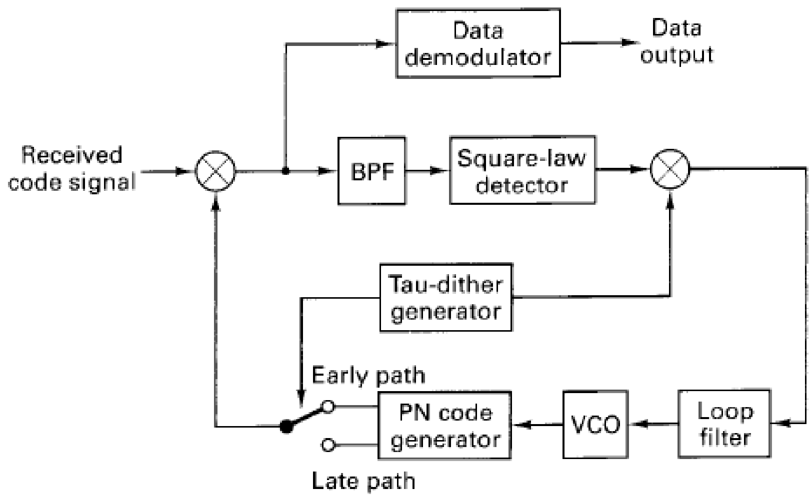

Figure 4.9 Tau-dither tracking loop.

4.5 JAMMING CONSIDERATIONS

4.5.1 The Jamming Game

The goals of a jammer are to deny reliable communications to his adversary and to accomplish this at minimum cost. The goals of the communicator are to develop a jam-resistant communication system under the following assumptions:

- complete invulnerability is not possible
- the jammer has a priori knowledge of most system parameters, such as frequency bands, timing, traffic, and so on
- the jammer has *no* a priori knowledge of the PN spreading or hopping codes

The signaling waveform should be designed so that the jammer cannot gain any appreciable jamming advantage by choosing a jammer waveform and strategy other than wideband Gaussian noise. The fundamental design rule in specifying a jam-resistant system is to make it as costly as possible for the jammer to succeed in jamming the system.

4.5.2 Jammer Waveforms

There are many different waveforms that can be used for jamming communication systems. The most appropriate choice depends on the targeted system. Figure 4.10 shows power spectral density plots of examples of jammer waveforms versus a communicator's frequency hopped M-ary FSK (FH/MFSK) tone. The range of the abscissa represents the spread-spectrum bandwidth W_{SS}. The three columns in the figure represent three instances in time (three hop times) when symbols having spectra G_1, *G2,* and *G3,* respectively, are being transmitted.

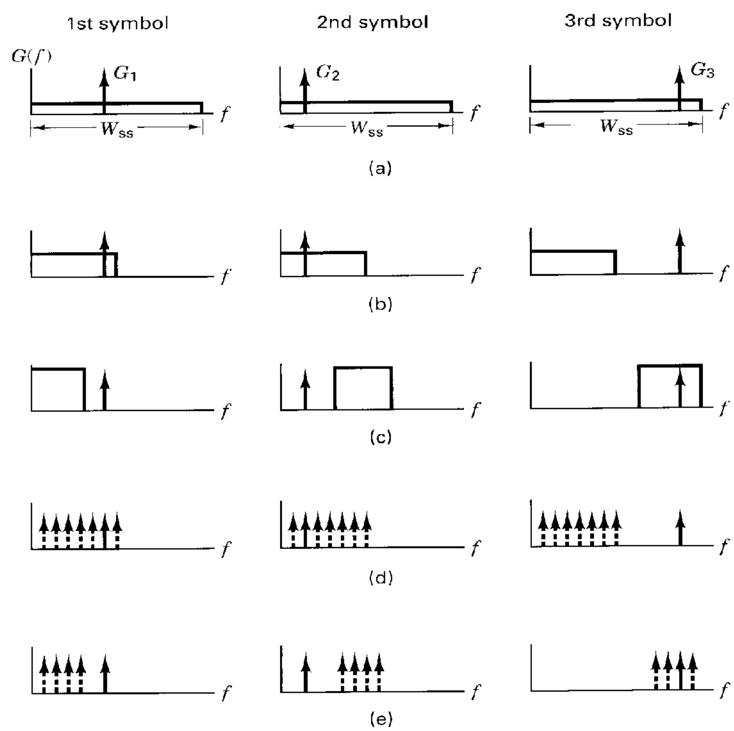

Figure 4.10 Jammer waveforms. (a) Full-band noise. (b) Partial-band noise. (c) Stepped noise. (d) Partial-band tones. (e) Stepped tones.

Figure 4.10a illustrates a relatively low-level noise jammer occupying the full spread-spectrum bandwidth. In Figure 4.10b the jammer strategy is to trade bandwidth occupancy for greater power spectral density (the total power, or area under the curve, remains the same). The figure indicates that in this case, the jammer noise does not always share the same bandwidth region as the signal, but when it does, the effect can be destructive. In Figure 4.10c the noise jammer strategy is again to jam only part of the hand, so that the jammer power spectral density can be increased, but in this case the jammer steps through different regions of the band at random times, thus preventing the communicator from using adaptive techniques to avoid the jamming. In Figure 4.10d and 4.10e the jammer uses a group of tones, instead of a continuous frequency band, in partial-band (Figure 4.10d) and stepped fashion (Figure 4.10e). This is a technique most often used against FH systems. Another jamming technique, not shown in Figure 4.10, is a pulse jammer, consisting of pulse-modulated bandlimited noise. Unless otherwise stated, the jammer waveform is wideband noise and that the jammer strategy is to jam the entire bandwidth W„ continuously.

4.5.3 Tools of the Communicator

The usual design goal for an anti-jam (AJ) communication system is to force a jammer to expend its resources over

(1) a wide-frequency band

(2) for a maximum time and

(3) from a diversity of sites.

The most prevalent design options are

1. frequency diversity, by the use of direct-sequence and frequency-hopping spread-spectrum techniques
2. time diversity, by the use of time hopping
3. spatial discrimination, by the use of a narrow-beam antenna, which forces a jammer to enter the receiver via an antenna sidelobe and hence suffer, typically, a 20- to 25-d13 disadvantage, and
4. combinations of the previous three options

4.5.4 *J/S Ratio*

Link error performance as a function of thermal noise interference places an emphasis on the signal-to-noise ratio parameters—required E_b/N_o and available E_b/N_o for meeting a specified error performance. Similarly concerned with link error performance as a function of interference, the source of interference is assumed to be wideband Gaussian noise power from a jammer in addition to thermal noise. Therefore, the SNR of interest is $E_b/(N_o + J_0)$, where J_0 is the noise power spectral density due to the jammer. Unless otherwise specified, J_0 is assumed equal to J/Wss where J is the average received jammer power (jammer power referred to the receiver front end) and Wss is the spread-spectrum bandwidth. Since the jammer power is generally much greater than the thermal noise power, the SNR of interest in a jammed environment is usually taken to be E_b/J_o. Therefore, similar to the thermal noise case, (E_b/J_0)reqd as the bit energy per jammer noise power spectral density *required* for maintaining the link at a specified error probability. The parameter E_b can be written as

$$E_b = ST_b = \frac{S}{R}$$

where S is the received signal power, T_b the bit duration, and R the data rate in bits/s. Then we can express (E_b/J_0)reqd as

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

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

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

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.The communicator would like the communication system *not* to degrade at all.

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 gain. 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.

4.5.5 Anti-Jam Margin

The ratio $(J/S)_{reqd}$ is referred to as the *anti-jam (AJ) margin,* since it characterizes the system jammer-rejection capability but the AJ margin usually means the safety margin against a *particular threat.* Using the same approach the AJ margin is defined as

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

where $(E_b/J_0)_r$ is the (E_b/J_0)*actually received* and $(E_b/J_O)_r$ is expressed as

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

where $(J/S)r$, or simply J/S, is the ratio of the actually received jammer power to signal power. The expression for received E_b/I_o is similar to the previous equation, 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. Combining all the equations we have

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

4.6 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, Wss, with its fixed power, the jammer is referred to as a wideband or *broadband jammer,* and the jammer power spectral density is

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

The bit error probability *P B* for a coherently demodulated BPSK system (without channel coding) is

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

The single-sided noise power spectral density N_o represents thermal noise at the front end of the receiver. The presence of the jammer increases this noise power spectral density from N_o to $(N_0 + J_0)$. Thus the average bit error probability for a coherent BPSK system in the presence of broadband jamming is

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

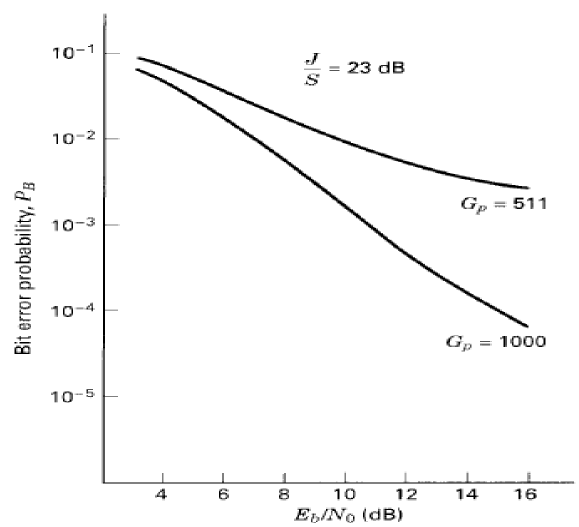

Figure 4.11 Bit-error probability Versus E_b/N_o for a given J/S ratio

When P_B is plotted versus E_b/N_o for a given *J/S* ratio, the resulting curves are such as shown in Figure 4.11. for two different 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 gain.

4.7 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 is given by

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

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 = pW,, the jammer noise power spectral density can be concentrated to a level $J_o/\ \rho$, 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 — ρ), and will be perturbed by jammer power with spectral density $J_o/\ \rho$, with probability p. Therefore, the average bit error probability is given by

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

Since, in a jamming environment, it is often the case that $J_o \gg N_o$, Equation is simplified to the form

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

Figure 4.12 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 $\rho = \rho_o$ that maximizes P_B. ρ_o decreases with increasing values of E_b/J_o. An expression for ρ_o is easily found by differentiation (setting $dP_B/dp = 0$ and solving for ρ). This yields

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

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 into the inverse linear.

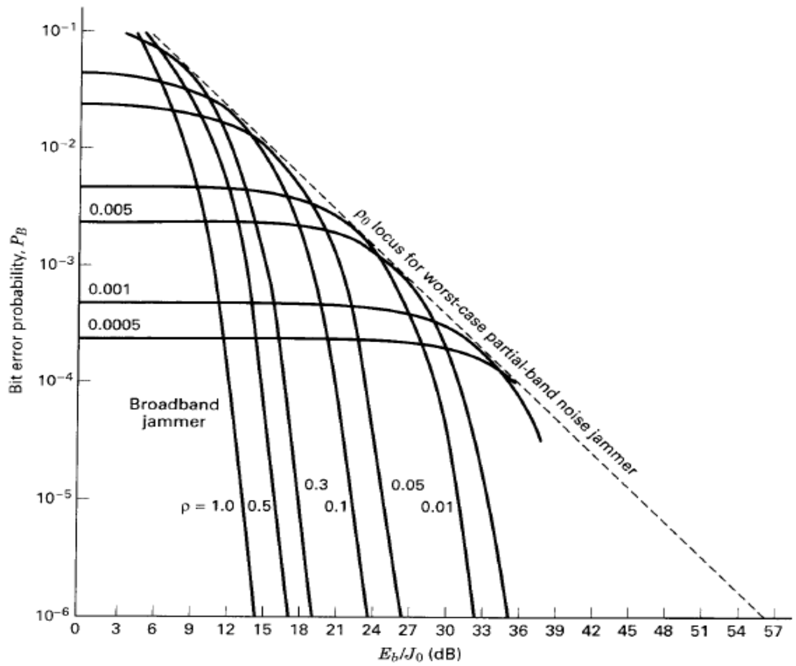

Figure 4.12 Partial band noise jammer (FH/BPSK signaling)

The ρ_o locus in Figure 4.12 illustrates the P_B versus E_b/J_0 performance for the worst-case partial-band jammer. Here at 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. 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. 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.

4.8 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}. 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. 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. 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.

In the FFH/MFSK demodulator of Figure 4.13, a chip-clipping circuit is shown between each envelope detector and accumulator. An 8-ary FSK frequency-hopping system with no diversity, indicated in Figure 4.13a, 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 14.13b. Each row in the figures represents one of the M = 8 accumulators. The presence of a signal in the accumulator is indicated by a vector. In Figure 4.13a 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 $J > S$, 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 4.13b, 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 4.13b, two of the jamming tones fall in the data band, but because they are clipped, there is no confusion about the correct symbol decision.

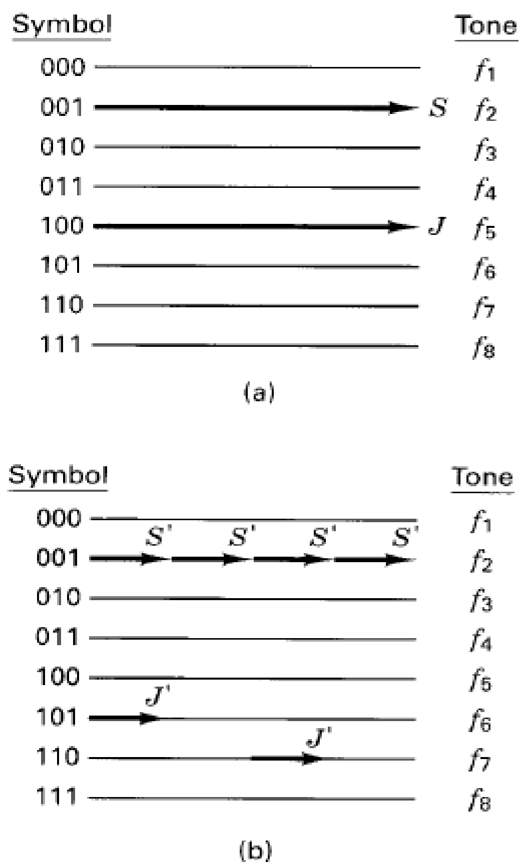

Figure 4.13 Fast hopping repeat versus tone jamming
a) One frequency hop b) Four frequency hops

4.9 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 < p < 1 of the time, then during this time, the jammer power spectral density is increased to a level J_o/p, thus maintaining a constant time-averaged power J (where $J = J_0 W_{ss}$ and W_s, is the system spread-spectrum bandwidth).

The bit error probability P_B for a coherently demodulated BPSK system (without channel coding) is given by

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

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

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

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

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

The jammer will, of course, attempt to choose the duty cycle p that maximizes *B.* Figure 12.30 illustrates P_B for various values of p. The value of $\rho = \rho_o$ that maximizes P_B decreases with increasing values of E_b/J_o, as was the case with partial-band jamming. By differentiating the Equation we have

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

which results in the maximum bit error probability

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

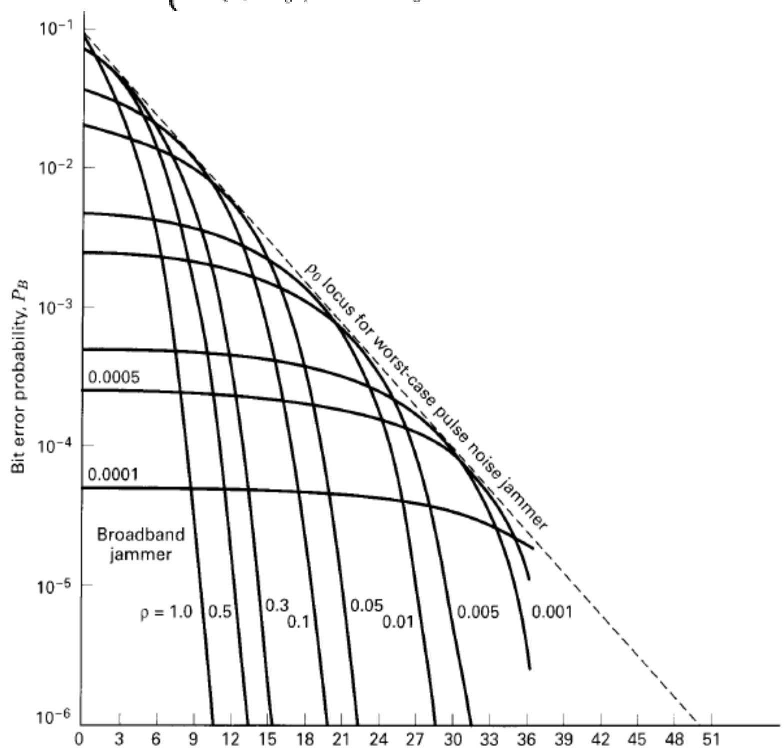

Figure 4.14 Pulse noise jammer (DS/BPSK signaling)

The effect of a worst-case pulse jammer upon a system with spread spectrum *but without coding* changes the complementary error function relationship into the inverse linear one. As a result, at an error probability of 10^{-6}, there is about a 40-dB difference in E_b/J_o between the broadband jammer and the worst-case pulse jammer. 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. 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.

4.10 REPEAT-BACK JAMMING

The measure of jammer-rejection capability, namely processing gain, G_r, 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. 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. 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.

One method to defeat the repeat-back jammer 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 will be unaffected by jamming at the frequency of the prior hop.