# CHAPTER 12.4
# SPREAD-SPECTRUM MODULATION

**Myron D. Egtvedt**

## *SPREAD-SIGNAL MODULATION*

In a receiver designed exactly for a specified set of possible transmitted waveforms (in the presence of white noise and in the absence of such propagation defects as multipath and dispersion), the performance of a matched filter or cross-correlation detector depends only on the ratio of signal energy to noise power density $E/n_o$, where $E$ is the received energy in one information symbol and $n_o/2$ is the rf noise density at the receiver input. Since signal bandwidth has no effect on performance in white noise, it is interesting to examine the effect of spreading the signal bandwidth in situations involving jamming, message and traffic-density security, and transmission security. Other applications include random-multiple-access communication channels, multipath propagation analysis, and ranging.

The information-symbol waveform can be characterized by its time-bandwidth (TW) product. Consider a binary system with the information symbol defined as a *bit* (of time duration $T$), while the fundamental component of the binary waveform is called a *chip*. For this *direct-sequence* system, the ratio (chips per bit) is equal to the TW product. An additional requirement on the symbol waveforms is that their cross-correlation with each other and the noise or extraneous signals be minimal.

Spread-spectrum systems occupy a signal bandwidth much larger (>10) than the information bandwidth, while the conventional systems have a TW of well under 10. FM with a high modulation index might slightly exceed 10 but is not optimally detectable and has a processing gain only above a predetection signal-to-noise threshold.

## *NOMENCLATURE OF SECURE SYSTEMS*

While terminology is not subject to rigorous definition, the following terms apply to the following material:

*Security and privacy*. Relate to the protection of the signal from an unauthorized receiver. They are differentiated by the sophistication required. Privacy protects against a casual listener with little or no analytical equipment, while security implies an interceptor familiar with the principles and using an analytical approach to learn the *key*. Protection requirements must be defined in terms of the interceptor's applied capability *and* the time value of the message. Various forms of *protection* include:

*Crypto security*. Protects the information content, generally without increasing the TW product.

*Antijamming (AJ) security*. Spreads the signal spectrum to provide discrimination against energy-limited interference by using cross-correlation or matched-filter detectors. The interference may be natural (impulse noise), inadvertent (as in amateur radio or aircraft channels), or deliberate (where the jammer may transmit continuous or burst cw, swept cw, narrow-band noise, wide-band noise, or replica or deception waveforms).

*Traffic-density security*. Involves capability to switch data rates without altering the apparent characteristics of the spread-spectrum waveform. The TW product (processing gain) is varied inversely with the data rates.

*Transmission security*. Involves spreading the bandwidth so that, beyond some range from the transmitter, the transmitted signal is buried in the natural background noise. The process gain (TW) controls the reduction in detectable range vis-à-vis a "clear" signal.

### Use in Radar

It is usual to view radar applications as a variation on communication; that is, the return waveforms are known except with respect to noise, Doppler shift, and delay. Spectrum spreading is applicable to both cw and pulse radars. The major differentiation is in the choice of cross-correlation or matched-filter detector. The TW product is the key performance parameter, but the covariance function properties must frequently be determined to resolve Doppler shifts as well as range delays.

## CLASSIFICATION OF SPREAD-SPECTRUM SIGNALS

Spread-spectrum signals can be classified on the basis of their spectral occupancy versus time characteristics, as sketched in Fig. 12.4.1. Direct-sequence (DS) and pseudo-noise (PN) waveforms provide continuous full

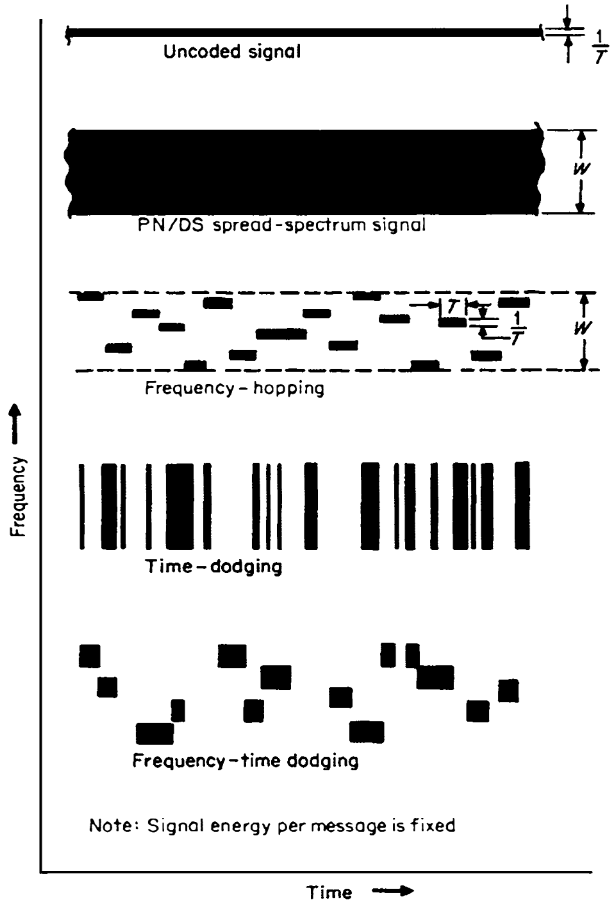


**FIGURE 12.4.1** Spectral occupancy vs. time characteristics of spread-spectrum signals.

coverage, while frequency-hopping (FH), time-dodging, and frequency-time dodging (F-TD), fill the frequency-time plane only in a long-term averaging sense.

DS waveforms are pseudo-random digital streams generated by digital techniques and transmitted without significant spectral filtering. If heavy filtering is used, the signal amplitude statistics become quite noiselike, and this is called a *PN waveform.* In either case correlation detection is generally used because the waveform is dimensionally too large to implement a practical matched filter, and the sequence generator is relatively simple and capable of changing codes.

In FH schemes the spectrum is divided into subchannels spaced orthogonally at $1/T$ separations. One or more (e.g., two for FSK) are selected by pseudo-random techniques for each data bit. In time-dodging schemes the signal burst time is controlled by pulse repetition methods, while F-TD combines both selections. In each case a jammer must either jam the total spectrum continuously or accept a much lower effectiveness (approaching 1/TW). Frequency-hopped signals can be generated using SAW chirp devices.

## CORRELATION-DETECTION SYSTEMS

The basic components of a typical DS type of link are shown in Fig. 12.4.2. The data are used to select the appropriate waveform, which is shifted to the desired rf spectrum by suppressed-carrier frequency-conversion

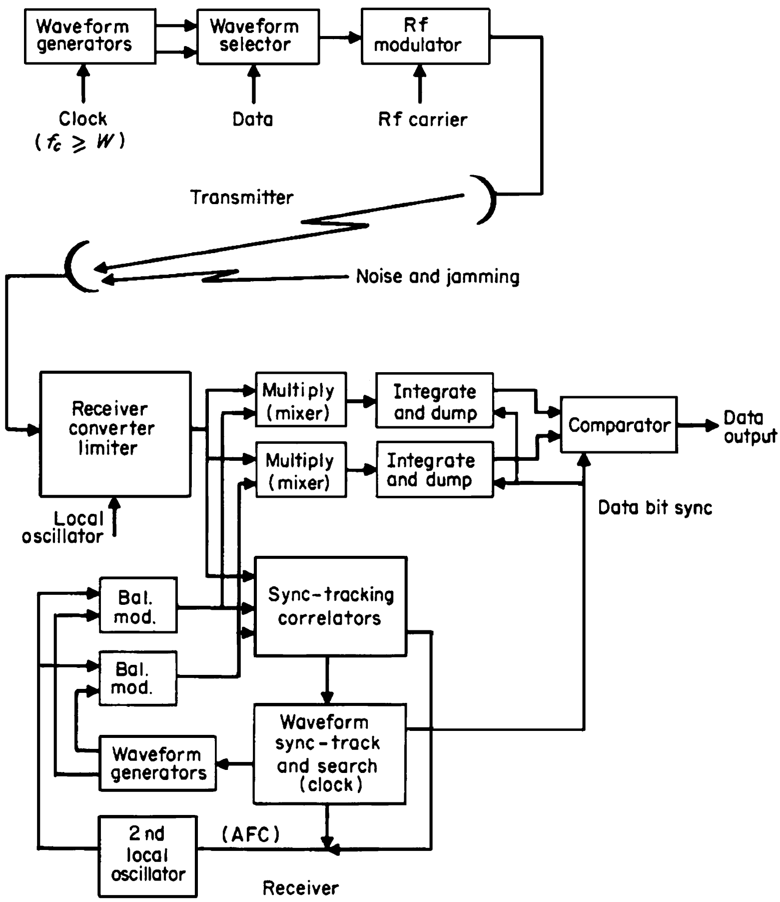


**FIGURE 12.4.2** Direct-sequence link for spread-spectrum system.

techniques, and transmitted. At the receiver identical locally generated waveforms multiply with the incoming signal. The stored reference signals are often modulated onto a local oscillator, and the incoming rf may be converted to an intermediate frequency, usually with rf or i.f. limiters.

The mixing detectors are followed by linear integrate-and-dump filters, with a "greatest of" decision at the end of each period. The integrator is either a low-pass or bandpass quenchable narrowband filter. Digital techniques are increasingly being used.

*Synchronization* is a major design and operational problem. Given a priori knowledge of the transmitted sequences, the receiver must bring its stored reference timing to within $\pm 1/(2W)$ of the width of the received signal and hold it at that value. In a system having a 19-stage *pn* generator, a 1-MHz *pn* clock, and a 1-kHz data rate, the width of the correlation function is $\pm^1/_2$ $\mu$s, repeating $^1/_2$ s separations, corresponding to 524.287 clock periods. In the worst case, it would be necessary to inspect each sequence position for 1 ms; that is, 524 s would be required to acquire sync. If oscillator tolerances and/or Doppler lead to frequency uncertainties equal to or greater than the 1-kHz data rate, then parallel receivers or multiple searches are required.

Ways to reduce the sync acquisition time include using jointly available timing references to start the *pn* generators, using shorter sequences for acquisition only; "clear" sync triggers; and paralleling detectors. Titsworth (see bibliography) discusses composite sequences which allow acquiring each component sequentially, searching $N_1 + N_2 + N_3$ delays, while the composite sequence has length $N_1N_2N_3$. These methods have advantages for space-vehicle ranging applications but have reduced security to jamming.

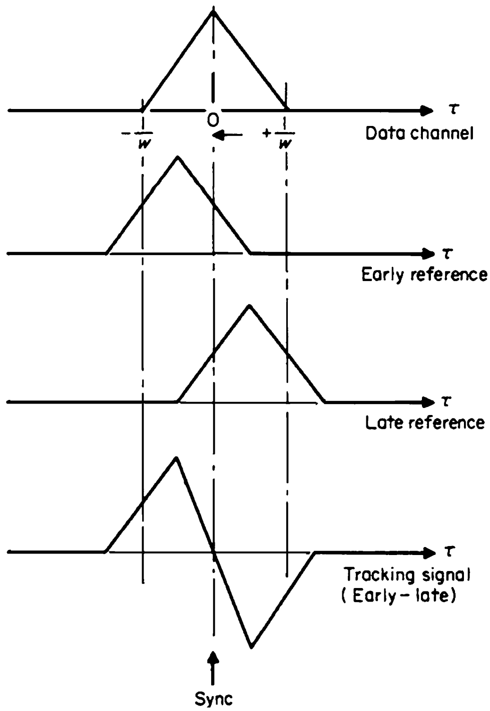


**FIGURE 12.4.3** Sync tracking by early-late correlators.

Sync tracking is usually performed by measuring the correlation at early and late times, $\pm\tau$, where $\tau \leq 1/W$, as shown in Fig. 12.4.3. Subtracting the two provides a useful time discrimination function, which controls the *pn* clock. The displaced values can be obtained by two tracking-loop correlators or by time-sharing a single unit. "Dithering" the reference signal to the signal correlator can also be used, but with performance compromises.

The tracking function can also be obtained by using the time derivative of one of the inputs

$$\frac{d}{d\tau}\,\varphi_{XY}(\tau) = \frac{dX(t)}{dt}\cdot Y(t+\tau) \tag{1}$$

A third approach has been to add by mod 2 methods the clock to the transmitted *pn* waveform. The spectral envelope is altered, but very accurate peak tracking can be accomplished by phase locking to the recovered clock.

## LIMITERS IN SPREAD-SPECTRUM RECEIVERS

Limiters are frequently used in spread-spectrum receivers to avoid overload saturation effects, such as circuit recovery time, and the incidental phase modulation. In the usual low-input signal-noise range, the limiter tends to normalize the output noise level, which simplifies the decision circuit design. In repeater applications (e.g., satellite), a limiter is desirable to allow the transmitter to be fully modulated regardless of the input-signal strength. When automatic gain control (AGC) is used, the receiver is highly vulnerable to pulse jamming, while the limiter causes a slight reduction of the instantaneous signal-to-jamming ratio and a proportional reduction of transmitter power allocated to the desired signal.

## DELTIC-AIDED SEARCH

The sync search can be accelerated by use of deltic-aided (delay-line time compression) circuits if logic speeds permit. The basic deltic consists of a recirculating shift register (or a delay line) which stores $M$ samples, as shown in Fig. 12.4.4. The incoming spread-spectrum signal must be sampled at a rate above $W$ ($W$ = bandwidth). During each intersample period the shift register is clocked through $M + 1$ shifts before accepting the next sample. If $M \geq 2W$, a signal period at least equal to the data integration period is stored and is read out at $M$ different delays during each period $T$, permitting many high-speed correlations against a similarly accelerated (but not time-advancing) reference.

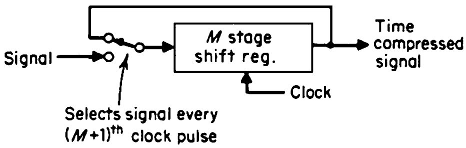


**FIGURE 12.4.4** Delay-line time compression (deltic) configuration.

For a serial-deltic and shift-register delay line the clock rate is at least $4TW^2$. Using a deltic with $K$ parallel interleaved delay lines, the internal delay lines are clocked at $4TW^2/K^2$ and the demultiplexed output has a bit rate of $4TW^2/K$, providing only $M/K$ discrete delays. This technique is device-limited to moderate signal bandwidths, primarily in the acoustic range up to about 10 kHz.

## WAVEFORMS

The desired properties of a spread-spectrum signal include:

- An autocorrelation function, which is unity at $\tau = 0$ and zero elsewhere
- A zero cross-correlation coefficient with noise and other signals
- A large library of orthogonal waveforms

### Maximal-Length Linear Sequences

A widely used class of waveforms is the maximal-length sequence (MLS) generated by a tapped refed shift register, as shown in Fig. 12.4.5*a* and as one-tap unit in Fig. 12.4.5*b*. The mod 2 half adder ($\oplus$) and EXCLUSIVE-OR logic gate are identical for 1-bit binary signals.

If analog levels +1 and –1 are substituted, respectively, for 0 and 1 logic levels, the circuit is observed to function as a l-bit multiplier.

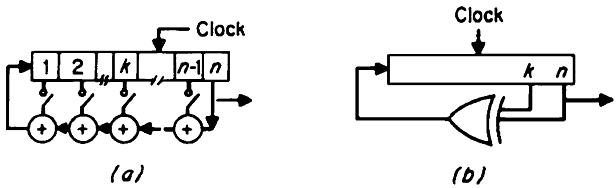


**FIGURE 12.4.5** Maximal-length-sequence (MLS) system.

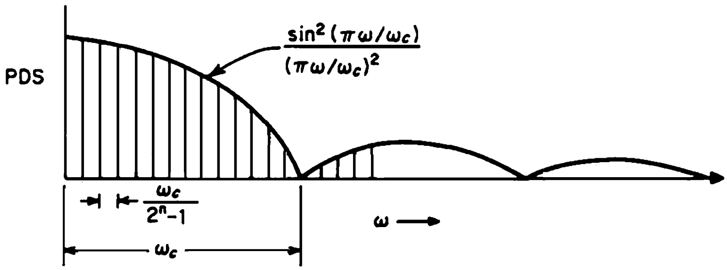


**FIGURE 12.4.6** Spectrum of MLS system.

Pertinent properties of the MLS are as follows. Its length, for an $n$-stage shift register, is $2^n - 1$ bits. During $2^n - 1$ successive clock pulses, all $n$-bit binary numbers (except all zeros) will have been present. The autocorrelation function is unity at $\tau = 0$, and at each $2^n - 1$ clock pulses displacement, and $1/(2n - 1)$ at all other displacements. This assumes that the sequences repeat cyclically, i.e., the last bit is closed onto the first. The autocorrelation function of a single (noncyclic) MLS shows significant time side lobes. Titsworth (see bibliography) has analyzed the self-noise of incomplete integration over $p$ chips, obtaining for MLSs,

$$\sigma^2(t) = (p - t)\,(p^2 - 1)/p^3 t \tag{2}$$

which approaches $1/t$ for the usual case of $p >> t$. Since $t \approx TW$, the self-noise component is usually negligible.

Another self-noise component is frequently present owing to amplitude and dispersion differences, caused by filtering, propagation effects, and circuit nonlinearities. In addition to intentional clipping, the correlation multiplier is frequently a balanced modulator, which is linear only to the smaller signal, unless deliberately operated in a bilinear range. The power spectrum is shown in Fig. 12.4.6. The envelope has a $(\sin^2 X)/X^2$ shape ($X = \pi\omega/\omega_{clock}$), while the individual lines are separated by $\omega_{clock}/(2^n - 1)$.

An upper bound on the number of MLS for an $n$-stage shift register is given in terms of the Euler $\phi$ function:

$$N_u = \phi(2^n - 1)/n \le 2^{(n - \log 2n)} \tag{3}$$

where $\phi(k)$ is the number of positive integers less than $k$, including 1, which are relatively prime to $k$.