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An Introduction to Spread Spectrum Systems

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Abstract

This report is a tutorial introduction to spread spectrum communication systems, and it should be accessible for a reader who has basic knowledge of digital modulation and channel coding. The authors have chosen to present the more advanced topics in the report from a signal space and channel coding point of view. While other approaches are certainly possible, the authors feel that the chosen approach is the most beneficial for the communications engineer. Since the original applications for spread spectrum systems were military, we will introduce spread spectrum links as a means of overcoming intentional jamming. The reader who dislikes military applications may be comforted by the fact that the problem of communicating in the presence of jamming is very much akin to the problem of communicating over fading channels. Hence, by finding out how to defeat jamming by spread spectrum will also reveal how to overcome fading. Today, spread spectrum links are also used in many civilian systems to overcome nonintentional jamming (or interference), and the report is concluded with an overview of current commercial spread spectrum systems.

The authors welcome comments and corrections to the material. Please send these to Erik Ström at `erik.strom@s2.chalmers.se`.

Contents

1 Introduction

Spread spectrum systems is a class of (primarily) wireless digital communication systems specifically designed to overcome a jamming situation, i.e., when an adversary intends to disrupt the communication. To disrupt the communication, the adversary needs to do two things, (a) to detect that a transmission is taking place and (b) to transmit a jamming signal which is designed to confuse the receiver. A spread spectrum system is therefore designed to make these tasks as difficult as possible. Firstly, the transmitted signal should be difficult to detect by an adversary, i.e., the signal should have a low probability of intercept (LPI). Secondly, the signal should be difficult to disturb with a jamming signal, i.e., the transmitted signal should possess an anti-jamming (AJ) property.

Clearly, the intentional jamming situation is most common in a military context, and spread spectrum systems were originally developed specifically for military applications. (For an interesting review of the history of the spread spectrum in the West in general and the US in particular, we can recommend the papers [1, 2, 3] or chapter 2 of [4].) However, in later years, spread spectrum systems have been introduced in many commercial applications that require good anti-jamming properties. An example of commercial spread spectrum systems are systems that are designed to be used in so-called unlicensensed bands, such as the Industry, Scientific, Medical (ISM) band around 2.4 GHz. Typical applications are here cordless telephones, wireless LANs, and cable replacement systems as Bluetooth. Since the band is unlicensed, there is no central control over the radio resources, and the systems have to function even in the presence of severe interference from other communication systems and other electrical and electronic equipment (e.g., microwave ovens, radars, etc.). Here the jamming is not intentional, but the interference may nevertheless be enough to disrupt the communication for non-spread spectrum systems.

Code-division multiple access systems (CDMA systems) use spread spectrum techniques to provide communication to several concurrent users. CDMA is used in one second generation (IS-95) and several third generation wireless cellular systems (e.g., cdma2000 and WCDMA). One advantage of using jamming-resistant signals in these applications is that the radio resource management (primarily the channel allocation to the active users) is significantly reduced.

The name spread spectrum stems from the fact that the transmitted signals occupies a much wider frequency band than what is necessary. This enables the transmitter to hide its signal in a large bandwidth. There are many different ways to use the bandwidth. The most common ones are called direct-sequence (DS) and frequency-hopping (FH) spread spectrum (SS). In FH-SS, the transmitter changes the carrier frequency of the relatively narrowband transmitted signal in a fashion which appears random to the jammer. At any given time, only a small fraction of the available bandwidth is used, and exactly which fraction is made a secret for the jammer. The jammer is therefore uncertain where in the system bandwidth the signal is being transmitted, and it is difficult for the jammer to detect and disturb the transmitted signal. In DS-SS, the power of the transmitted signal is spread over the entire system bandwidth in a way that looks random for the jammer.

Again, this makes the signal hard to detect and to jam. Several other spread spectrum strategies are available; however, the clear majority of the implemented systems are either frequency-hopping or direct-sequence (or hybrids of these basic schemes).

The bandwidth necessary for the transmission of a digital communications signal is determined by the data rate, R_b, (measured in the number of information bits transmitted per second) and the chosen modulation format. For binary passband modulation (suitable for wireless transmission), the minimum required bandwidth is approximately $W_{\min} = R_b$ Hz. If we denote the actual bandwidth of the transmitted signal by W_{ss}, then for a spread spectrum system $W_{ss} \gg R_b$. The spectral efficiency of the spread spectrum communication link is R_b/W_{ss} bits/second/Hz. By definition, the spectral efficiency of a spread spectrum system is very low. This seems to render spread spectrum techniques useless for systems that need to use spectrum efficiently. However, this is not necessarily the case since several users using spread spectrum signals can share the same bandwidth (at the same time), and the system's spectral efficiency (measured in the total number of information bits transmitted per second) may be still be very good, even if the individual links have low spectral efficiencies.

The literature in the field of spread spectrum communications is quite voluminous and ranges from text books to specialized conference and journal papers. Among the available books, we would like to especially mention the text by Simon, Omura, Scholtz, and Levitt [4] which covers quite a lot of the classical spread spectrum techniques and which has inspired this presentation to a large degree. The books by Peterson, Ziemer, and Borth [5] and by Dixon [6], which cover more of the current commercial applications, are also recommended. The reference lists of the abovementioned books contain several thousand entries. Among the tutorial-style papers that are available in the literature, we would like to especially mention the 1982 paper by Pickholtz, Schilling, and Milstein [7].

In this report, we will concentrate on describing the anti-jamming property of spread spectrum systems. Moreover, we will limit our discussion to the two main forms of spread spectrum: direct-sequence and frequency-hopping spread spectrum. We will try to cover DS spread spectrum in some detail, but many issues will be skimmed and the FH discussion will be quite brief.

2 A Simple Jamming Game

Let us study a simple example to illustrate the basic concepts and ideas of spread spectrum. Let the average received communication signal power and received jamming signal power be denoted by S and J, respectively. We ignore any other interference, such as thermal noise, at this stage. The basic question is "How can we communicate reliably even when $J \gg S$?".

Actually, we know the answer from basic communication theory. We know that we can communicate over a channel disturbed by additive white Gaussian noise (an AWGN channel). White noise has infinite power, but since the power is spread over an infinite number of signal space dimensions (or infinite bandwidth), the power per signal space dimension is finite. Hence, by concentrating the transmitter power to a finite-dimensional signal space, we can gain a power advantage over the noise.

The same idea is used in a jamming situation. However, we must make the choice of signal space dimensions used for transmission a secret for the jammer. Otherwise, the jammer can concentrate its power to the same dimensions, and nothing is gained. This implies that we need to hide the transmitted signal in a space with many more dimensions than what is needed for the transmitted signal.

From the work by Landau, Pollak, Slepian and others [8, 9, 10, 11], we know that the dimensionality of a signal space depends on duration and bandwidth of the signals in the space. Suppose the spread spectrum bandwidth (or system bandwidth) is W_{ss}. That is, the transmitted signal must reside in a frequency band of width W_{ss} Hz. If the data rate (number of information bits transmitted per second) is $R_b = 1/T_b$, then the transmission of a (long) packet of P information bits will take approximately T_p seconds, where

$$T_p = PT_b = \frac{P}{R_b}.$$

It has been shown that the set of all signals that are (essentially) time-limited to T_p seconds and (essentially) bandlimited to W_{ss} Hz spans a signal space of (approximately) $N_p = 2W_{ss}T_p$ dimensions. Hence, the number of dimensions available for the transmission of one information bit is

$$N_d = \frac{N_p}{P} = 2\frac{W_{ss}}{R_b}.$$

A characteristic of a spread spectrum system is that ratio W_{ss}/R_b is very large. As a matter of the fact, the larger the ratio the more resistant to jamming the system can be made. The number W_{ss}/R_b is often called the processing gain; however, the reader should be cautioned that processing gain is not a well-defined term. Indeed, in the literature many other definitions exist.

If the jammer has no idea which subset of the N_d dimensions that is used for the transmission, it may decide to spread its power equally over all dimensions. The jamming signal can then be viewed as a noise waveform with flat power spectrum over the system bandwidth, see Figure 1. The spectral height of the jamming signal is denoted $N_J/2$, where

$$J = 2W_{ss}\frac{N_J}{2} = W_{ss}N_J.$$

This type of jamming is called broadband noise jamming, and from now on we will assume that the noise is Gaussian. As we will see later, broadband noise jamming is perhaps the most benign form of jamming, and much effort is put into the design of spread spectrum systems to force the jammer into this jamming strategy.

To quantify the effect of broadband noise jamming, let us consider the effect on a binary antipodal modulation format, such as binary phase-shift keying (BPSK). The bit error probability is (assuming perfect synchronization and ideal coherent detection)

$$P_b = Q\left(\sqrt{2\frac{E_b}{N_J}}\right),$$

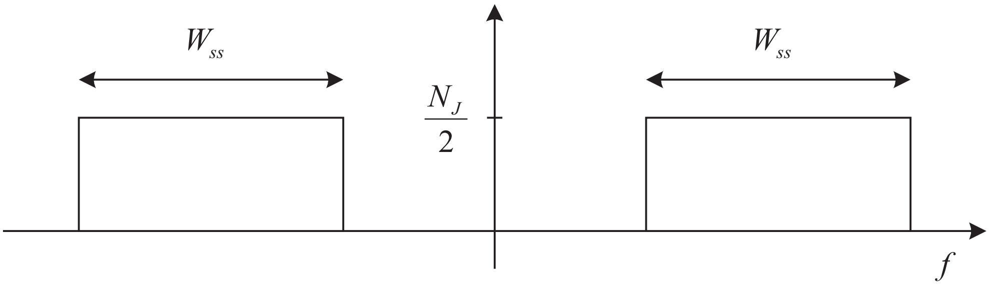

Figure 1: Power spectral density of a broadband (nonpulsed) noise jammer.

where $E_b = ST_b = S/R_b$ is the received energy per information bit and

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{x} e^{-t^2/2}\, dt.$$

As seen from Figure 2, the bit error probability is decreasing exponentially with E_b/N_J, and since

$$\frac{E_b}{N_J} = \frac{S/R_b}{J/W_{ss}} = \frac{S}{J}\frac{W_{ss}}{R_b},$$

we can reach any bit error probability for any jammer-to-signal power ratio, J/S, by making the processing gain large enough. However, we should remember that we have reached this conclusion under idealized circumstances. For instance, if the jamming causes the synchronization or front-end electronics or processing to fail, then this will effectively disrupt the communication.

We are now ready to characterize a spread spectrum system as a system for which the follwing holds.

1. The system (spread spectrum) bandwidth is (much) larger than the information bit rate.
2. The bandwidth expansion achieved by a process which is independent of the transmitted data.

The second requirement has to do with the fact that the choice of dimensions must be made a secret for the jammer. This definition may not be completely unambiguous and has been criticized by, e.g., Massey who in [12] offers a different definition.

For any given type of spread spectrum system, there will be a worst-case jamming strategy (worst-case from the communicator's point of view). However, there is no jamming strategy which is worst-case for all conceivable spread spectrum formats. The vice versa is also true; there exist no spread spectrum format which is optimum for all jamming strategies. Hence, many types of spread spectrum systems have been proposed and are used. As mentioned earlier, the most common types are direct-sequence spread spectrum (DS-SS), frequency-hopping spread spectrum (FH-SS), and hybrids of these.

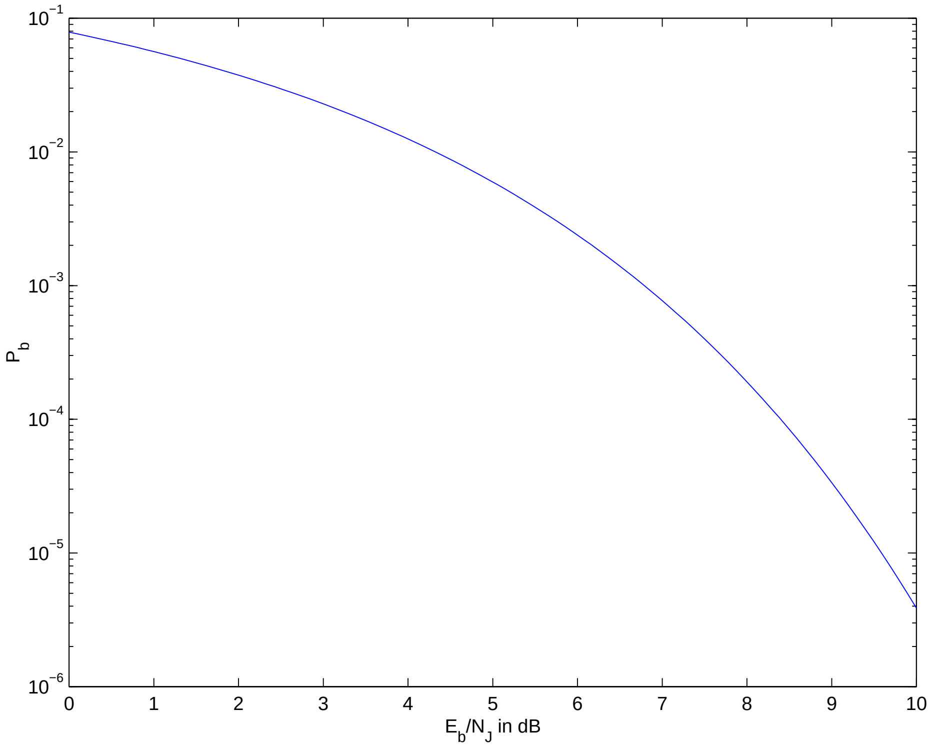

Figure 2: Bit error probability for BPSK-modulated DS-SS and broadband (non-pulsed) noise jamming. Coherent detection and perfect synchronization is assumed.

3 Direct-Sequence Spread Spectrum

3.1 Transmitted Signal and Receiver Structure

We will present two receiver structures and two ways to produce the transmitted signal for a DS-SS system. The first approach is to view the bandwidth expansion as result of repetition coding followed by a randomizing scrambling. The second approach views DS-SS as regular BPSK modulation where the bandwidth expansion is a result of (randomly) choosing a wideband pulse shape. The two models are equivalent in that the transmitted signal is the same and that the receivers have the same performance. Nevertheless, it is worthwhile to be aware of both models since they both add insight to the problem.

Let us first consider an ordinary uncoded BPSK system. The transmitted signal can be written as

$$s(t) = \sqrt{2E_b}\cos(2\pi f_c t)\sum_{n=-\infty}^{\infty} b[n]p(t-nT_b),$$

where E_b is the energy per information bit, f_c is the carrier frequency, $b[n] \in \{\pm 1\}$ is the nth information bit, $p(t)$ is the unit-energy pulse shape, and the data rate is $R_b = 1/T_b$. The bandwidth of $s(t)$ is determined by the pulse shape and data rate (assuming that the data sequence is white). For instance, if the pulse has a root-raised cosine spectrum with roll-off factor α, the bandwidth of the transmitted signal is $(1+\alpha)R_b \approx R_b$ for small α.

If we instead use a rectangular pulse shape

$$p(t) = \begin{cases} \frac{1}{\sqrt{T_b}}, & 0 \leq t < T_b \\ 0, & \text{otherwise} \end{cases},$$

the transmitted signal will no longer be strictly bandlimited. However, if we use the null-to-null bandwidth measure, the bandwidth is $2R_b$ Hz, see Figure 3. It can be shown that more than 90% of the total power of the transmitted signal is contained in the null-to-null bandwidth.

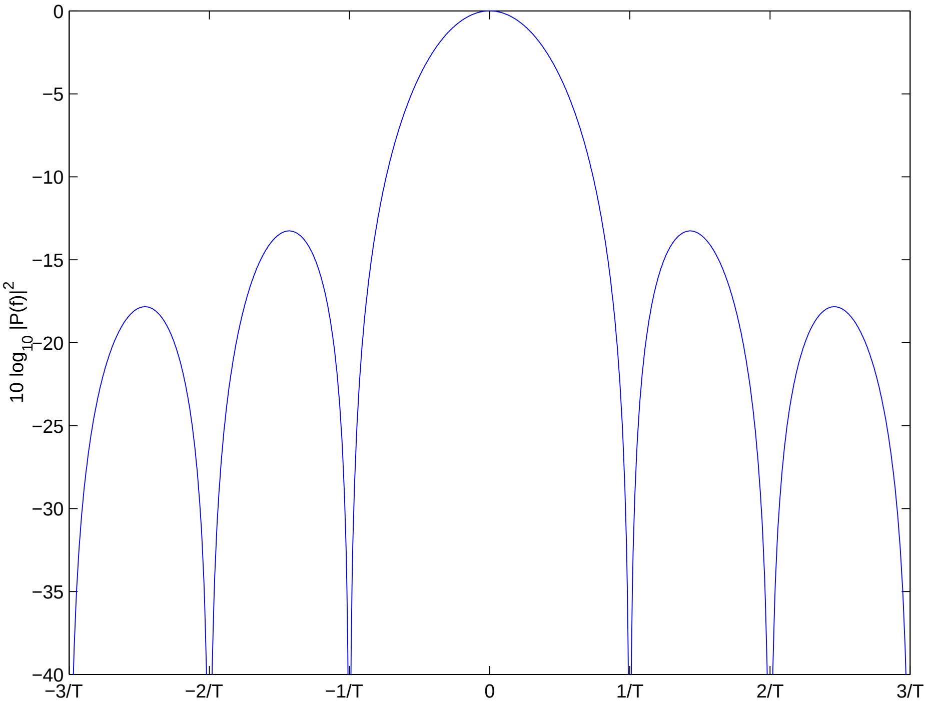

Figure 3: Plot of the magnitude square of the Fourier transform of a rectangular pulse of width T seconds (in dB-scale). Clearly, the null-to-null bandwidth is $1/T$ for the baseband signal and $2/T$ for a passband signal.

We can achieve an increase of the bandwidth to $W_{ss} = N2R_b$ by (a) repeating every information bit N times and (b) randomizing the repeated information bits with a scrambling code (or spreading code) $c[m]$. The procedure is illustrated in Figure 4. Since the bandwidth is increased with a factor of N, the number N is known as the spreading factor.

The rate after the repetition encoder is $NR_b = N/T_b$ and the transmitted (channel) bits are

$$d[m] = c[m]b\big[\lfloor m/N \rfloor\big],$$

where $\lfloor \cdot \rfloor$ denotes the floor function (rounds down to closest integer). The channel bits are also known as chips, and the channel bit rate is called the chip rate $1/T_c = NR_b$. The scrambling code should appear random to the jammer, but

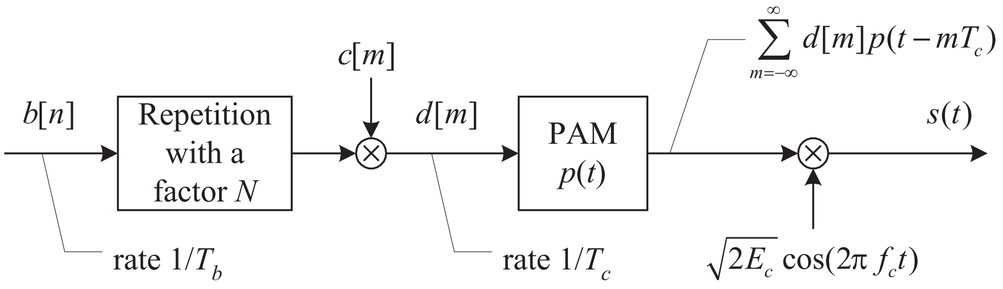
$b[n]$
Repetition with a factor N
rate $1/T_b$
$c[m]$
$d[m]$
rate $1/T_c$
PAM $p(t)$
$\sum_{m=-\infty}^{\infty} d[m]p(t-mT_c)$
$s(t)$
$\sqrt{2E_c}\cos(2\pi f_c t)$

Figure 4: Modulator for DS-SS with BPSK modulation with an arbitrary chip waveform $p(t)$. (PAM stands for pulse amplitude modulation.)

must also be reproducible by the receiver. Pseudo-random noise sequences fits these requirements and are often used as scrambling codes [13].

The transmitted signal of a DS-SS system with BPSK modulation is

$$s(t) = \sqrt{2E_c}\cos(2\pi f_c t) \sum_{m=-\infty}^{\infty} d[m]p(t - mT_c),$$

where $E_c = E_b/N$ is the energy per chip and $p(t)$ is the chip pulse shape (assumed to have unit energy).

This arrangement now fulfills the requirement for being a spread spectrum system: the bandwidth of the transmitted signal is larger than R_b and the bandwidth expansion is done by the scrambling code, which is independent of the transmitted data. For the special case of rectangular chip pulses, we can rewrite the transmitted signal as

$$s(t) = \sqrt{\frac{2E_b}{T_b}}\cos(2\pi f_c t)b(t)c(t)$$

where the data signal $b(t)$ is defined as

$$b(t) = \sum_{n=-\infty}^{\infty} b[n]\Pi_{T_b}(t - nT_b)$$

and the scrambling waveform is defined as

$$c(t) = \sum_{m=-\infty}^{\infty} c[m]\Pi_{T_c}(t - mT_c),$$

and $\Pi_T(t)$ denotes a unit-amplitude rectangular pulse of duration T seconds. A block diagram of the modulator is found in Figure 5.

We observe that the scrambling waveform $c(t)$ is found by pulse amplitude modulating the scrambling code $c[m]$ with the chip pulse shape. The scrambling code is designed to behave as discrete-time white noise, and we can model the chips as being independent and equally likely ± 1. Hence, the bandwidth of $c(t)$ is determined by the bandwidth of $\Pi_{T_c}(t)$ which has null-to-null bandwidth $1/T_c$ Hz

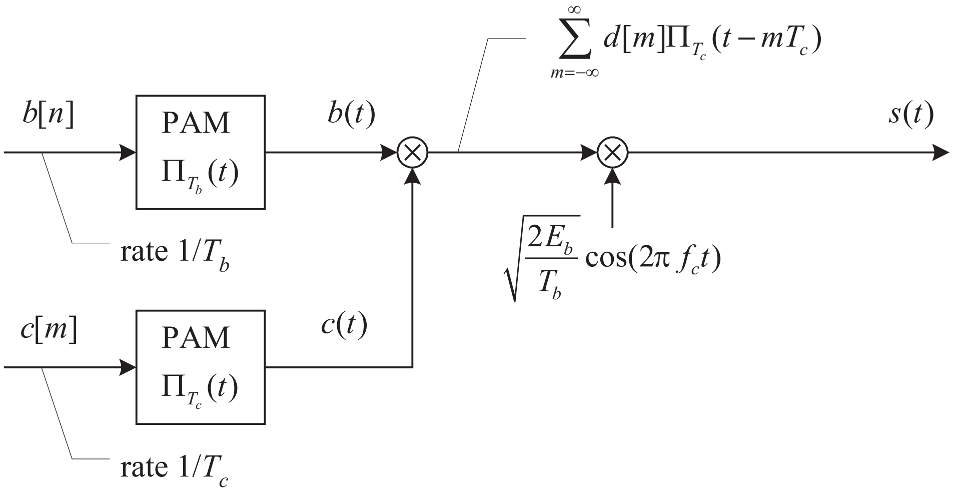
$\sum_{m=-\infty}^{\infty} d[m]\Pi_{T_c}(t-mT_c)$
$b[n]$
PAM
$\Pi_{T_b}(t)$
$b(t)$
$s(t)$
rate $1/T_b$
$\sqrt{\frac{2E_b}{T_b}}\cos(2\pi f_c t)$
$c[m]$
PAM
$\Pi_{T_c}(t)$
$c(t)$
rate $1/T_c$

Figure 5: Modulator for DS-SS with BPSK modulation and rectangular chip waveforms.

(recall that $c(t)$ is a baseband signal and its power spectral density is proportional to the magnitude squared of the Fourier transform of $\Pi_{T_c}(t)$, see Figure 3). A similar argument reveals that $b(t)$ has bandwidth $1/T_b$, and that the product $c(t)b(t)$ has the same bandwidth as $c(t)$, i.e., $1/T_c = N/T_b = NR_b$ Hz. Finally, after modulation with the carrier, the transmitted waveform will have a null-to-null bandwidth of $2/T_c = 2N/T_b = 2NR_b$ Hz.

It is now clear that by multiplying $b(t)$ with $c(t)$, we obtain a bandwidth expansion, or spreading, of a factor of $T_b/T_c = N$. This is why $c(t)$ is also called the spreading waveform and $c[m]$ is called the spreading code.

We have presented two views on how the DS-SS transmitted signal is produced. The first one says that the bandwidth expansion is achieved through channel coding (repetition coding), and the second says that the bandwidth expansion is done through a change of the pulse shape. We can also form the receiver from these two perspectives. The receiver from the channel coding perspective is a demodulator followed by a channel decoder, see Figure 6. Hence, the receiver consists of a down-mixing stage (multiplication by a locally generated carrier) followed by a filter which is matched to the chip pulse shape and sampled at chip rate. The sampled output of the chip matched filter is descrambled by multiplying with a locally generated copy of the scrambling code. The descrambled chips are fed to the repetition code decoder which outputs the information bit decisions.

The receiver from a modulation perspective is just a down-mixing stage followed by a filter which is matched to consecutive T_b-segments of $c(t)$ (a so-called code matched filter). A block diagram of the receiver where the matched filter is implemented with a correlator is found in Figure 7.

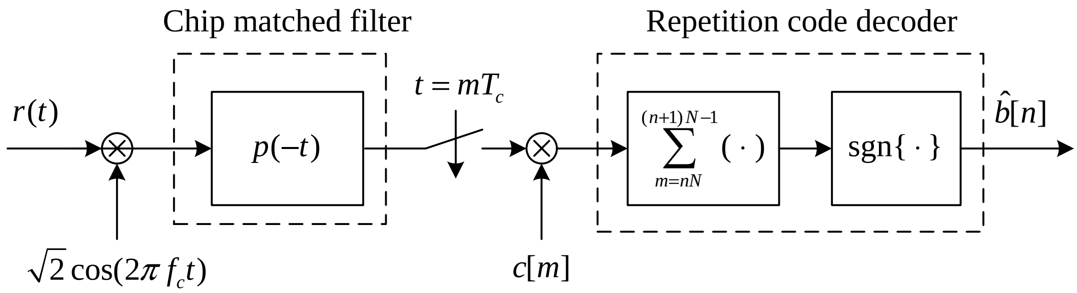

Figure 6: Demodulator for DS-SS with BPSK modulation with an arbitrary chip waveform $p(t)$.)

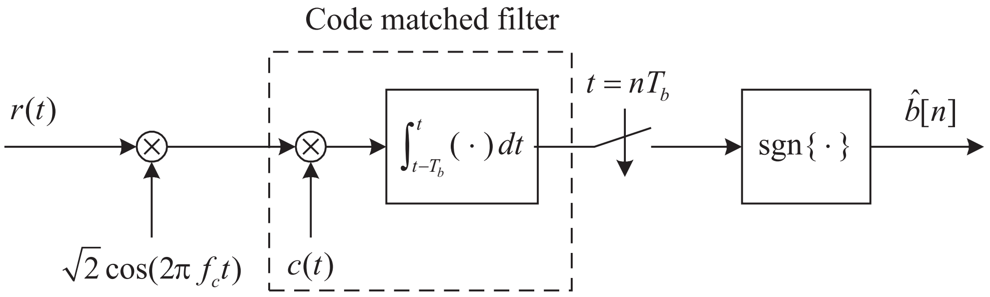

Figure 7: Demodulator for DS-SS with BPSK modulation with rectangular chip waveform.

3.2 DS-SS and Broadband Continuous Noise Jamming

For rectangular chip pulses, the receivers presented in Figures 6 and 7 are equivalent and optimum, in the sense that the bit error probability is minimized if the received signal is equal to the transmitted signal plus white Gaussian noise. Hence, if the jammer waveform is Gaussian noise that is spectrally white over the system bandwidth and if we ignore any other interference (such as thermal noise), the bit error probability is

$$P_b = Q\left(\sqrt{2\frac{E_b}{N_J}}\right). \tag{1}$$

If we assume that the channel also adds white Gaussian noise with power spectral density $N_0/2$, then the resulting bit error probability is

$$P_b = Q\left(\sqrt{2\frac{E_b}{N_J + N_0}}\right),$$

which is illustrated in Figure 8.

Note that the processing gain only affects $N_J = JR_b/W_{ss}$. Hence, the bandwidth expansion does not help at all to combat the white channel noise. However, by replacing the repetition code with a better channel code, we can combat both the jamming and the channel noise more efficiently.

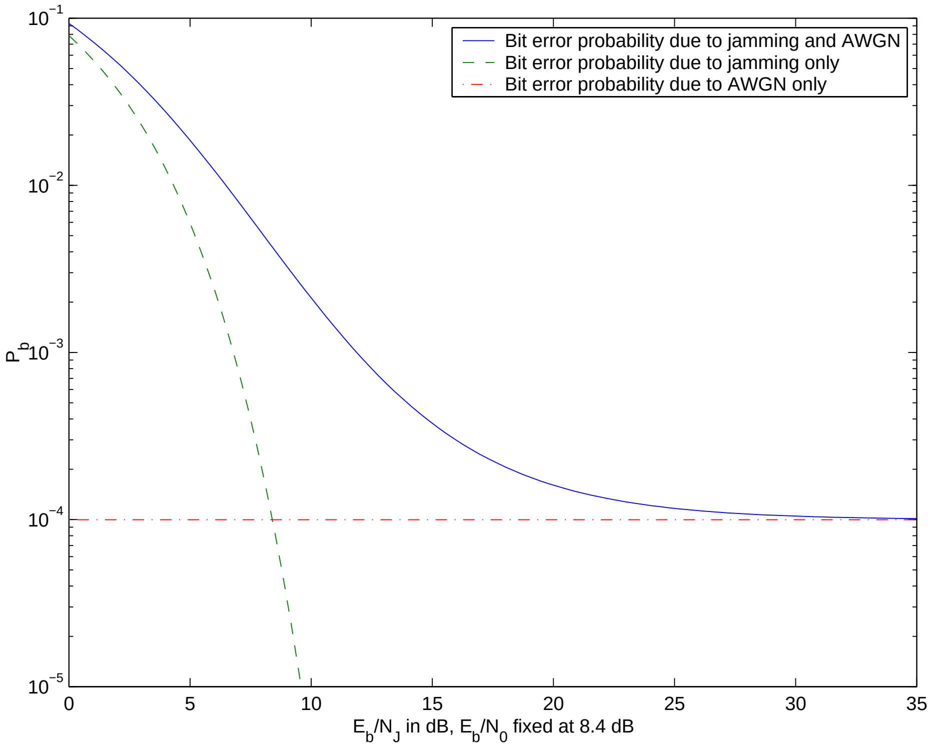

Figure 8: Bit error probability for DS-SS with BPSK modulation over an AWGN channel with $E_b/N_0 = 8.4$ dB and with broadband noise jamming of varying power.

3.3 DS-SS and Narrowband Jamming

DS-SS is also effective against narrowband jamming and interference. This is perhaps easiest seen from the modulation point of view.

Let us study a system with rectangular chip waveforms and transmitter and receiver as in Figures 5 and 7, respectively. Furthermore, suppose the jamming signal is a pure cosine at the carrier frequency with power J and phase θ, i.e.,

$$j(t) = \sqrt{2J}\cos(2\pi f_c t + \theta).$$

The contribution from the jammer to the input to the integrator block in Figure 7 is

$$\begin{aligned} j(t)\sqrt{2}\cos(2\pi f_c t)c(t) &= \sqrt{J}c(t)2\cos(2\pi f_c t + \theta)\cos(2\pi f_c t) \\ &= \sqrt{J}c(t)[\cos(\theta) + \cos(4\pi f_c t + \theta)] \\ &= \sqrt{J}c(t)\cos(\theta) + \sqrt{J}c(t)\cos(4\pi f_c t + \theta). \end{aligned}$$

The second term in the equation above will have its power centered around twice the carrier frequency, and since the integrator is a lowpass filter, this term will be suppressed almost completely. The first term has its power centered around DC and is spread over the entire system bandwidth (maximum frequency approximately $1/T_c$ Hz). Again, since the integrator is a lowpass filter with a cut-off frequency of approximately $1/T_b$ Hz, only a fraction of the jammer power will remain after the integrator.

The same type of argument can be made for a more general narrowband jamming signal. The receiver will spread the power of the jamming signal to span approximately the entire system bandwidth, and the integrator will lowpass filter the spread jamming signal. Hence, only a small fraction of the jammer power will effect the decisions on the information bits. However, the desired signal component will be despread by the receiver. The desired signal component at the input to the integrator in Figure 7 is

$$c^2(t)b(t)\sqrt{\frac{E_b}{T_b}}2\cos^2(2\pi f_c t) = b(t)\sqrt{\frac{E_b}{T_b}} + b(t)\sqrt{\frac{E_b}{T_b}}\cos(4\pi f_c t),$$

and the integrator serves as a matched filter for the data signal $b(t)$. The spreading of the jammer signal and despreading of the desired signal operation is conceptually illustrated in Figure 9.

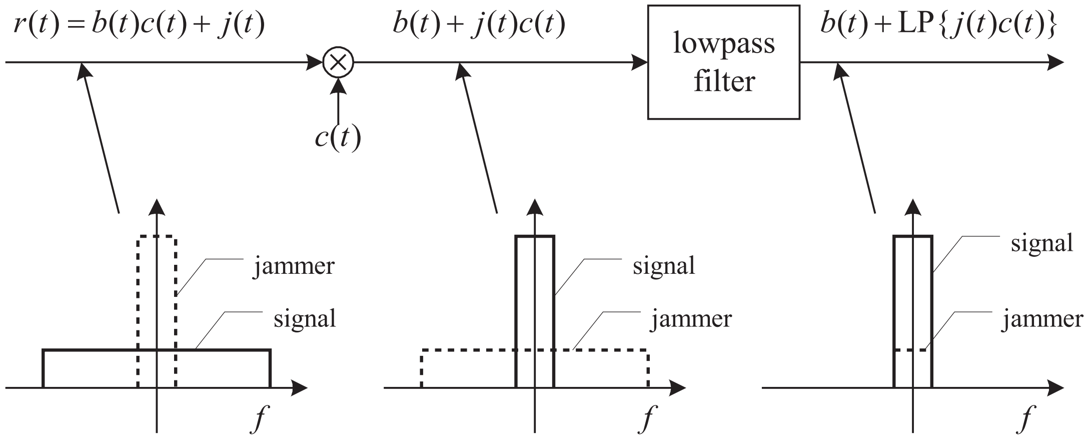

Figure 9: Despreading operation in the presence of narrowband jamming. The plots show the power spectral densities at various points in the despreading circuit (double frequency terms are neglected in this figure).

3.4 Pulsed Jamming

One very effective jamming strategy for DS-SS is a broadband pulsed noise jammer. A broadband pulsed noise jammer transmits noise whose power is spread over the entire system bandwidth. However, the transmission is only on for a fraction ρ of the time (i.e., ρ is the duty cycle of the jammer transmission and $0 < \rho \leq 1$). This allows the jammer to transmit with a power of J/ρ when it is transmitting (remember that J is the average received jammer power), and the equivalent spectral height of the noise is $N_J/2\rho$.

To make a simple analysis of the impact of a pulsed jammer we start by assuming that the jammer affects an integer number of information bits. That is, during the transmission of a certain information bit, the jammer is either on (with probability ρ) or off (with probability $1 - \rho$). Furthermore, if we assume that the jammer waveform is Gaussian noise and ignore all other noise and interference, the bit error probability for a DS-SS system with BPSK modulation (coherent detection and perfect synchronization) is

$$P_b = (1-\rho)\times 0 + \rho Q\left(\sqrt{\frac{2E_b}{N_J}\rho}\right) = \rho Q\left(\sqrt{\frac{2E_b}{N_J}\rho}\right).$$

For a fixed E_b/N_J (e.g., for a fixed processing gain and fixed average received jammer and signal powers), the worst case jammer duty cycle can be found to be

$$\rho_{wc} = \begin{cases} \frac{0.709}{E_b/N_J}, & \frac{E_b}{N_J} > 0.709 \\ 1, & \frac{E_b}{N_J} \leq 0.709 \end{cases},$$

and the corresponding bit error probability is

$$P_b = \begin{cases} \frac{0.083}{E_b/N_j}, & \frac{E_b}{N_J} > 0.709 \\ Q\left(\sqrt{\frac{2E_b}{N_J}}\right), & \frac{E_b}{N_J} \leq 0.709 \end{cases}.$$

This situation is illustrated in Figure 10. If the jammer uses the worst-case duty cycle, then the bit error probability is only decaying as $1/(E_b/N_J)$ rather than exponentially in E_b/N_J. As seen from the figure, to reach $P_b = 10^{-4}$ we have to spend almost 21 dB more signal power compared to case for a broadband continuous noise jammer.

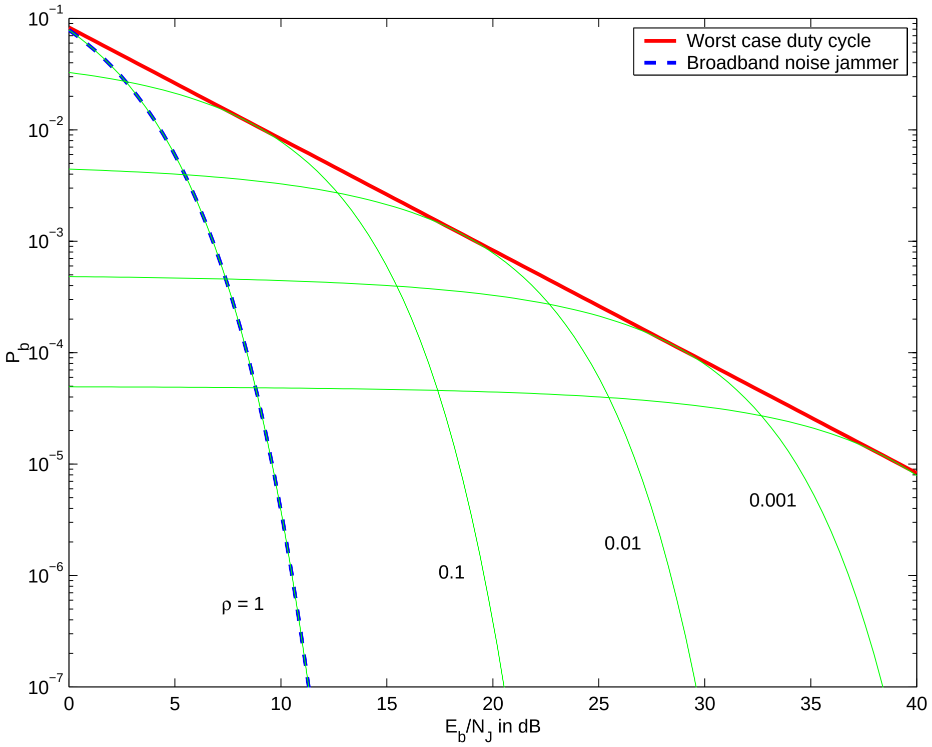

Figure 10: Bit error probability for BPSK-modulated DS-SS with pulsed noise jamming with duty cycle ρ.

3.5 Repetition Coding and Interleaving

The pulsed jammer situation is similar to a flat fading channel: at some times the channel is very bad, and this dominates the bit error probability. The remedy to overcome a pulse jammer is the same as for a flat fading channel. We introduce channel coding and interleaving to gain diversity.

Luckily, the DS-SS format has a built-in channel code (the repetition code), and we just need to introduce an interleaver before the BPSK modulator in Figure 4 and a deinterleaver in front of the descrambler in Figure 6. We will assume a block interleaver. That is, N consecutive chips (that corresponds to an information bit) are divided into m blocks of N/m chips (assumed to be an integer), and the blocks are interleaved. The parameter m must of course be less or equal to N.

With a sufficiently large interleaver depth, each chip will see an independent channel. The input to the channel decoder (i.e., the signal after deinterleaving) in Figure 6 during a certain bit interval is

$$r[k] = \sqrt{E_c}b + z[k]n[k],$$

where b is the information bit, $z[k]$ is binary random variable which is 1 if the chip was jammed and 0 if the chip was not jammed, and $n[k]$ is a Gaussian random variable with variance $N_J/2\rho$. We make the assumption that $z[k] = 1$ with probability ρ, i.e., we assume that during the transmission of a chip, the jammer is either on or off. Moreover, we make the assumption that an interleaved block of chips, i.e., a block of N/m consecutive chips are either jammed or not jammed with the probability ρ and $1-\rho$, respectively. This means that each interleaved block will see an independent channel, and that the channel is bad with probability ρ and good (perfect) with probability $1-\rho$. Hence, the parameter m is also the diversity order of the system.

The variable $z[k]$ represents the channel state information (CSI), which may or may not be available for the receiver (it is quite conceivable that the receiver is able to estimate which chips that have been jammed by, e.g., monitoring the received signal power). In the absence of CSI, the receiver can form the decision as the sign of

$$\sum_{n=0}^{N-1} r[n] = \sum_{l=0}^{m-1} \sum_{k=lN/m}^{(l+1)N/m-1} r[k],$$

where the right-hand side shows that the contributions from all interleaved blocks are simply added together (so-called equal gain combining). The above strategy is known as soft decoding without CSI, and the bit error probability is

$$P_b = \sum_{l=1}^{m} \Pr\{l \text{ blocks are jammed}\} \Pr\{\text{bit error if } l \text{ blocks are jammed}\},$$

and since (if $\rho < 1$)

$$\Pr\{l \text{ blocks are jammed}\} = \binom{m}{l} \rho^l (1-\rho)^{m-l}$$

and

$$\Pr\{\text{error if } l \text{ blocks are jammed}\} = Q\left(\sqrt{\frac{2E_b\rho}{N_J}\frac{m}{l}}\right),$$

we have that

$$P_b = \sum_{l=1}^{m}\binom{m}{l}\rho^l(1-\rho)^{m-l}Q\left(\sqrt{\frac{2E_b\rho}{N_J}\frac{m}{l}}\right), \qquad \rho < 1.$$

For $\rho = 1$, the use of interleaving will not affect the bit error probability, and we will have the same error probability as for broadband nonpulsed noise jamming, see equation (1). As seen from Figure 11, even though the curves for fixed ρ are improved for high E_b/N_J as the diversity increases, nothing is gained in terms of the worst-case performance.

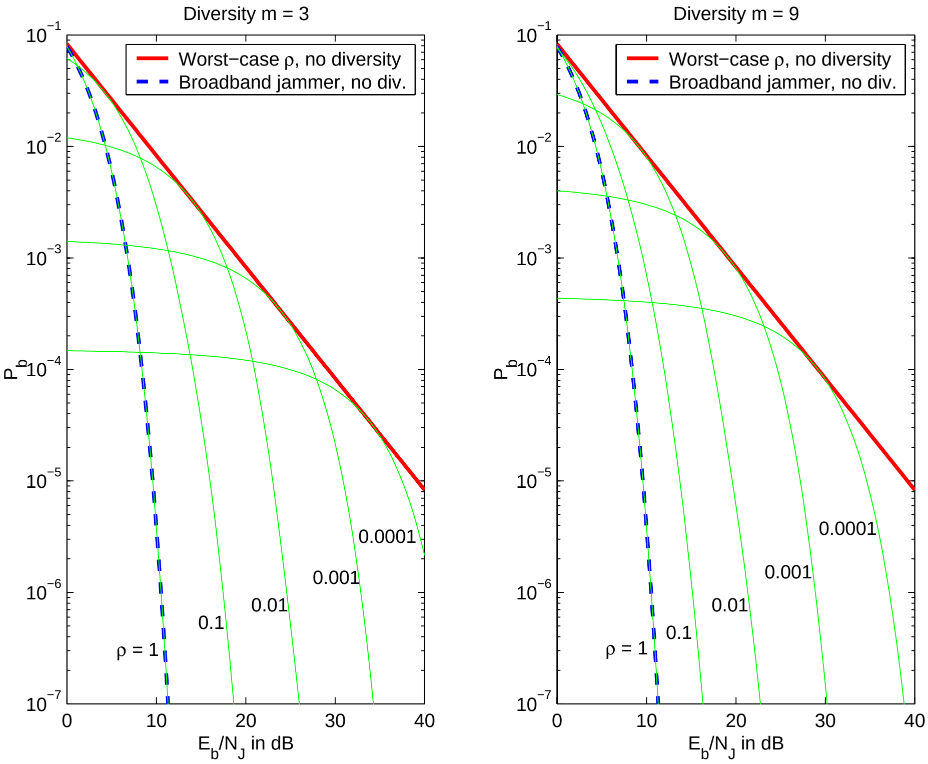

Figure 11: Bit error probability for BPSK DS-SS with diversity and soft decoding without channel state information (jammer information).

To improve the performance, we can do hard decoding instead. With hard decoding, we mean that we make a decision on the interleaved block of chips (coded bits) before decoding the channel code. In other words, the decision on the information bit is taken as the sign of

$$\sum_{l=0}^{m-1}\operatorname{sgn}\left\{\sum_{k=lN/m}^{(l+1)N/m-1} r[k]\right\}.$$

It is easy to show that

$$\operatorname{sgn}\left\{\sum_{k=lN/m}^{(l+1)N/m-1} r[k]\right\} = \begin{cases} b, & \text{if the chips are not jammed} \\ b, & \text{with probability } 1-q \text{ if the chips are jammed} \\ -b, & \text{with probability } q \text{ if the chips are jammed} \end{cases},$$

where q is the probability of error when decoding a block of N/m chips when the block is jammed, i.e.,

$$q = Q\left(\sqrt{\frac{2E_c(N/m)\rho}{N_J}}\right) = Q\left(\sqrt{\frac{2E_b\rho}{N_J}\frac{1}{m}}\right).$$

Hence, the probability that a block of interleaved chips will be decoded incorrectly is $p = \rho q$. The information bit error probability is the same as for a rate $1/m$ repetition code when the channel is a binary symmetric channel with crossover probability p. That is, the information bit error probability is, for odd m

$$\begin{aligned} P_b &= \Pr\{(m+1)/2 \text{ or more interleaved block errors}\} \\ &= \sum_{l=(m+1)/2}^{m} \binom{m}{l} p^l (1-p)^{l-k}, \end{aligned}$$

and for even m

$$P_b = \frac{1}{2}p^{m/2}(1-p)^{m/2} + \sum_{l=m/2+1}^{m} \binom{m}{l} p^l (1-p)^{m-l}.$$

Here, the extra term is to take in account the case when exactly half the chips in a code word are in error. The Hamming distance from the received word to both code words will then be the same ($m/2$), and we will make a decoding error with probability 1/2 in this case.

As seen in Figure 12, there is a significant performance gain compared with the worst-case jamming case, and the larger m is, the better the performance will get. However, for the $\rho = 1$ case, soft decoding is better than hard decoding, which is expected since soft decoding is optimum for $\rho = 1$.

It may be a little bit surprising that hard decoding yields better worst-case performance than soft decoding. For an ordinary AWGN channel it is known that soft decoding is approximately 2 dB better than hard decoding [14]. The reason for why hard decoding is preferable to soft decoding on a jamming channel is that we have not assumed any knowledge of the channel state in the decoding. Hence, we pick the code word which is closest to the received word in Euclidean and Hamming distance for soft and hard decoding, respectively. A jammed block of chips can add significant Euclidean distance but only a limited amount of Hamming distance. As a matter of fact, for hard decoding, more than half of the blocks of interleaved chips that corresponds to an information bit must be jammed before erroneous decoding is even possible, while it may be enough if just one block of chips is jammed in the soft decoding case.

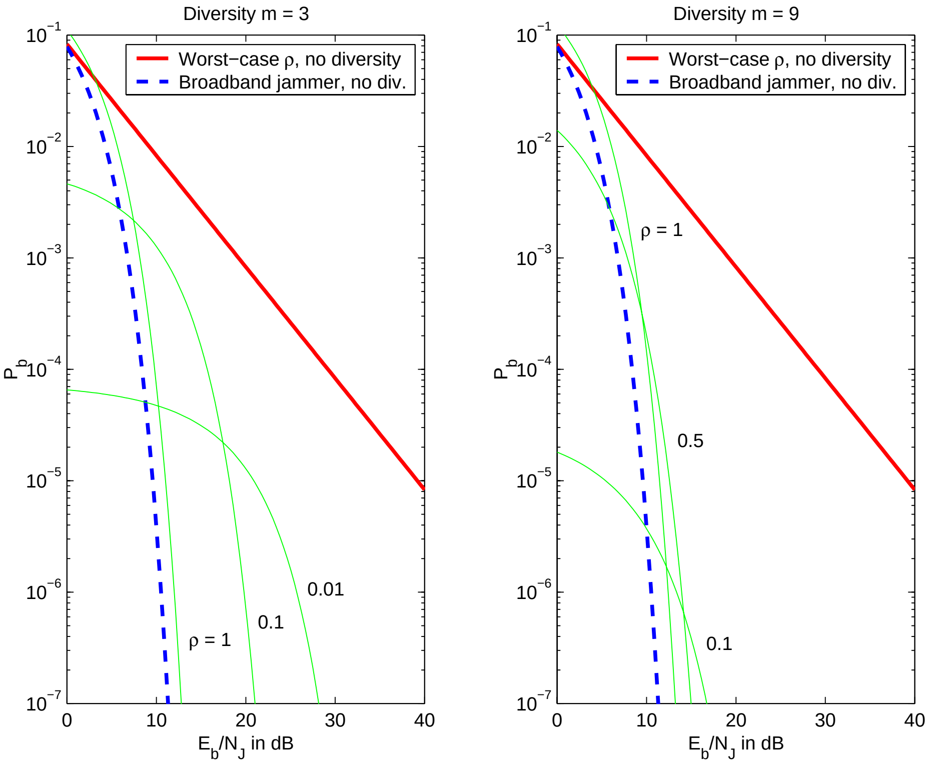

Figure 12: Bit error probability for BPSK DS-SS with diversity and hard decoding without channel state information (jammer information).

The performance for hard and soft decoding is significantly improved if we have access to CSI. Since we assume a perfect channel when the jammer is in the off state, we can always decode the information bit correctly if at least one chip is not jammed. Hence, the only time a decoding error can occur is when all chips are jammed, which happens with probability ρ^m. The resulting information bit error probabilities are for soft decoding

$$P_b = \rho^m Q\left(\sqrt{\frac{2E_b}{N_J}\rho}\right),$$

and for hard decoding and odd m

$$P_b = \rho^m \sum_{l=(m+1)/2}^{m} \binom{m}{l} q^l (1-q)^{m-l},$$

and for even m

$$P_b = \rho^m \frac{1}{2} q^{m/2}(1-q)^{m/2} + \rho^m \sum_{l=m/2+1}^{m} \binom{m}{l} q^l (1-q)^{m-l}.$$

The resulting bit error probabilities are illustrated in Figure 13. We note that soft decoding is roughly 2 dB better than hard decoding, and that for a moderate

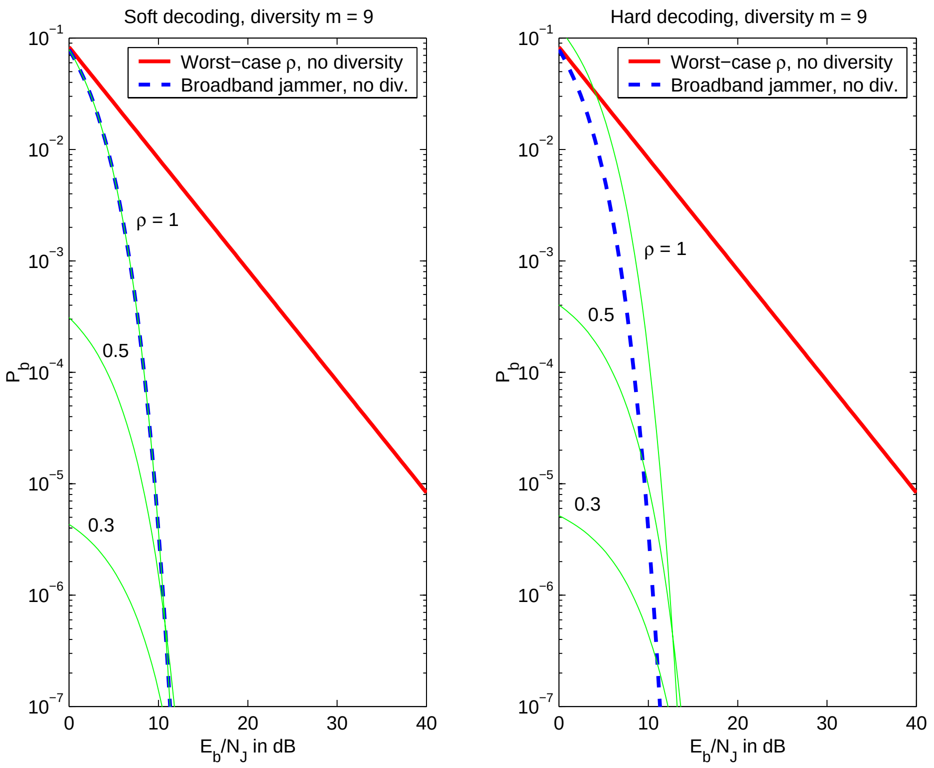

Figure 13: Bit error probability for BPSK DS-SS with diversity using soft and hard decoding with perfect channel state information (jammer information).

m, the worst-case duty cycle is $\rho = 1$ for most bit error probabilities of practical interest. Hence, the jammer is forced into becoming a broadband nonpulsed noise jammer and is thereby effectively defeated.

For the more realistic case when there is also white Gaussian noise added by the channel in addition to the jamming, we need to modify the decoding strategies for known CSI. For instance, in the soft decoding case, we should preferably use maximum-ratio combining, that is the soft decisions on the non-jammed chips should be weighted by $E_c^{1/2}/N_0$ and the jammed chips by $E_c^{1/2}/(N_0 + N_J)$. Moreover, for a fixed E_b/N_0, the bit error curves will exhibit error floors, i.e., the curves will approach an asymptotical bit error rate determined by E_b/N_0 as E_b/N_J becomes large (similar to the behavior illustrated in Figure 8).

To further improve the performance, we can use a better channel code than the implicit repetition code. Both block and convolutional codes are possible candidates; however, convolutional codes are usually the preferred choice if we want to do soft decoding. In a practical setting, a spread spectrum system with spreading factor N can be achieved by cascading a rate $1/n_{cc}$ convolutional code with a rate $1/n_{rc}$ repetition code, where $N = n_{cc} + n_{rc}$. The output from the last encoder should then be scrambled by a pseudo-random sequence for the system to be called a spread spectrum system.

3.6 Signal Space Interpretation

To reflect back on Section 2, we will now examine the basis used in DS-SS in more detail.

For the case when the chip waveforms are rectangular, we can form a set of orthonormal basis functions as

$$\begin{aligned} \phi_{I,k}(t) &= \Pi_{T_c}(t-kT_c)\sqrt{\frac{2}{T_c}}\cos(2\pi f_c t) \\ \phi_{Q,k}(t) &= \Pi_{T_c}(t-kT_c)\sqrt{\frac{2}{T_c}}\sin(2\pi f_c t) \end{aligned} \qquad k=0,1,\ldots \tag{2}$$

(The subscripts I and Q stand for inphase and quadrature phase, respectively.) The functions will not be strictly bandlimited, but if we allow for some spectral leakage outside the system bandwidth, we may set the chip rate to $1/T_c = W_{ss}/2$ (the system bandwidth is then equal to the null-to-null bandwidth of the transmitted signal).

Now consider the case when we want to transmit a (long) packet of P bits with bit rate $1/T_b$. We can find $2PT_b/T_c$ basis functions of the form described in (2) that are confined to the time interval $[0, PT_b]$. Hence, the number of basis function available per transmitted bit is

$$N_d = \frac{2PT_b}{T_c}\frac{1}{P} = 2\frac{T_b}{T_c}.$$

It is this fact that allows for channel coding with very low code rates to be used in DS-SS without loss of spectral efficiency.

The basis is illustrated in a time-frequency diagram as shown in Figure 14. Note that only one of the N_d possible linearly independent combinations of the basis functions is actually used for transmitting a bit (assuming BPSK modulation). Exactly which linear combination is used depends on the code sequence and changes from bit to bit. This is what signifies a spread spectrum system: the transmission uses a small dimensional space which is hidden in a large dimensional space, and the transmission space is chosen in a way which is independent of the transmitted data and appears random and time-varying to the jammer.

From Figure 14, we also see why a pulsed jammer is so effective against a noninterleaved DS-SS system. A certain information bit is concentrated to an interval of T_b seconds along the time axis. The jammer can concentrate its power to jam a subset of the transmitted bits by transmitting pulsed noise. The jammed bits are likely to be in error, and this will dominate the average bit error probability. The countermeasure is equally logical. By introducing interleaving, we scatter the chips over a wide time interval, and a pulsed jammer is unlikely to hit all chips corresponding to an information bit. Diversity is therefore introduced, and the probability of correct decoding is consequently increased (especially if the receiver has access to reliable channel state information).

For the case when we have root raised cosine pulse shapes, a set of orthonormal functions can be formed as

$$\begin{aligned} \phi_{I,k}(t) &= x_{RRC}(t-kT_c)\sqrt{2}\cos(2\pi f_c t) \\ \phi_{Q,k}(t) &= x_{RRC}(t-kT_c)\sqrt{2}\sin(2\pi f_c t) \end{aligned} \qquad k=0,1,\ldots \tag{3}$$

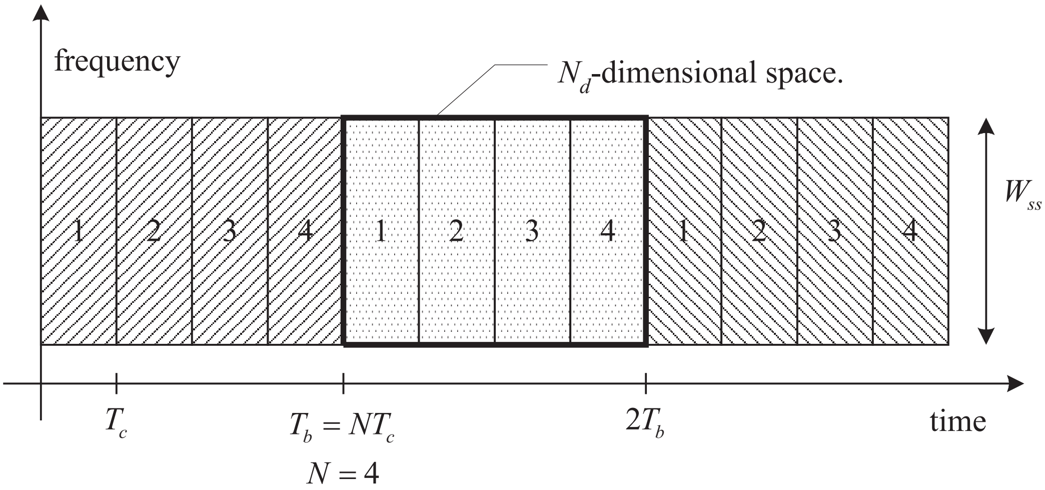

Figure 14: Time-frequency plot of basis used for DS-SS (only the positive frequencies are shown). The boxes with the same background pattern are used for the same bit. Note that this figure shows the noninterleaved case.

where $x_{RRC}(t)$ is a pulse with a root raised cosine spectrum with roll-off factor α and f_c is the carrier frequency (located in the middle of the system bandwidth). The spectrum of the basis functions will span the entire system bandwidth if

$$\frac{1+\alpha}{T_c} = W_{ss}.$$

Since $x_{RRC}(t)$ has infinite extension in time, we have to truncate the pulses in a practical implementation. If the pulse after truncation is sufficiently long, we will not lose too much of the spectral compactness and we may still regard the pulses as bandlimited to the system bandwidth.

Now consider the case when we want to transmit a (long) packet of P bits with bit rate $1/T_b$. Ignoring the edge effects due to that the pulses have longer extension than T_c seconds and letting α be very small, we can find $2PT_b/T_c$ basis functions of the form described in (3). Hence, the number of basis function available per transmitted bit is

$$N_d = \frac{2PT_b}{T_c}\frac{1}{P} = 2\frac{T_b}{T_c},$$

which is the same as in the rectangular chip waveform case.

4 Frequency-Hopping Spread Spectrum

The basis functions used in the DS case are designed to be time-limited and to span the entire frequency band. The larger the system bandwidth, the shorter the pulses must be. Hence, for DS spread spectrum, W_{ss} is limited to few hundred MHz due to hardware constraints. This implies that the processing gain is also limited. Moreover, the system bandwidth must be contiguous.

Both these limitations can be avoided by using a frequency-hopped spread spectrum (FH-SS) system. If we let the baseband signal be denoted by $x(t)$, the transmitted signal in a FH-SS system is

$$s(t) = x(t)\sqrt{2}\cos(2\pi f_k t + \theta_k), \qquad kT_h \leq t < (k+1)T_h$$

where $1/T_h$ is the frequency hopping rate and f_k and θ_k is the carrier frequency and phase after the kth hop. The reason for including a carrier phase is that it may be difficult to implement a system with continuous carrier phase. As we will see, it is important that the hopping rate is relatively high, and it may therefore be difficult for the receiver to estimate the carrier phase, and noncoherent methods are therefore widely used. In the following, we will assume that M-ary frequency-shift keying (FSK) is used, and that the frequency spacing will allow for noncoherent detection.

A system can be either fast hopping or slow hopping. The system is said to be fast hopping if the frequency hop rate is larger than the symbol rate, i.e., if $1/T_h > 1/T_s$, where $1/T_s$ is the symbol rate. Logically, a system is said to be slow hopping if $1/T_h \leq 1/T_s$. We should here interpret $1/T_s$ as the rate of the possibly coded symbols. For an uncoded system, the symbol rate is $\log_2(M)/T_b$ where $1/T_b$ is the information bit rate, while for a system which employs a rate R_c error-control code, the symbol rate is $\log_2(M)/(R_c T_b)$.

For a slow hopping system, the frequency spacing between the signal alternatives must be a multiple of $1/T_s$ to allow for noncoherent detection (we want the signal alternatives to be orthogonal, regardless of the carrier phase). However, if the system is fast hopping, a symbol will span more than one hop, and we must ensure that the partial symbols within a hop are orthogonal (regardless of the carrier phase). Hence, the signal alternatives must be spaced by a multiple of $1/T_h$.

To account for both fast and slow hopping systems, we define the chip rate as the maximum rate with which the transmitted signal can change frequency, i.e.,

$$\frac{1}{T_c} = \max\left\{\frac{1}{T_h}, \frac{1}{T_s}\right\}.$$

The chip rate will decide the minimum frequency spacing between the signal alternatives as $\Delta_f = 1/T_c$.

For slow hopping systems, we must therefore change the subcarrier spacing when introducing coding (assuming constant M and information bit rate). Depending on the code, symbol, and hop rates, this may or may not be necessary for fast hopping systems.

To summarize, we construct the transmitted signal from orthonormal functions from the set

$$\phi_{k,l}(t) = \Pi_{T_c}(t - kT_c)\sqrt{\frac{2}{T_c}}\cos(2\pi[f_0 + l\Delta_f]t + \theta_k), \qquad (4)$$

where k is the hop index, $l = 0, 1, \ldots, N_f - 1$ is the subcarrier index, N_f is the total number of subcarriers, and θ_k is the carrier phase after hop k (assumed to be unknown to the receiver). If we ignore the spectral leakage due to the use of rectangular pulses, the number of subcarriers is $N_f = W_{ss}T_c$. We have here assumed that the system bandwidth is contiguous, but we can easily redistribute the subcarriers to allow for a noncontiguous system bandwidth. Furthermore, since the fundamental pulse length, T_c, is not directly coupled to W_{ss}, we can allow for very large processing gains with reasonable hardware complexity.

There exist many ways to compile a transmitted signal from the functions in (4). We can let the M frequencies (corresponding to one symbol) to be adjacent or nonadjacent, and we can let the carrier frequency jump a multiple of $M\Delta_f$ or to any subcarrier (or perhaps even to any frequency in the system bandwidth). Which method to use depends on which hardware complexity we can afford. The least structure in the transmitted signal is clearly when we allow any subcarrier to be transmitted in any given hop. However, this will also require the most complex hardware.

Time-frequency plots for a fast hopping and a slow hopping FH-SS system are shown in Figure 15. In both plots, $M = 4$ adjacent subcarriers are used as signal alternatives in a 4-ary FSK constellation, and any subcarrier can be the main carrier. In the fast hopping system (the plot to the left), a symbol is split over two hops, and in the slow hopping system, two symbols are transmitted per hop.

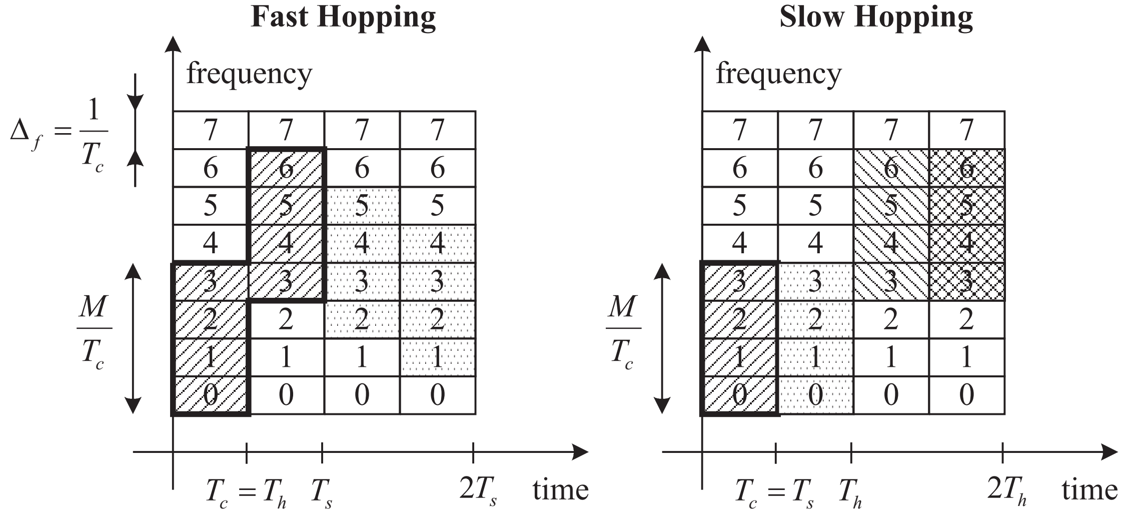

Figure 15: Time-frequency plot of FH-SS (only the positive frequencies are shown) with 4-ary FSK modulation. The boxes with the same background pattern are used for the same symbol.

4.1 Broadband Noise Jamming

Again, the most benign form of jamming is broadband nonpulsed noise jamming. For uncoded binary FSK with noncoherent detection, the resulting bit error probability is

$$P_b = \frac{1}{2}e^{-\frac{E_b}{2N_J}},$$

where $N_J = J/W_{ss}$.

4.2 Partial-Band Noise Jamming

Compared to broadband noise jamming, the performance for the uncoded system becomes much worse for pulsed broadband noise jamming or nonpulsed partial-band noise jamming. The reason is the same as for the DS case: when the jammer can concentrate its power to affect a fraction of the transmitted symbols, the jammed symbols will be erroneously decoded with a high probability, and this conditional error probability will dominate the average error probability. Just as in the DS case, we can affect certain symbols by pulsing the jammer along the time axis. However, we can also "pulse" the jammer along the frequency axis. This is called partial-band jamming.

A jammer that uses partial-band noise jamming will concentrate its power to a bandwidth ρW_{ss}, where ρ is the frequency domain duty cycle ($0 < \rho \leq 1$). The jammer signal is Gaussian noise with a flat power spectral density over the jammed bandwidth, i.e., in the jammed band the power spectral density is

$$\frac{J}{2W_{ss}\rho} = \frac{N_J}{2\rho}$$

If we assume that the jammed bandwidth is placed such that all signal alternatives (subcarriers) used for a certain symbol are either jammed or not jammed, then the probability that a certain symbol will be jammed is ρ. The resulting bit error probability for BFSK with noncoherent detection is then

$$P_b = (1-\rho) \times 0 + \rho \frac{1}{2} \exp\left(-\frac{E_b \rho}{2N_J}\right) = \frac{\rho}{2} \exp\left(-\frac{E_b \rho}{2N_J}\right).$$

The worst-case duty cycle is easily found to be

$$\rho_{wc} = \begin{cases} \frac{2}{E_b/N_J}, & E_b/N_J > 2 \\ 1, & E_b/N_J \leq 2 \end{cases},$$

and the worst-case bit error probability is

$$P_{b,wc} = \begin{cases} \frac{1}{eE_b/N_J}, & E_b/N_J > 2 \\ \frac{\rho}{2} \exp\left(-\frac{E_b \rho}{2N_J}\right), & E_b/N_J \leq 2 \end{cases}.$$

We note that, just as for the case of DS-SS with worst-case pulsed broadband noise jamming, the error probability is made inversely proportional to E_b/N_J for sufficiently large E_b/N_J. This is a significant degradation from the broadband noise jamming case, for which the error probability falls of exponentially with E_b/N_J.

4.3 Multitone Jamming

An efficient jamming method for FH-SS is multitone jamming. Recall the number of subcarriers is $N_f = W_{ss}T_c$. A jammer (which is assumed to have knowledge of where the subcarriers are located) can choose to jam certain subcarriers by transmitting pure tones at the selected subcarriers. To describe the worst-case jamming strategy, i.e., how the jammer should choose which subcarriers to jam and how to assign the power to each jammer tone, is actually rather involved. The reader is referred to [4] for a detailed discussion on this. In short, it is shown in [4] that a simple but effective jamming strategy is to randomly select which subcarriers to jam, and to assign a power to each jamming tone which is slightly larger than the power of the received information tone. This strategy, known as independent multitone jamming, is more harmful than worst-case partial-band noise jamming.

A more complicated jamming strategy, which under some circumstances is more effective than independent multitone jamming, is called band multitone jamming. Band multitone jamming can be used if the jammer knows exactly how the transmitter partitions the N_f possible subcarriers into bands of M subcarriers for transmission of a symbol. Note that this is not the same as that the jammer knows which band is used during a certain symbol interval; the jammer only knows to what band any subcarrier belongs. The jammer can now choose to jam exactly n or none of the subcarriers in a band. This means that the jammer either jams a band or not, and if the band is jammed, exactly n of the M subcarriers are jammed in that band ($0 < n \leq M$).

4.4 Repetition Coding and Interleaving

To combat partial-band or multitone jamming, we can use channel coding and interleaving to spread the information bits over several transmissions. With the appropriate decoding strategy, we can gain back most or all of the performance loss from broadband noise jamming to partial-band or multitone jamming. The details on how to form the decision metric and the resulting performance is a bit involved and will be excluded here. Again, the interested reader is referred to [4] for a detailed discussion.

5 Commercial Spread Spectrum Systems

As mentioned earlier, the number of nonmilitary spread spectrum systems have increased rapidly the last decades. The applications are quite diverse: underwater communications [15], wireless local loop systems [16], wireless local area networks, cellular systems, satellite communications [17], and ultra wideband systems [18]. Spread spectrum is also used in wired application in, e.g., power-line communication [19] and have been proposed for communication over cable-TV networks [16] and optical fiber systems [20, 21]. Finally, spread spectrum techniques have been found to be useful in ranging, e.g., radar and navigation, e.g., the Global Positioning System (GPS) [22]. Other applications are watermarking of multimedia [23] and (mentioned here as a curiosity) in clocking of high-speed electronics [24, 25]. Due to space constraint, we will only briefly mention some of the hot wireless applications here.

The wireless local area network (WLAN) standard IEEE 802.11 was originally designed to operate in the ISM band at approximately 2.4 GHz. The standard supports several different coding and modulation formats and several data rates. The first version of the standard was released in 1997 and supports both FH and DS spread spectrum formats with data rates of 1 or 2 Mbit/s [26]. The FH modes are slow hopping and use so-called Gaussian FSK (GFSK) modulation (binary for 1 Mbit/s and 4-ary for 2 Mbit/s). The system hops over 79 subcarriers with 1 MHz spacing. The DS-SS modes use a 11-chip long Barker sequence which is periodically repeated for each symbol. The chip rate is 11 Mchips/s, and the symbol rate is 1 Msymbols/s. The modulation is differentially encoded BPSK or QPSK (for 1 and 2 Mbit/s, respectively). We note that the processing gain is rather low, especially for the DS-SS modes.

The 802.11 standard has since 1997 been extended in several directions (new bands, higher data rates, etc). In 1999, the standard was updated to IEEE 802.11b (also known as Wi-Fi, if the equipment also passes an interoperability test). In addition to the original 1 and 2 Mbit/s modes, IEEE 802.11b also supports 5.5 and 11 Mbit/s DS-SS modes [27] and several other optional modes with varying rates. The higher rate DS-SS modes uses so-called complementary code keying (CCK). The chip rate is still 11 Mchips/s and each symbol is represented by 8 complex chips. Hence, for the 5.5 Mbit/s mode, each symbol carries 4 bits, and for the 11 Mbit/s mode, each symbol carries 8 bits. Hence, the processing gains is reduced compared to the 1 and 2 Mbit/s modes. As a matter of fact, the 11 Mbit/s is perhaps not even a spread-spectrum system. The CCK modulation is a little bit complicated to describe, but in essence it forms the complex chips by combining a block code and differential QPSK [26, Section 18.4.6.5].

Bluetooth is primarily a cable replacement system, i.e., a system for short range communication with relatively low data rate. It is designed for the ISM band and uses slow hopping FH-SS with GFSK modulation ($BT = 0.5$ and modulation index between 0.28 and 0.35). The systems hops over 79 subcarriers with a rate of 1600 hops/s. The subcarrier spacing is 1 MHz, and in most countries the subcarriers are placed at $f_k = 2402 + k$ MHz for $k = 0, 1, \ldots, 78$. Bluetooth supports both synchronous and asynchronous links and several different coding and packet schemes. The user data rates varies from 64 kbits/s (symmetrical and synchronous) to 723 kbits/s (asymmetrical and asynchronous). The maximum symmetrical rate is 434 kbits/s. The range of the system is quite short, probably less than 10 m in most environments. It is likely that future versions of Bluetooth will support higher data rates and longer ranges.

The first cellular system with a distinct spread spectrum component was IS-95 (also known as cdmaOne or somewhat pretentiously as CDMA). Although the Global System for Mobile Communications (GSM), has a provision for frequency hopping, it is not usually considered to be a spread spectrum system. Often, spread spectrum and code-division multiple access (CDMA) are used as synonyms, although they really are not. A multiple access method is a method for allowing several links (that are not at the same geographical location) to share a common communication resource. CDMA is a multiple access method where the links are spread spectrum links. A receiver that is tuned to a certain user relies on the anti-jamming properties of the spread spectrum format to suppress the other users' signals.

IS-95 uses DS-SS links with a chip rate of 1.2288 Mchips/s and a bandwidth of (approximately) 1.25 MHz. In the downlink (forward link or base-station-to-terminal link) the chips are formed by a combination of convolutional encoding, repetition encoding, and scrambling. The chips are transmitted both in inphase and quadrature (but scrambled by different PN sequences). In the uplink (reverse link), the transmitted chips are formed by a combination of convolutional coding, orthogonal block coding, repetition coding, and scrambling. In the original IS-95 (IS-95A), the uplink was designed such that the detection could be done noncoherently. In the third generation evolution of IS-95, known as cdma2000, the modulation and coding has changed and the transmitted bandwidth tripled to allow for peak data rates exceeding 2 Mbit/s [28, 29, 30].

Another third generation system is Wideband CDMA (WCDMA) [31]. WCDMA is a rather complex system with many options and modes. We will here only briefly describe the frequency-division duplex (FDD) mode. The FDD mode uses direct-sequence spreading with a chip rate of 3.84 Mchips/s. The chip waveform is a root-raised cosine pulse with roll-off factor 0.22, and the bandwidth of the transmitted signal is approximately 5 MHz. WCDMA supports many different information bit rates by changing the spreading factor (from 4 to 256 in the uplink and from 4 to 512 in the downlink) and the error control scheme (no coding, convolutional coding, or turbo coding); however, the modulation is QPSK with coherent detection in all cases. Today, the maximum information data rate is roughly 2 Mbits/s, but it is likely that future revisions of the standard will support higher rates through new combinations of spreading, coding, and modulation.

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