Chapter 1

Spread Spectrum Concepts

1.1 Introduction

Spread spectrum communications systems are often used when there is a need for message security and confidentiality, or where there is a requirement that the message be received error free [6], [7]. A spread spectrum system is able to offer a very high degree of message security in a number of ways, depending on the system implementation. Such a system may spread the data over a very wide bandwidth, making it almost impossible for a narrow band receiver to decipher any useful information. Along a similar vein, the same system may be able to offer a very high degree of interference rejection, both from intentional, and unintentional sources.

A spread spectrum communications system is usually characterised as one in which the transmission bandwidth is much greater than that necessary to transmit the required information [8]. In addition, demodulation must be accomplished by correlating the received signal with a replica of the signal used to spread the information. There are several communications systems which satisfy the first criteria, for example, wideband frequency modulation (W-FM), however, they do not qualify as spread spectrum systems. There are a number of advantages in spreading the transmitted data over a wide range of frequencies, but there are also some disadvantages. Most disadvantages center around the increased hardware complexity, and the software algorithms needed for efficient operation.

There are several different types of spread spectrum communications systems, and these are usually classified according to their modulation techniques. Some of the more commonly used systems include:

- Direct sequence (DS),
- Frequency hopping (FH),

- Time hopping,
- Chirp, and
- a combination of the above, usually known as a hybrid DS/FH system.

The more commonly used modulation techniques shall be briefly examined in order to provide a background for the hardware designs presented later in this thesis.

1.2 Direct Sequence Modulation

A direct sequence spread spectrum system achieves its spreading capability by modulating a narrow bandwidth data signal with a wide bandwidth spreading signal. Figure 1.1 shows a functional block diagram for a binary phase modulated (BPSK) direct sequence system, with the spectrum of the transmitted signal shown in Figure 1.2.

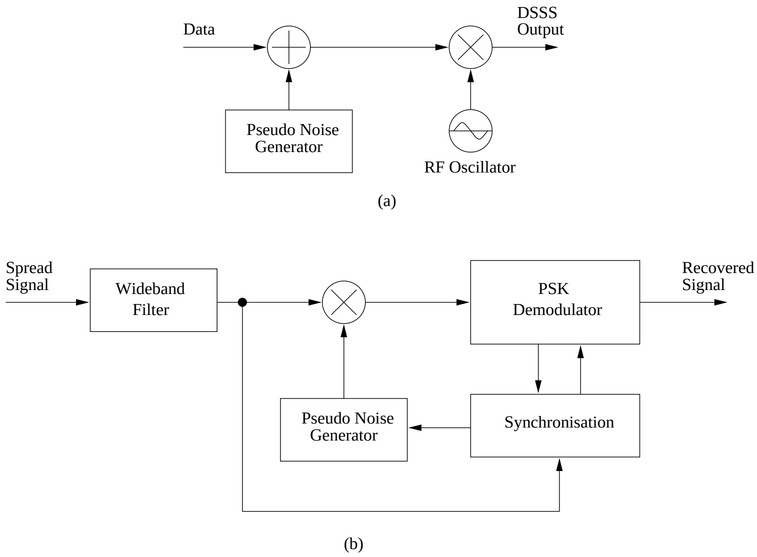

Figure 1.1: Binary phase modulated direct sequence system, (a) Transmitter and (b) Receiver

This type of system is probably the most commonly utilised spread spectrum technique, where the binary data is modulo-2 added with a pseudo-random spreading signal before being phase modulated. The receiver design is usually more complicated, but generally a coherent phase shift keying (PSK) demodulator is used.

At the receiver a phase shift modulated waveform is multiplied by a locally generated pseudo-random sequence, identical to that used in the transmitter to encode the modulation. The effect of this is complementary to that seen in the transmitter, in that the input signal is phase-reversed each time the locally generated

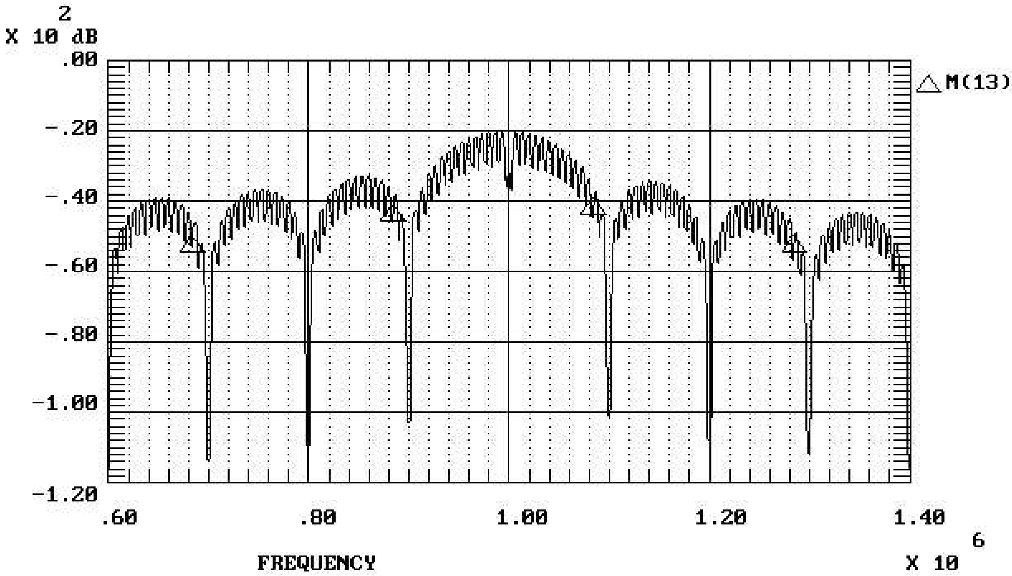

Figure 1.2: A typical direct sequence spectrum

code passes through a binary level transition. This is shown diagrammatically in Figure 1.3, where it is seen that at each phase reversal of the transmitted signal, the receiver also reverses the phase. These two complementary inversions cancel each other out, restoring the original signal.

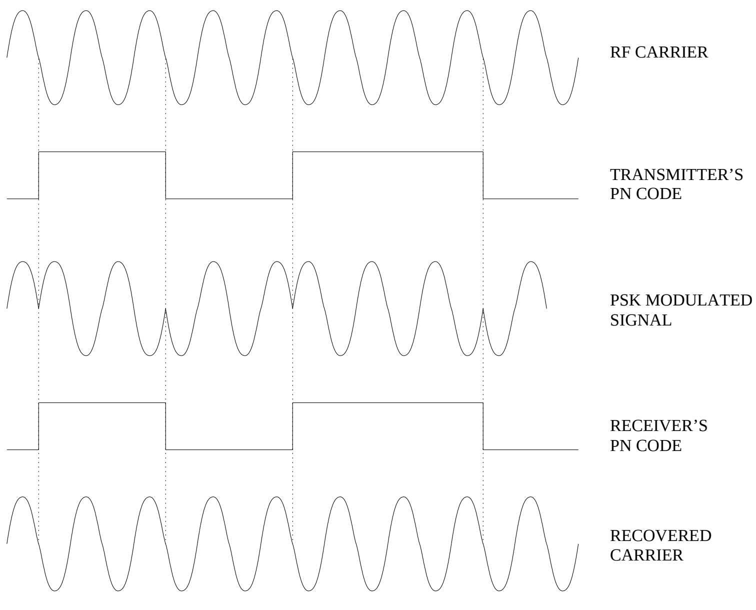

Figure 1.3: Carrier recovery using an in-line correlator

Although simple, this method of correlation is not often used in practice. It's major downfall is inherent in the demodulation, where the input frequency, as well as the demodulated signal, are both centered at a common frequency, say f_0. Although not a problem in itself, this in-line method of correlation is very susceptible to

interference, as any signal in the vicinity of f_0 will be mixed directly into the receiver [9]. A more reliable method is to use a heterodyne correlator, such that the output is at a different center frequency from the input signal.
At the receiver, the spread spectrum signal can be represented by

$$\psi(t) = \sqrt{2P}m(t)p(t)\cos(\omega_o t + \theta) \tag{1.1}$$

where P is the signal power, $m(t)$ represents the data modulation, $p(t)$ is the pseudo-random spreading waveform, ω_0 is the carrier frequency and θ is the phase angle at $t = 0$. Each pulse of $m(t)$ has a duration of T_m, while each pulse of $p(t)$ has a duration of T_p. As message privacy requires that the transitions of the data signal and the spreading waveform coincide on both sides of the symbol, the ratio of T_m to T_p must be an integer. Now, if W if the bandwidth of $\psi(t)$ and B is the bandwidth of $m(t)\cos(\omega_0 t)$, the spreading due to $p(t)$ results in $W \gg B$.

Assuming that the receiver has synchronised to the transmitter's spreading code, the received signal passes through a wideband filter, and is multiplied by a locally generated replica of $p(t)$. Given that $p(t) = \pm 1$, then $p^2(t) = 1$, and this operation yields the de-spread signal,

$$\psi_1(t) = \sqrt{2P}m(t)\cos(\omega_0 t + \theta) \tag{1.2}$$

at the input to the demodulator. Since $\psi_1(t)$ follows the definition of a PSK signal, the process of demodulation recovers $m(t)$.

1.2.1 Interference Rejection

The process of interference rejection is shown diagrammatically in Figure 1.4. Figure 1.4(a) shows the relative spectra of the desired signal and the unwanted interference at the receiver before being de-spread. After multiplication by the spreading waveform, the de-spread signal is shown in Figure 1.4(b).

The desired signal's bandwidth is reduced to B, whilst the unwanted interference has had it's bandwidth further spread over a bandwidth exceeding W. The demodulation stage acts to filter most of the interference spectrum that does not coincide with the desired signal spectrum. Hence, most of the interfering signal is eliminated and does not affect the operation of the receiver. A term commonly found in spread spectrum literature is the processing gain, and is an approximate measure of a receivers' interference rejection. It is given by the ratio $G_p = \frac{T_m}{T_p}$ and is equal to the number of pseudo-random bits in a data symbol interval. Now since W and B are usually proportional to $1/T_p$ and $1/T_m$ respectively, then

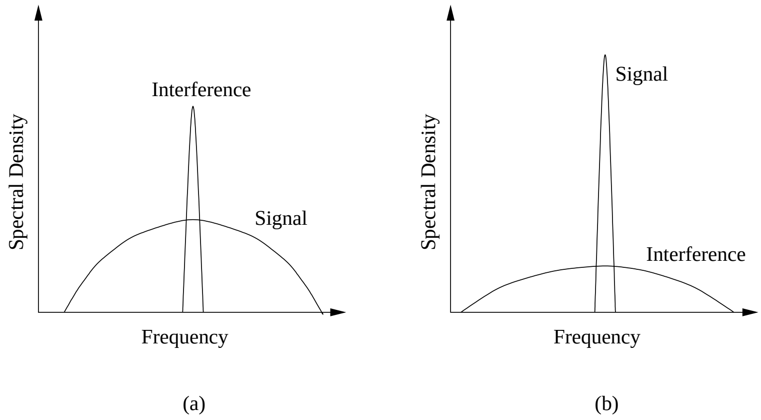

Figure 1.4: Effect of de-spreading: (a) Spectra before PN multiplication and (b) after multiplication

$$\text{Processing gain, } G_p = \frac{W}{B} \tag{1.3}$$

thus relating the interference rejection shown in Figure 1.4 with the system's processing gain.

1.2.2 QPSK Modulation

Phase quadrature modulation allows the simultaneous transmission of data using two carriers. The principal reason for doing this is to conserve spectrum, since, for the equivalent total transmitted power, the same bit error probability is achieved using half the transmission bandwidth [3]. Quadrature modulation schemes are used in spread spectrum systems since they are more difficult to detect in low probability of detection applications [1], and are less sensitive to certain methods of jamming. Both the spreading modulation and data modulation can be combined on quadrature carriers using a number of techniques. One example is shown in Figure 1.5. Note that the power at either output of the quadrature hybrid is half of the input power. The output of the QPSK modulator is given by

$$s(t) = \sqrt{P}c_1(t)\cos[\omega_0 t + \theta_d(t)] - \sqrt{P}c_2(t)\sin[\omega_0 t + \theta_d(t)] \tag{1.4}$$

where $c_1(t)$ and $c_2(t)$ are the in-phase quadrature spreading waveforms. It is assumed that both spreading waveforms only take on values of ± 1.

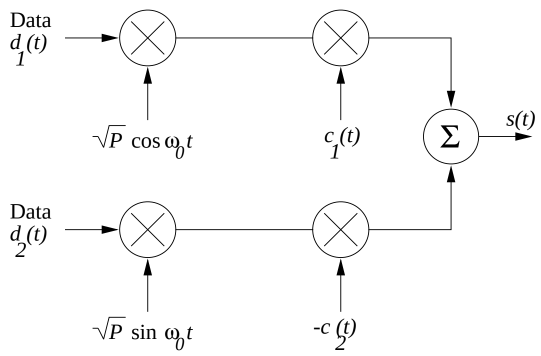

Figure 1.5: Quadrature phase modulated DS transmitter

1.3 Frequency Hopping

A frequency hopping communications system differs markedly from a direct sequence system. Whereas in a DS system, the carrier frequency remains constant, and the data is spread over a wide band of frequencies, in a frequency hopped system, the data is transmitted using a conventional narrow-band technique, but the carrier frequency is changed in discrete hops over a wide bandwidth. A typical FH spectrum is shown in Figure 1.6. Note that the bandwidths of each individual carrier may or may not overlap depending on the system design.

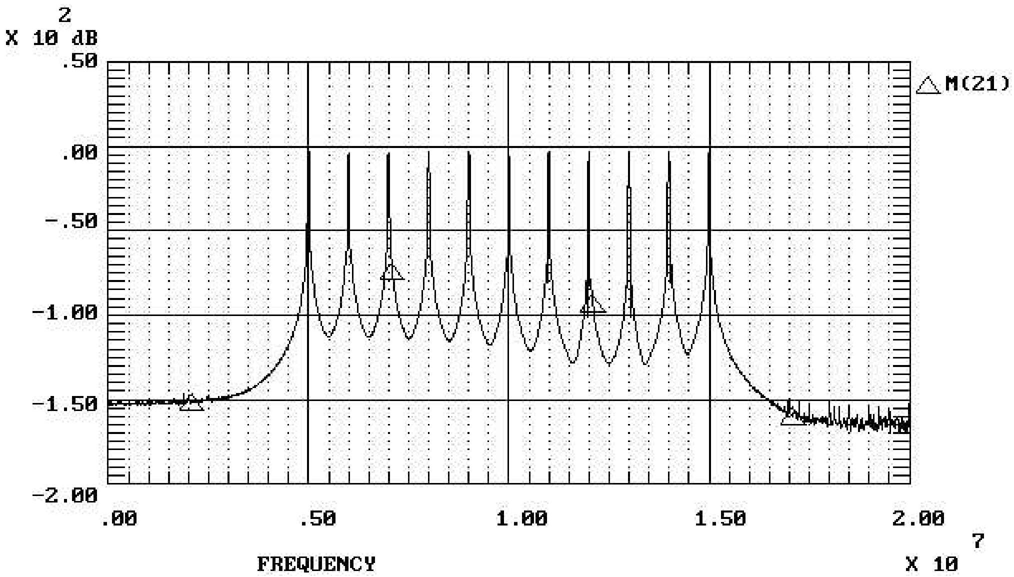

Figure 1.6: A typical frequency hopping spectrum

This technique is also very secure in that a narrow band receiver will be unable to detect any useful information, receiving just a short burst of information on odd occasions. Frequency hopping systems also have the ability to reject both intentional and unintentional interference, although somewhat more involved than the direct sequence case. A frequency hopping transmitter is shown diagrammatically in Figure 1.7. The receiver uses the same technique as the transmitter, but in reverse, along a similar line to the DS SS transmit/receive pair. Note that although a second mixer stage and RF oscillator are shown, their application depends on the desired carrier frequency. If the synthesizer is able to function at the required rate and generate the appropriate frequencies, then there is no need for a subsequent mixing stage.

The basic FH SS system has a number of disadvantages, which can be overcome quite elegantly [10]. Since a FH system can essentially be regarded as an instantaneous narrow band communications system, it suffers from the same problems with interference as any other conventional narrow-band modulation system, although the period of such degradation is considerably reduced. The classical example used to describe this aspect of FH SS is a peak, or spike of narrow-band noise somewhere in the desired spectrum. If the amplitude of this interference is sufficiently high, reception of a narrow-band signal at the same point in the spectrum is made difficult, if not impossible. This is the same for either a conventional narrow-band system, or a FH SS system. In order to reduce the possibility of narrow-band interference degrading a FH SS signal, a redundant method of transmission is used. Such a system generally transmits the same information at a number of different, discrete frequencies, over a short time period, with the receiver comparing the received signal at each frequency in order to compensate for any errors. There are two common methods of transmitting the redundant information. The first system, shown in Figure 1.8(a), transmits information at only one point in the spectrum at any one time, but repeats the transmission at a number of different frequencies. Although only one carrier frequency has been shown here, it is feasible to use more than one frequency, although with a corresponding increase in complexity, and a reduction in the overall speed of operation. The second method, shown in Figure 1.8(b), transmits the same information at several different points in the spectrum simultaneously. Obviously, the second method requires more hardware and is slightly more complicated than the first system, but offers a higher degree of interference rejection. The method shown in Figure 1.8(b) also offers a higher speed of operation, since there is no redundancy in time, only in frequency.

1.4 Hybrid Spread Spectrum Systems

A hybrid spread spectrum system generally consists of a combination of a direct sequence system, and a frequency hopping system. Such a system, shown in Fig-

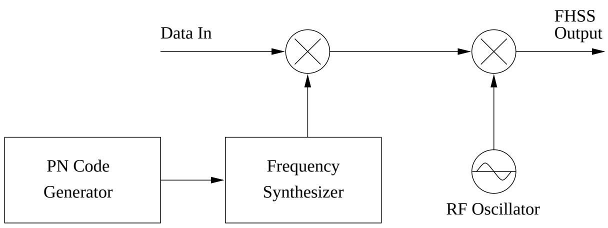

Figure 1.7: Frequency hopping spread spectrum transmitter

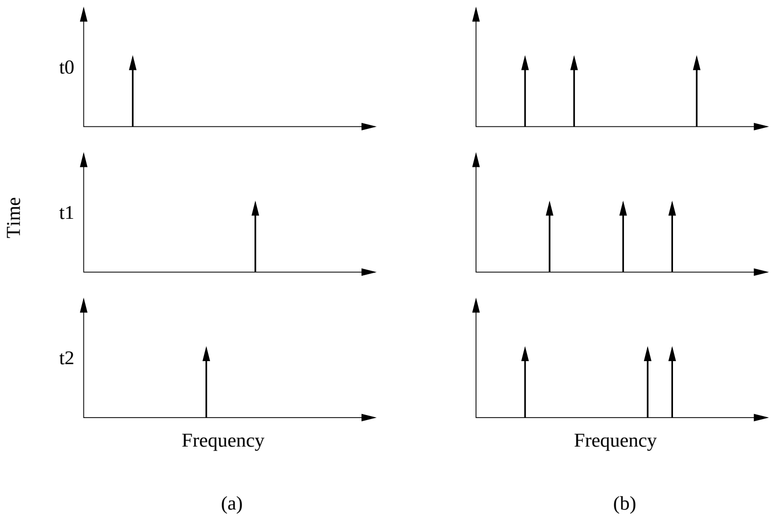

Figure 1.8: Comparison between different methods of frequency hopping

ure 1.9, will usually possess a very high degree of information security, as well as a high ability to reject interference.

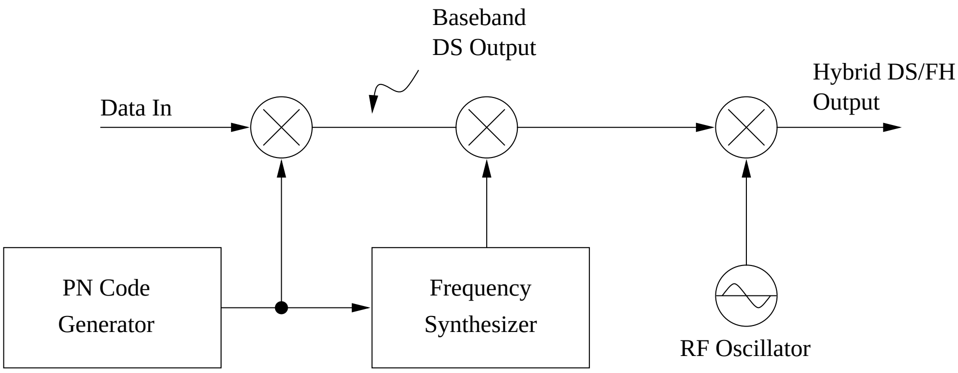

Figure 1.9: Hybrid DS/FH spread spectrum transmitter

The disadvantage of this type of system is the overall increase in complexity and operation. A hybrid system can be thought of as a direct sequence system in which the carrier frequency is changed periodically. The information to be transmitted is spread by mixing with a PN sequence, but the band of frequencies over which the data is spread is changed at a rapid rate. It is very difficult for a narrow band listener to intercept and gather information from a direct sequence transmission, but when the entire spread bandwidth is hopping around the spectrum, this task becomes almost impossible.

Figure 1.10 shows the spectrum one can expect from a hybrid SS system. In this case the spectra are shown equally spaced, although this need not be the case. Each individual direct sequence spectrum may be positioned anywhere in the frequency hoppers range, and may even overlap with one another.

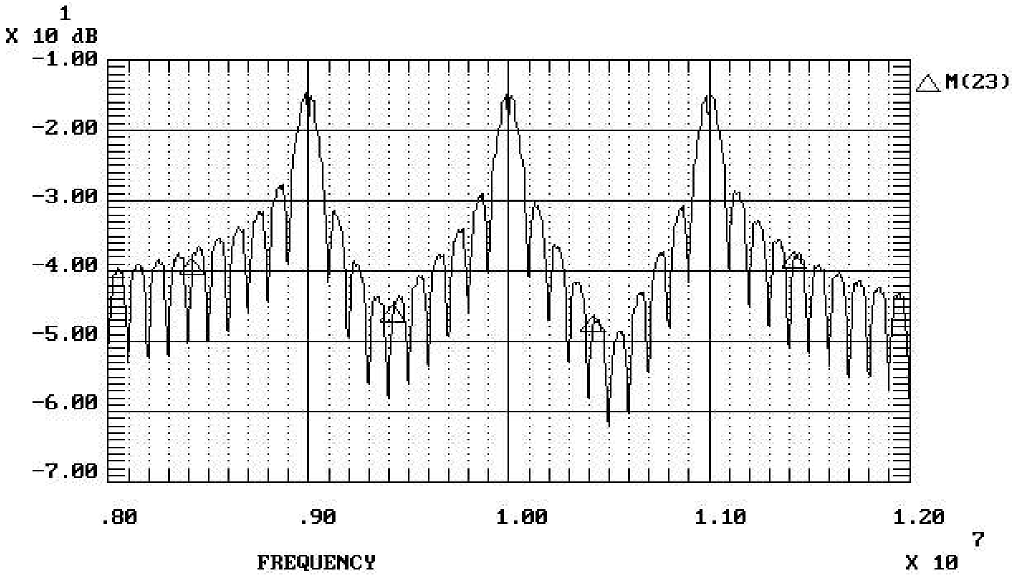

Figure 1.10: Spectrum of a hybrid DS/FH system

1.5 Time Hopping and Chirp Modulation

Both time hopping and chirp modulation schemes have been used to implement spread spectrum systems. Their use in communications systems is not as common as direct sequence or frequency hopping, however they are mentioned here for completeness.

1.5.1 Time Hopping

Consider the time hopping waveform shown in Figure 1.11. Note that the time axis has been divided into intervals known as frames, with each frame divided into M time slots. During each frame only one time slot may be modulated by a message, with each particular time slot being chosen according to the output of a PN generator. There are a total of $M = 2^m$ time slots in each frame, and all of the message bits accumulated in the previous frame are sent in a burst during the selected time slot.

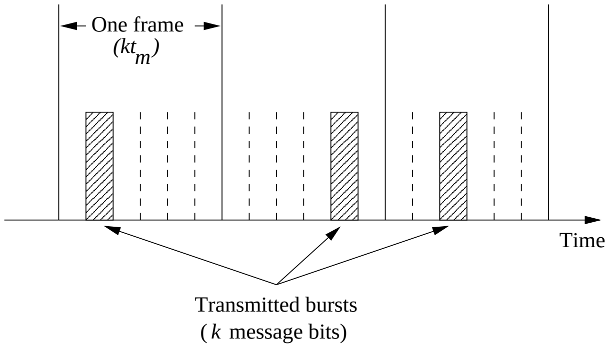

Figure 1.11: A typical time hopping waveform

If we denote the frame duration by T_f, let k be the number of messages bits in one frame, and $T_f = kt_m$, then the width of each time slot is $\frac{T_f}{M}$, and the width of each bit in the time slot is $\frac{T_f}{kM} = \frac{t_m}{M}$. This indicates that the transmitted signal bandwidth is $2M$ times the message bandwidth, and hence the processing gain of a time hopping system is simply twice the number of time slots in each frame when BPSK modulation is used, and half this when QPSK modulation is used.

Interference between simultaneous users in a time hopping system can be minimized by co-ordinating the times at which each user is allowed to transmit. This also has a secondary implication, in that it avoids the possibility of the near-far problem occurring.

1.5.2 Chirp

A spread spectrum system using chirp modulation varies the frequency of the carrier in a linear fashion to spread the bandwidth. Linear frequency modulation is a technique very common in radar systems, and is occassionaly used for communications systems. Assume that T is the duration of a given signal waveform, and B is the bandwidth over which the frequency is varied. In this case, the processing gain is given simply by the product BT.

1.6 The Importance of Processing Gain

One of the most important aspects of any spread spectrum communications system is it's processing gain. This was defined in Section 1.2 as the ratio of the PN code clock rate to the input data rate, for a direct sequence system.

In this example, and in all following examples in this chapter, a 10 bit, maximal length, pseudo random sequence generator has been used. This generator has a code length of $2^{10} - 1 = 1023$ bits, which is sufficient for the simulation purposes being discussed. As explained in Chapter **??**, a 13 stage Gold code generator is ideally suited for CDMA networks, but although such a sequence generator is readily constructed in hardware, the software simulation time would be excessive. The results obtained using a 10 bit, maximal length generator are, for the purposes of this discussion, entirely adequate.

Figure 1.12 shows the effect on the transmitted spectrum as the processing gain is varied. The input data was BPSK modulated onto a 10MHz carrier frequency, and PN clock rates of 10, 50, 100 and 1000 times the data rate have been shown. It is clear that the lower orders of gain do not result in a desirable spectrum, and as outlined previously, the spectrum of the main lobe starts to become clearly defined as the PN clock rate approaches 100 times the data rate. At 1000 times the data rate, a very clean and well defined spectrum is seen. In order to draw a comparison between these diagrams, the RF bandwidth has been keep constant. Since the bandwidth of the main lobe is directly proportional to the PN clock rate, it would be expected that the width of this lobe increases as the clock rate increases. However, this would make a direct comparison more difficult, and so in order to allow this, the data rate has been changed accordingly.

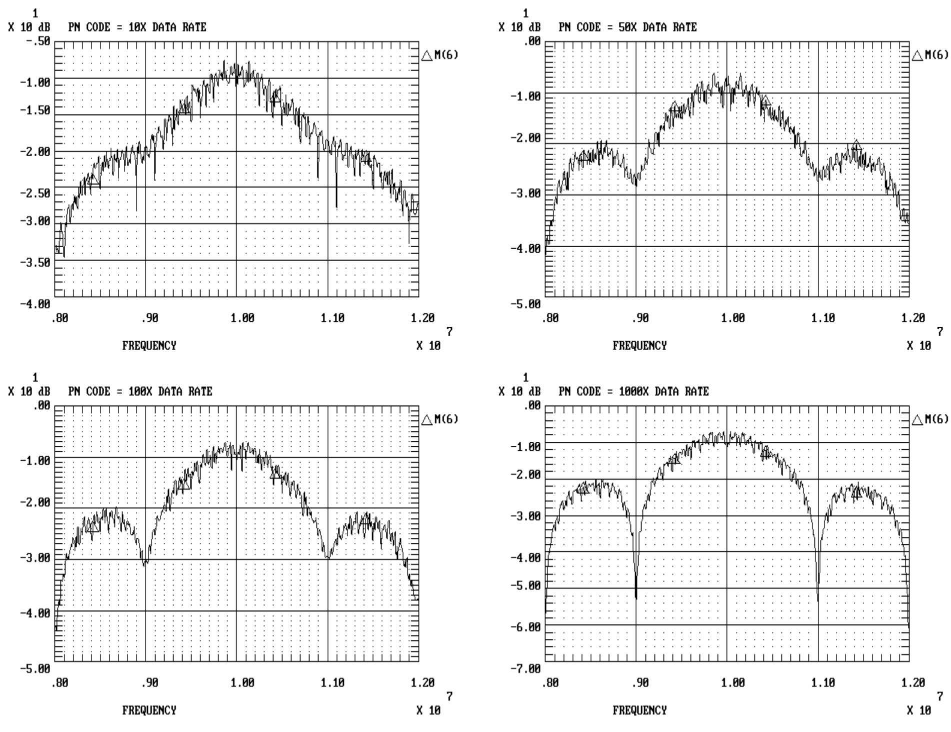

Figure 1.12: The effect of varying the processing gain

1.7 CDMA Networking

A Code Division Multiple Access (CDMA) system generally consists of a network of several spread spectrum transmitters and a single receiver. An often quoted example is the GSM mobile phone network, with a common base station controller and many mobile transceivers. In CDMA, the users access the channel in a random manner. Hence, the signal transmissions among the multiple users overlap both in time and in frequency. The separation and demodulation of the required signal at the receiver is facilitated by the fact that each signal is spread in frequency by a known pseudo-random sequence.

In a system employing CDMA, each user transmits a pseudo-random signal having a bandwidth W with average power P. The total capacity of the network varies, depending on the level of cooperation among the K users. At one extreme is non-cooperative CDMA in which the receiver chooses to ignore the spreading code from the unwanted transmitters. In this case, the other user's signals appear as interference at the receiver of each user and the multiuser receiver consists of a bank of K single-user receivers. Consider the case where each user's spreading code waveform is Gaussian. In this situation, each user's signal is corrupted by Gaussian interference of power $(K-1)P$ plus AWGN of power WN_0, where W is the additive Gaussian noise channel bandwidth, and N_0 is twice the power spectral density of the additive noise. Thus, the capacity per user is given by

$$C_k = W \log_2[1 + \frac{P}{WN_0 + (K-1)P}] \tag{1.5}$$

On the other hand, consider a network where the K users cooperate by transmitting synchronously in time, and the multiuser receiver knows the spreading waveforms of all the users, and jointly detects and demodulates all the user's signals. Here, each user is allocated a rate R_i, where $1 \leq i \leq K$, and a codebook containing a table of 2^{nR_i} codewords having power P. During each signal interval, a user chooses an arbitrary codeword and all users transmit their codewords simultaneously. The achievable rate region for the K users, assuming equal power for each user, in an AWGN channel is given by

$$R_i < W \log_2(1 + \frac{P}{WN_0}), 1 \leq i \leq K \tag{1.6}$$

[.....compare the two....]

1.7.1 CDMA Channel Models

In the following sections, the demodulation and detection of multiuser CDMA signals will be considered. It will be shown that the optimum maximum-likelihood detector has a computational complexity that grows exponentially with the number of users and that such a high complexity serves as a motivation to devise suboptimum detectors having lower complexities.

Consider the case of K users sharing a CDMA channel simultaneously. Each user is allocated a unique signature waveform $g_k(t)$ of duration T, where T is the inter-symbol period. This signature waveform may be expressed as

$$g_k(t) = \sum_{n=0}^{L-1} a_k(n)p(t - nT_c), 0 \leq t \leq T \tag{1.7}$$

where $\{a_k(n), 0 \leq n \leq (L-1)\}$ is a pseudo random sequence consisting of L chips that take values $\{\pm 1\}$, $p(t)$ is a pulse of duration T_c, where T_c is the chip interval. Thus, the code consists of L chips per symbol and $T = LT_c$.

The cross-correlations between pairs of signature waveforms play an important role in the operation of the signal detector.

[.....expand on this....]

1.7.2 An Optimum Receiver for CDMA Systems

An optimum receiver is one which is defined such that the receiver selects the most probable sequence of bits $\{b_k(n), 1 \leq n \leq N, 1 \leq k \leq K\}$ given the received signal $r(t)$ when observed over the time interval $0 \leq t \leq NT + 2T$. An optimum receiver can be devised for both synchronous and asynchronous transmissions.

[.....this whole section (up to The Near-Far Problem) needs to be rewritten....]

Synchronous Transmission

In synchronous transmission, each user produces exactly one symbol which interferes with the desired symbol. In AWGN, it is sufficient to consider the signal received in one signal interval, say $0 \leq t \leq T$, and determine the optimum receiver. Hence, $r(t)$ may be expressed as

$$r(t) = \sum_{k=1}^{K} \sqrt{\varepsilon_k} \cdot b_k(1)g_k(t) + n(t), 0 \leq t \leq T \tag{1.8}$$

Asynchronous Transmission

In this case, there are exactly two consecutive symbols from each interferer that overlap a desired symbol. It is assumed that the receiver knows the received signal energies $\{\varepsilon_k\}$ for the K users and the transmission delays $\{\tau_k\}$. Clearly, these parameters must be measured at the receiver or provided to the receiver via some control channel.

[.....expand....]

Using a block processing approach, the optimum maximum likelihood detector must compute 2^{NK} correlation metrics and select the K sequences of length N that correspond to the largest correlation metric. Clearly, such an approach is much too complex computationally to be implemented in practice, especially when K and N are large. An alternative approach is maximum likelihood sequence estimation employing the Viterbi algorithm. In order to construct a sequential-type detector, it is necessary to use the fact that each transmitted symbols overlaps at most with $2K-2$ symbols. Thus, a significant reduction in computational complexity is obtained with respect to the block size parameter, N, but the exponential dependence on K cannot be reduced.

It is apparent that the optimum maximum likelihood receiver employing the Viterbi algorithm involves such a high computational complexity that its use in practice is limited to communication systems where the number of users is extremely small, for example, $K < 10$. For larger values of K, a sequential-type detector is more appropriate.

1.7.3 Suboptimum Detectors for CDMA Receivers

In the previous section, it was observed that the optimum detector for K CDMA users had a computational complexity that grows exponentially with K. In this section, suboptimum detectors with computational complexities (measured by the number of arithmetic operations) that grow linearly with the number of users will be discussed.

Conventional Single User Detector

In a system employing a conventional single user detection method, the receiver for each user consists of a demodulator that correlates the received signal with the signature sequence of the user and passes the correlator output to the detector. Thus, the presence of the other users is neglected, or equivalently, the conventional single user detector assumes that the aggregate noise plus interference is Gaussian.

Consider a system using synchronous transmission. In this case, the output of the correlator for the kth user for the signal in the interval $0 \leq t \leq T$ is given by

$$15 - 3 - 26 = 15 - 3 - 27 \tag{1.9}$$

It is obvious that if the signature sequences are orthogonal, the interference from the other users given by the middle term in Equation **??** disappears and the conventional single user detector is optimum. However, if one or more of the other signature sequences are not orthogonal to the user signature sequence, the interference from the other users can become excessive if the power levels of the signals of one or more of the other users is sufficiently larger than the power level of the kth user. This situation is generally referred to as the *near-far problem* and will be discussed in a later section.

In asynchronous transmission, the conventional detector is more vulnerable to interference from other users. This is because it is not possible to design signature sequences for any pair of users that are orthogonal for all time offsets. Consequently, interference from other users is unavoidable in asynchronous transmission with the conventional single user detection. In such a case, the near-far problem resulting from unequal power in the signals transmitted by the various users is particularly serious. One practical solution requires a power adjustment method that is controlled by the receiver via a separate channel.

A Decorrelating Detector

[.....expand....]

Minimum Mean-Square-Error Detector

The multiuser MMSE detector can be implemented by employing a tapped delay-line filter with adjustable coefficients for each user and selecting the filter coefficients to minimise the MSE for each user.

Characterising the Performance of a Multiuser Detector

In multiuser communications system, the bit error probability is generally cited as a desirable performance measure. In evaluating the effect of multiuser interference on the performance of a detector for a single user, the probability of a bit error for a single user receiver in the absence of other channel users is often used. This is given by

$$P_k(\gamma_k) = Q(\sqrt{2\gamma_k}) \tag{1.10}$$

where $\gamma_k = \varepsilon_k/N_0$, ε_k is the signal energy per bit and N_0 is twice the power spectral density of the additive noise. In the case of the optimum detector for either synchronous or asynchronous transmission, the probability of error is extremely difficult and tedious to evaluate. In this case it is possible to use Equation 1.10 as a lower bound and the performance of a suboptimum detector as an upper bound.

1.7.4 The Near-Far problem

The near-far problem occurs when a signal from one transmitter arrives at the receiver with a much greater signal strength than any other transmitter. This may be due to an increased RF power level, or because of a closer proximity to the receiving antenna [20]. The apparent increase of signal strength can effectively jam the receiver, thereby locking all other users out of the network. This type of problem is especially relevant in mobile communications systems where many independent transmitters are free to move closer to, or further from, the receiver [21].

Consider a CDMA network comprising of binary direct-sequence transmitters. At a particular receiver, the multiple-access interference signal from transmitter j has the form

$$i_j(t) = \sqrt{2P_j}m_j(t)s_j(t)\cos(\omega_0 t + \theta_j) \tag{1.11}$$

where m_j is the data modulation, $s_j(t)$ is the spreading waveform, P_j is the received power, and ω_0 is the same for all transmitters in the network.

Assume a CDMA network capable of sustaining 200 independent transmitters, all transmitting with equal power levels. If one transmitter were to move from a distance d, to a closer distance at $d/4$, and assuming that propagation loss varies as the fourth power of the range, then this transmitter has effectively increased it's RF signal level by 24dB. Such an increase in signal level at the receiver would be sufficient to isolate any other transmitter, thereby losing any benefits of code division multiplexing, and reducing the network to a single transmit-receive pair. It is often necessary to employ some type of adaptive power-level feedback to ensure this type of non-deliberate jamming does not occur [22].

1.8 The Effects of Noise

This section looks at the effects of introducing Additive White Gaussian Noise (AWGN) into the communications channel. This can be used to simulate quiet atmospheric interference, as well as the effects of other co-channel users who are using properly correlated PN codes. Atmospheric interference varies markedly in it's spectral representation, ranging from slight AWGN through to major bandwidth pollution. The results presented here are not meant to account for all possible types of atmospheric interference, since this type of analysis can be found elsewhere [8]. It was shown by Weber *et al.* [38] that, under certain conditions, the effects of other simultaneous spread spectrum transmissions can effectively be modeled as a Gaussian random variable. As such, the simulations presented here are very relevant for CDMA networking, and although they use a discrete noise generator, this can be thought of as one or more separate transmitters using a properly correlated PN code.

The system used to analyse the introduction of white noise into the system is shown in Figure 1.13. The left hand side of this block diagram consists of a BPSK transmitter, modulating a 10MHz sinusoidal signal with a 10 stage (1023 bit) PN sequence. This signal is then mixed with an 800MHz carrier. Node 6 represents the transmitter output. From here, AWGN is added to the signal at various amplitudes, before entering the receiver. Both the carrier frequency and PN code phase are assumed to be synchronous, and this is simulated simply by using the same oscillators as the transmitter. The receiver de-spreads and down-converts the signal, which is then low pass filtered at 20MHz. The recovered signal can be seen at node 12.

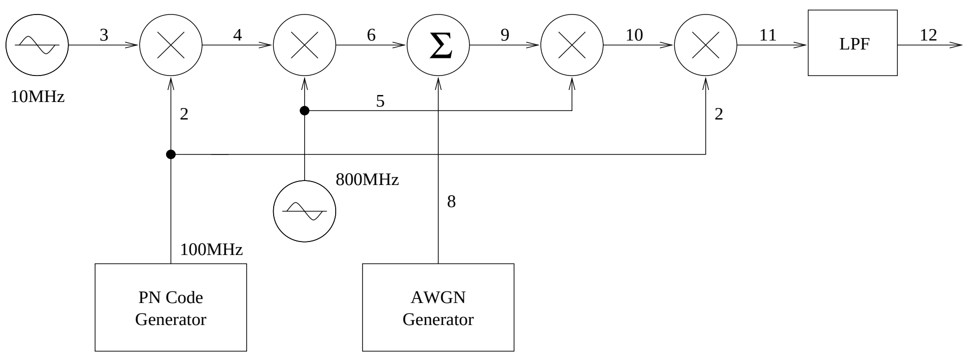

Figure 1.13: System diagram showing the addition of AWGN

Four levels of noise were introduced into the system, and the spectra for these simulations are shown in Figure 1.14.

The first system uses a noise level significantly lower than the signal amplitude, and as would be expected, does not have any noticeable effects on the system. As the noise level is increased to just below the first major null, then midway into the main lobe, a degradation in the received signal level becomes apparent. Finally, as the noise level is increased such that it exceeds the transmitted signal level, a further reduction in received signal level is observed. Note that as the noise level increases, the noise floor at the receiver also increases. These results are summarised in Table 1.1.

What is particularly noteworthy here, is that a signal can still be recovered with excellent phase and amplitude characteristics, even when the noise level exceeds the signal level. The concept of interference spreading was explained in Section 1.2, and this is shown very clearly here. The noise spectrum is spread over a wide bandwidth, allowing the desired narrow band signal to be recovered.

Table 1.1: Measured signal levels with AWGN interference

Transmitter Signal Power (dB)	Channel Noise Power (dB)	SNR at the Receiver	Recovered Signal Power (dB above the noise floor)
-25	-72	***	Δ45
-25	-52	***	Δ45
-25	-32	***	Δ30
-25	-22	***	Δ20

1.9 VLSI, FPGA & ASIC Implementations

This section gives a brief overview of current research into reprogrammable gate array and VLSI implementations of spread spectrum circuitry.

A finite state machine generating an oversampled signal from a binary data stream is proposed by Pagden in [23]. This state machine is able to successfully spread the spectrum of the input signal using a processing gain of 12. The design is implemented in VHDL, and is achieved using approximately 50 lines of source code.

Duncan [24] discusses a reconfigurable correlator implemented using a Xilinx XC6200. This correlator uses a pipelined design, and although specifically targeted at image correlation using a 32x16 pixel mask, the underlying theory is still relevant. As with PN code correlation, image correlation uses the idea of matching pixels instead of bits, and when the number of matching pixels in an image exceeds a preset threshold, a hit is detected. This corresponds to a synchronised bit stream in a spread spectrum receiver.

A matched filter for a digital spread spectrum modem is discussed in [26]. This matched, finite impulse response, filter is designed such that the input filter on the receiver side is identical to the output filter on the transmitter side. It is implemented using $1.2\mu m$ CMOS technology, and packaged in a 68 pin grid array package. It is capable of operating at a clock frequency of 65 MHz, with an output data rate of 65 MWords/s.

A VLSI implementation for a correlator/demodulator chip suitable for direct sequence, spread spectrum operation is presented by Zimmermann and Neeracher in [27]. This design integrates the code correlation and data demodulation of a RAKE receiver into a single package. It allows the use of either BPSK or QPSK encoding, with PN code lengths of between 15 to 1023 chips, at a clock rate of up to 16 MHz.

Lingwood *et al.* [28] presents two ASIC designs for a spread spectrum wireless local area network. The first implements a matched filter, performing square root raised cosine filtering, while the second calculates complex correlations and coherent BPSK/QPSK demodulation. A PN code rate of 20 Mchips/s is cited as being typical, with a resulting bit rate from 16 kb/s to 2 Mb/s. Both devices are clocked at 65 MHz.

A single ASIC CDMA digital receiver for space applications is proposed in [29]. It is assumed that the receiver is used by a low earth orbit satellite, and Doppler effects are taken into account during the acquisition phase. QPSK modulation is used, with Gold code and maximal length sequences operating from 250 bps to 300 kbps.

A design for a code phase shift keying spread spectrum receiver is presented by Chan and Leung in [30]. An FPGA is used to implement a baseband code PSK, M-ary decoder using a double threshold detection scheme. An analog IF and demodulator stage are constructed using discrete components and interfaced to the FPGA, with carrier recovery being performed by a Costas loop.

Kong *et al.* [31] presents a fast serial Viterbi decoder ASIC for CDMA cellular base stations. Using 0.8μm CMOS technology, a variable decoding depth Viterbi decoder is implemented with four add-compare-select pairs and internal static RAM to reduce the decoding time and to store path metrics.

A comparison of arithmetic architectures for Reed-Solomon decoders implemented in reconfigurable hardware is discussed in [32]. Using a design entry based on a VHDL description, several architectures are mapped to popular FPGA and EPLD devices. For each mapping, an area and a speed optimisation are performed. It is shown that composite field architectures can have great advantages using both types of reconfigurable platforms.

*** Add more relevant and up-to-date research (eg - Nova Eng)

1.10 Conclusion

This chapter has introduced the basic concepts of spread spectrum communications systems. Direct sequence, frequency hopping and hybrid systems were discussed, and examples of typical spectra were presented. The relevance of a system's processing gain was analysed, and magnitude plots were presented showing the effect on a transmitted signal as the processing gain was varied.

BEGIN EDIT

Brief mention was made of code division multiple access systems, and the near-far problem was outlined.............

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Varying levels of AWGN were introduced into the system, and comparisons were made for different transmitted signal to noise ratios. It was shown that a DS SS receiver is capable of recovering a desired signal even when the introduced noise exceeds the magnitude of the transmitted signal. To conclude the chapter, a brief discussion of current research into the use of re-programmable logic devices such as FPGAs and ASICs, as used in the field of spread spectrum communications, was presented.

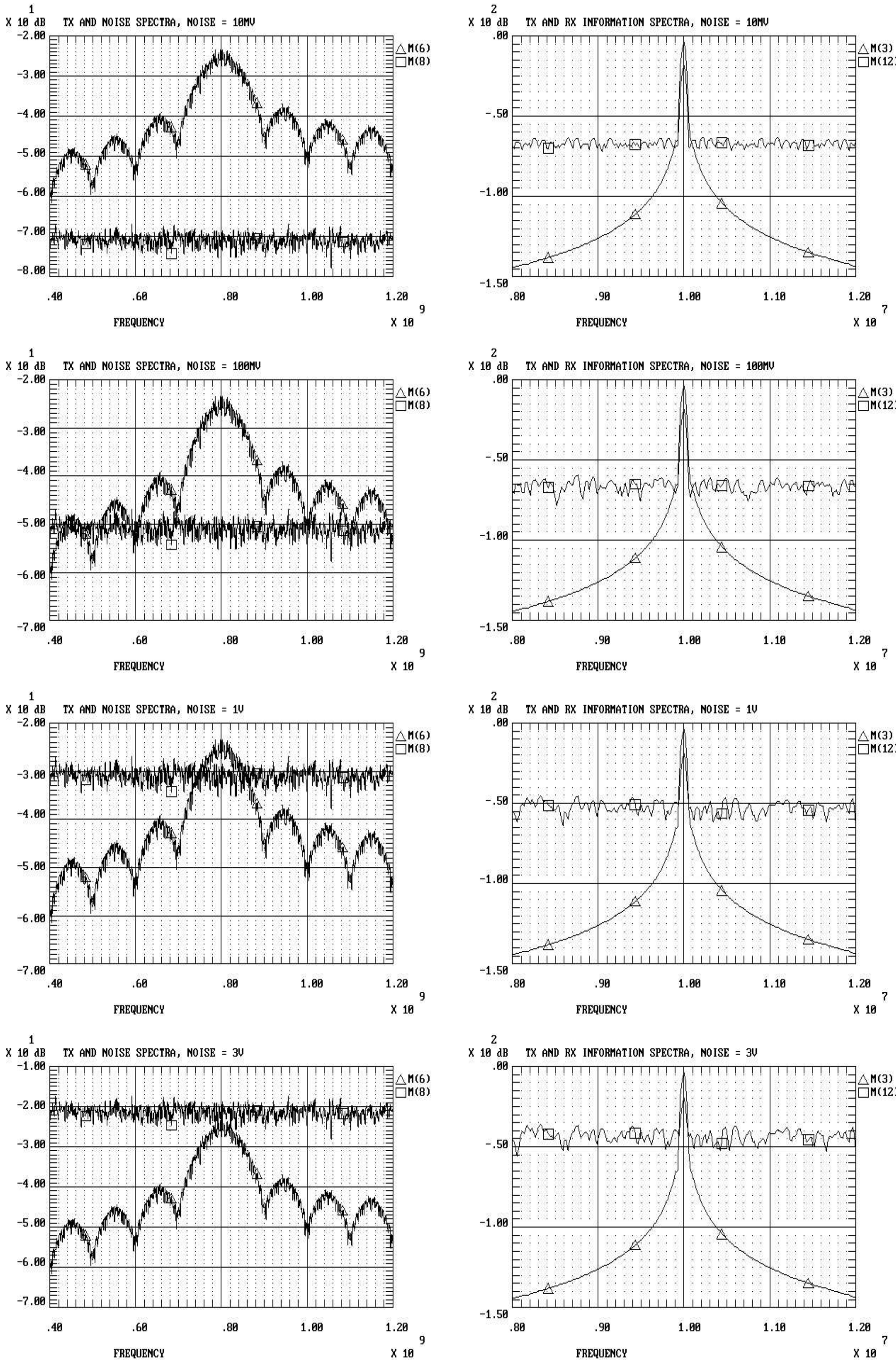

Figure 1.14: Transmitted spectrum with recovered signal under the influence of varying noise levels

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