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third-generation cdma systems for enhanced data services

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Third-Generation CDMA Systems for Enhanced Data Services

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Preface

CDMA has been quite successful as a second-generation cellular system, having achieved widespread use in particular in North America and Korea by the turn of the twenty-first century. As the new century begins, CDMA systems will once again find widespread use in the form of third-generation cellular systems. However, two third-generation systems based on CDMA, cdma2000 and WCDMA, have emerged. As a result, many people both inside and outside the wireless industry have a distinct desire to actually be able to distinguish between these two systems and study the relative features of each.

Moreover, work was initiated within the standardization bodies responsible for cdma2000 and WCDMA (namely the 3GPP2 and 3GPP respectively) to evolve these third-generation CDMA systems to provide improved services for cellular packet data users. This effort has resulted in the 1X-EV system for cdma2000 and HSDPA for WCDMA.

This book is intended to give the reader an overview of the CDMA systems presently being deployed and used for second- and third-generation cellular telephony. The authors hope that any reader, after reading this text, will not only be able to determine exactly what distinguishes second-generation CDMA systems such as IS-95 from their third-generation counterparts, but also what distinguishes cdma2000 from WCDMA.

In addition, it is intended that the reader come away with a better understanding of the evolution of WCDMA and cdma2000. As packet data services are expected to play a more prominent role in cellular telephony during the first decade of the twenty-first century, these technologies will become more important.

Chapter 1 of the text provides an introduction to cellular telephony and the various technologies available for cellular multiple-access communications; it also provides a description as to what technologies are used in first-generation, second-generation and third-generation systems. Chapter 2 provides the basic theory behind spread spectrum communications and CDMA systems. Chapter 3 introduces concepts of the mobile wireless channel and CDMA receiver theory for diversity reception. Chapter 4 provides an overview of the second-generation IS-95 CDMA system in the first half, and its third-generation evolution cdma2000 in the second half. Chapter 5 discusses the 1X-EV evolution technology for cdma2000. Chapter 6 provides an overview of the WCDMA system. Chapter 7 provides performance comparisons between IS-95, cdma2000, 1X-EV and WCDMA. Chapter 8 discusses handover in all of these CDMA systems. Finally, Appendix A provides a detailed analysis of CDMA transceiver design, using an IS-95 handset as an example.

The authors would like to thank many of our Nokia colleagues for contributing to this work. In particular, the authors would like to thank Alan Hsu, Mark Cheng, and Craig Greer for reviewing the manuscript and providing many insightful suggestions. In addition, the authors would like to thank Zhigang Rong, Lin Ma, Petteri Luukkanen, and Zhoyue Pi for many of the results provided in Chapter 7. The authors would also like to thank Tom Derryberry and Mitch Tseng for much of their insight into the current standardization status of CDMA systems, and to Alan Brown for his support and encouragement in writing this book. The authors would also like to acknowledge Joel Claypool from Academic Press and Prof. Jerry Gibson from Southern Methodist University for their encouragement and advice in preparing the text. Finally, the authors would like to thank their wives, Chitra Mandyam and Ying-Erh Lee, for their support and patience.

Chapter 1

Introduction to Cellular Systems

Key Topics: Cell, mobile station, base station, PSTN, BSC, MSC, cluster, FDMA, TDMA, CDMA

Overview: An overview of cellular telephony systems is provided in this section. A brief history of cellular telephony is provided, and a basic description of a cellular system and its underlying components is given.

1. INTRODUCTION

Cellular telephony systems are radio systems that involve distributed transmission. Therefore, rather than having a single transmitter service many different users over a wide area of coverage (e.g., commercial FM radio), the coverage area is divided into smaller areas known as *cells*. Each cell has one stationary transceiver known as a *base station*.

A user of a cellular system communicates with the base station to place a call. The call can be data or voice, and the base station routes the call to either a terrestrial network to the termination point or to another user of the same cellular network. Normally, for voice calls, the base station either directly or indirectly routes the call to a *public switched telephony network* (PSTN).

Each user of a cellular system is also sometimes called a *subscriber*. The basic relationship between a subscriber and the base station is shown in Figure 1-1. The communications link from the base station to the subscriber is referred to as the *downlink* or *forward link*, while the link from the subscriber to the base station is referred to as the *uplink* or *reverse link*.

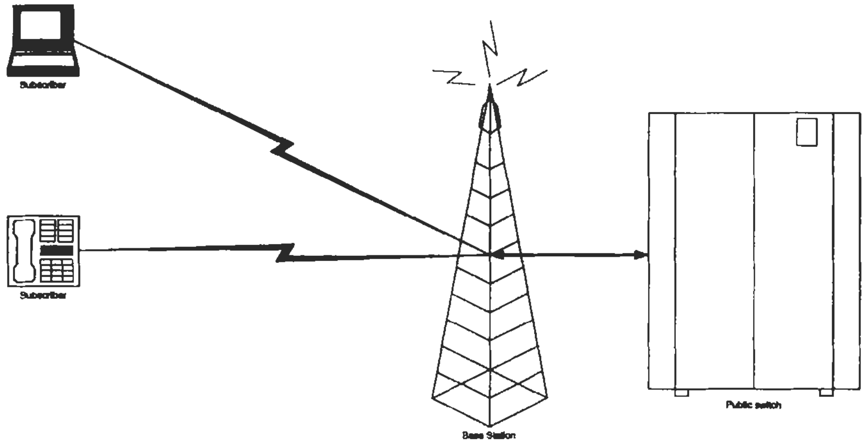

Figure 1-1: Subscriber/Base Station Link

Cellular subscribers can be stationary or mobile. If the subscriber is mobile, then the cellular network must be able to handle the situation in which a mobile subscriber (also known as a *mobile station*) moves from one cell to another. This event is known as *handoff* or *handover*. If the mobile station can engage in simultaneous communication with multiple base stations, then it is said to be in *soft handoff*.

In order to ensure that a call is not dropped when a handoff occurs, information about the mobile station is usually known to the base stations involved in the handoff. Due to this and for other reasons, some communication exists in the network that connects base stations together in a cellular system. This network is known as the *backbone network* or simply the *backhaul*.

The backbone network consists of several entities between the PSTN and the base station. The base station usually interfaces with a *base station controller* (BSC), which networks a cluster of base stations to ensure that call admission and handover can function in a coordinated manner among base stations within a geographical region. A *cluster* is a group of cells that use the complete set of available telephony channels in a cellular network.

One or more BSCs are usually connected to a *mobile switching center* (MSC), which interfaces directly with the PSTN. The MSC contains information about the cellular subscriber that can be used to route other information to that user during the call. Moreover, a *home location register* (HLR) may be co-located with the MSC; this entity contains user-specific information used primarily for authentication of the subscriber during call initialization. The intercommunication between the mobile station, base station, BSC and MSC is shown in Figure 1-2.

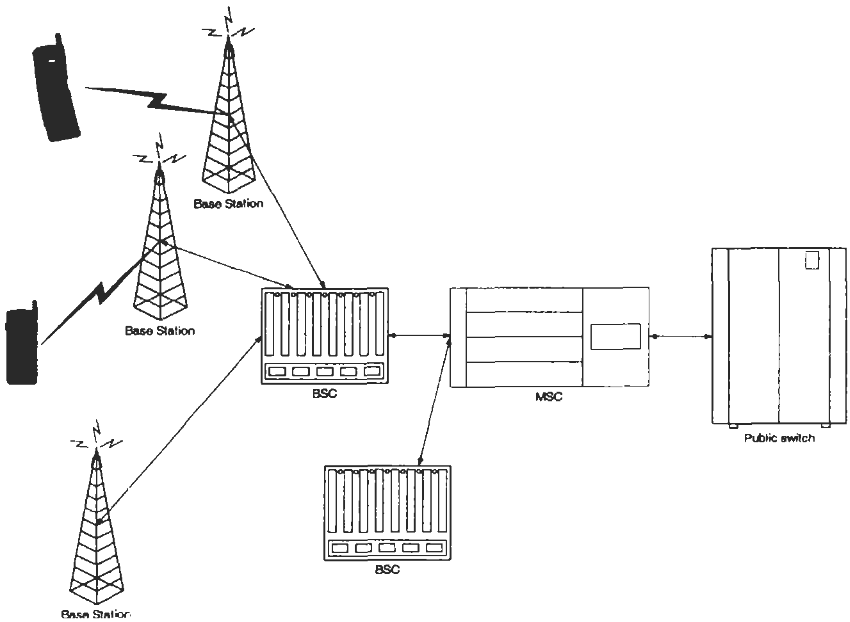

Figure 1-2: Mobile Station/BSC/MSC Architecture

A cellular network is comprised of many cells in a particular geographical arrangement. Often, a base station will use a different frequency (or frequencies) for communication than the base stations in neighboring cells use. This gives rise to the *frequency reuse factor*, which is the minimum number of frequencies needed for a given cellular network to ensure that interference within the same frequency (i.e., *co-channel interference*) is below a tolerable level. For analysis' sake, cells are often represented as hexagonal in shape to describe an ideal cellular network (see Figure 1-3). This type of representation often results in a frequency reuse factor of 7 (see Figure 1-4), as that is the minimum number of frequencies needed to ensure that no neighboring base stations have to occupy the same frequency.

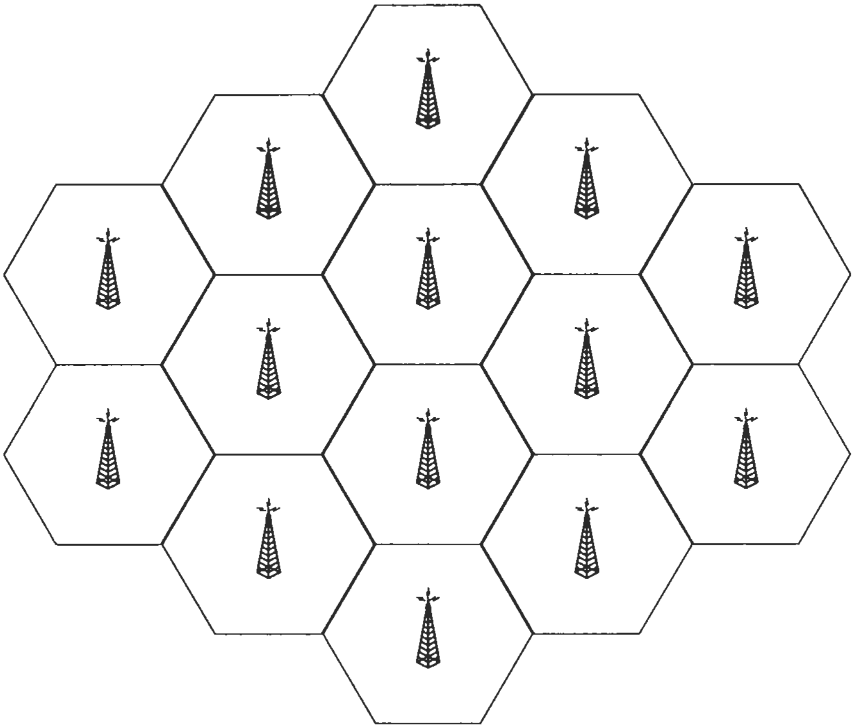

Figure 1-3: Hexagonal Cell Representation

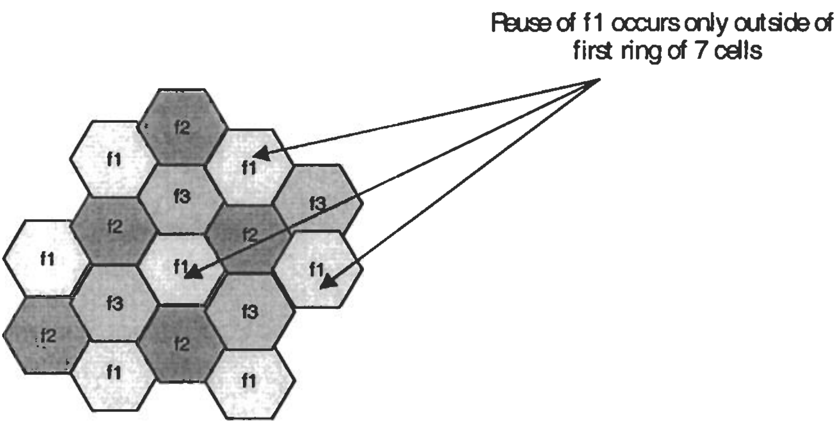

Figure 1-4: Frequency Reuse Factor of 7

The hexagonal type of representation is adequate for preliminary analysis of a network. In reality, several factors such as terrain, restrictions on base station deployment, coverage holes, and high-density (in terms of number of subscribers) areas will most likely prevent a uniform hexagonal cell structure.

2. MULTIPLE-ACCESS CELLULAR COMMUNICATIONS

Multiple-access communications are critical for a cellular system to be commercially viable. The term *multiple access* refers to the fact that multiple users are able to use a cellular system simultaneously. Multiple-access wireless systems normally can be classified into three categories:

1. Frequency division multiple access (FDMA)
2. Time division multiple access (TDMA)
3. Code division multiple access (CDMA)

This is not to say that other types of multiple access systems for cellular communications do not exist. However, these three categories encompass the evolution of almost all cellular systems around the world.

2.1 Frequency Division Multiple Access

FDMA systems formed the basis for the first widely deployed public cellular systems in North America. In particular, the Advanced Mobile Phone System (AMPS) ([1],[2]), developed primarily by AT & T Inc. was initially deployed in North America with a small rollout in Mexico City in 1981. The first United States deployment was in the Chicago area in 1983, marking the beginning of a nationwide rollout of cellular services for the public. This system was deployed in the 800 MHz band ("cellular band") using 30 kHz channel spacing that still exists in this band today.

ETACS (European Total Access Communication System) was deployed in Europe, with the slight difference from AMPS in that the channel spacing was 25 kHz. Similarly, N-AMPS (Narrowband AMPS) was developed by Motorola to work within a 10 kHz channel spacing, thus increasing the AMPS capacity. These initial FDMA systems normally are referred to as "first-generation" cellular systems.

FDMA systems generally operate with each base station in a cluster of cells occupying a separate frequency in both the transmit and receive bands (limited frequency reuse factor), and with each frequency band servicing only one cellular subscriber. A high-level example of the spectrum allocation for an FDMA system is given in Figure 1-5.

These first-generation systems based on FDMA are primarily analog systems, such as AMPS. This does not mean that it is not possible to carry digital information on an FDMA carrier, however. In fact, the AMPS system transmits voice in an analog format but the control information is transmitted in a digital format in a *blank and burst* manner (meaning that the receiver is receiving either voice or control information but never both simultaneously).

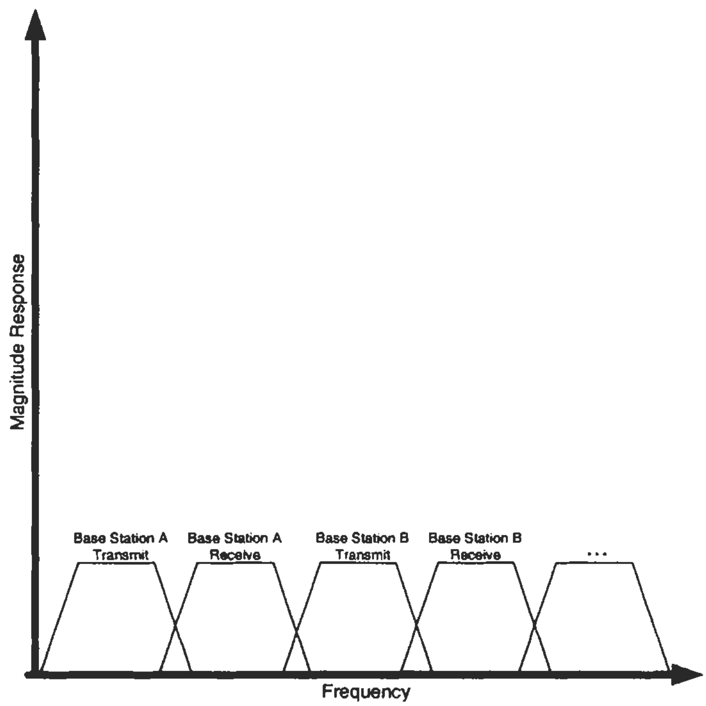

Figure 1-5: FDMA Cellular Communications Spectrum Allocation

FDMA systems have inherent capacity limitations due to the fact that each spectral channel can be allocated to only one user. Therefore, assuming a base station has only a single transmit and single receive frequency, it can only accommodate a single user. In addition, due to the limitations of the frequency reuse factor in a given deployment, co-channel interference is an issue.

2.2 Time Divison Multiple Access

In the early 1980s, cellular operators and wireless equipment vendors around the world recognized the capacity limitations of FDMA-based analog cellular systems. There was a concern that with the growing popularity of cellular services, systems such as AMPS would not be able to efficiently meet the demand. AMPS was a system conceived in the mid-1960s, and at that time AMPS took advantage of the state-of-the-art technology for circuit design. However, two decades later, real-time digital signal processing in integrated circuits was possible. This meant that digital communications could become a reality for cellular systems.

This led to the development of the first digital cellular systems based on TDMA. In North America, there was a desire to remain compatible with the spectrum allocation already in existence in the cellular band for AMPS. As a result, the 30-kHz channel TDMA system known as US Digital Cellular (USDC) or Digital AMPS (D-AMPS) was developed in the late 1980s [1]. Also in the 1980s, the Groupe Special Mobile (GSM) in Europe developed a digital TDMA standard to work within 200 kHz channels. The first GSM deployments were in 1991, and the first D-AMPS deployments in North America were in Canada by CANTEL in 1992. These TDMA systems are also grouped under a general classification for initial digital cellular technologies known as "second generation."

TDMA cellular systems utilize spectrum in a similar fashion to FDMA systems, with each base station in a cluster occupying a distinct frequency from transmission and reception. However, each of the two spectral bands is also allocated in time to each user in a round-robin fashion. For instance, 3-slot TDMA divides transmission into three fixed time periods (slots), each of equal duration, with a particular slot assigned for transmission to (or from, in the case of the uplink) one of three possible users (see Figure 1-6). This type of approach requires tight synchronization between the mobile station and base station.

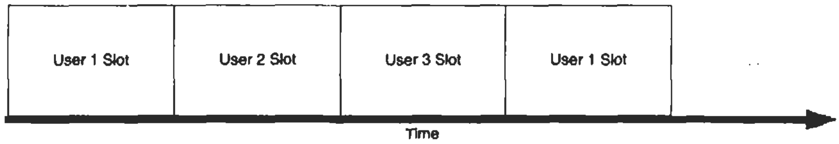

Figure 1-6: 3-slot TDMA

A simple rule of thumb is that the number of slots allocated in a TDMA system is also the number of times the capacity is increased when compared to an FDMA system with identical bandwidth. However, this does not always bear out in deployed networks, primarily due to differences in digital processing in TDMA systems versus analog processing in FDMA systems, including suppression of co-channel interference and mitigation of wireless channel effects (the wireless propagation environment will be examined in more detail in Chapter 3).

2.3 Code Division Multiple Access

In the mid-1980s, several researchers saw the potential for a technology primarily used in military applications to also be used for cellular communications. This technology, spread spectrum communications, which involve transforming narrowband information to a wideband signal for transmission, was seen as a mean of addressing potential capacity limitations of TDMA systems (which result from the fact that the number of users on any single frequency is restricted by the number of available time slots).

A spread spectrum system operates by transforming the narrowband information of an individual user into wideband information by using high-frequency codes, each unique for that particular user. By assigning different users unique codes, a multiple-access system is possible, i.e., *code division multiple access* (CDMA). Moreover, in a CDMA system, frequency reuse limitations seen in FDMA and TDMA systems are not quite so critical, as multiple mobile stations and base stations can occupy the same frequencies at once.

Qualcomm Incorporated in San Diego, California, developed the first CDMA cellular system for widespread deployment in the early 1990s, culminating with the standardization of Qualcomm's CDMA solution by the Telecommunications Industry Association (TIA) in 1992 [3].

More recently, CDMA has formed the basis for enhancing cellular systems around the world. As part of the International Telecommunications Union's (ITU) International Mobile Telephony-2000 (IMT-2000) project, third-generation systems were developed to improve wireless multimedia services to cellular subscribers. Three primary technologies were approved in 1998 by the ITU as third-generation technologies:

1. Wideband CDMA [4], developed by the European Telecommunication Standardization Institute (ETSI).
2. cdma2000, developed by the TIA. cdma2000 is backwards-compatible to IS-95.
3. EDGE (Enhanced Data rates for GSM Evolution), which was co-sponsored by ETSI and the TIA.

CDMA spread spectrum systems come in two types: frequency hopped and direct sequence. CDMA using frequency hopping involves a user transmitting over multiple frequencies consecutively in time in a *pseudorandom* manner. Pseudorandom in this case refers to the fact that the sequence of transmission frequencies is known at the transmitter and receiver, but appears random to any other receiver. An example of a frequency hopping sequence is given in Figure 1-7.

Slow-hopping systems involve a changing of frequencies at a slower rate than the information bit rate, whereas *fast-hopping* requires a much faster change of the transmission frequency than the information bit rate. Frequency hopped systems are limited by the total number of hopping frequencies available. If two users hop to the same frequency at once, they will interfere with one another.

Direct-sequence systems work by modulating the user's information signal with a sequence known to the receiver and transmitter. This sequence is generated at a much higher rate than the user signal, literally "spreading" the user's signal bandwidth. This process is illustrated in Figure 1-8. All commercial cellular CDMA systems use direct-sequence spreading as opposed to frequency hopping.

In the next chapter, more detail will be provided on the technical aspects of direct-sequence spread spectrum systems.

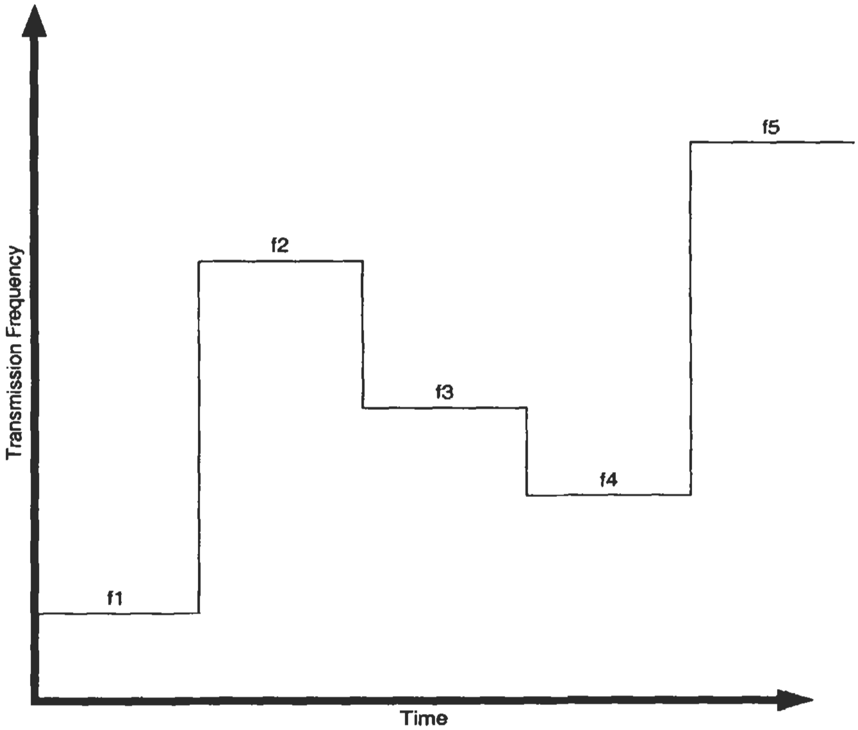

Figure 1-7: Frequency Hopping Sequence

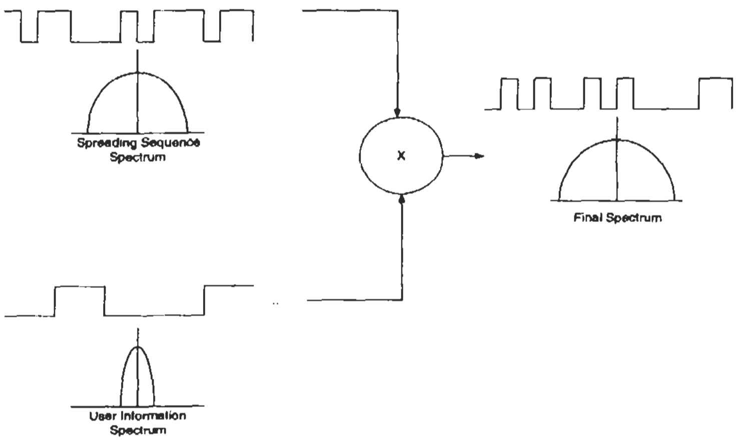

Figure 1-8: Direct-Sequence Spreading of Information

3. CONCLUSIONS

In this chapter, the cellular concept, which makes wireless multiple-access communications possible, was introduced. A brief overview of the different types of multi-access technologies available for wireless cellular systems was also provided. With this background, the reader is now prepared to understand the underlying direct-sequence spread spectrum technology that makes CDMA possible.

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Chapter 2

Direct-Sequence Spread Spectrum Systems

Key Topics: Spread spectrum, direct-sequence spread spectrum, PN sequence

Overview: This chapter will focus on fundamental principles covering the design of direct-sequence spread spectrum systems. The basic concepts covered here will be used in later chapters when describing existing CDMA wireless standards.

1. INTRODUCTION

Direct-sequence spread spectrum (DSSS) systems form the basis for the most widely used CDMA standards for public wireless systems today. These types of systems have their genesis in the significant amount of work performed in the area of secure communications during the early and middle parts of the 20th century; a detailed history may be found in [1] and [2].

In the 1930s, wideband frequency modulation was proposed as a means of wireless transmission of information without the information distortion associated with narrowband FM systems. In 1933, Edwin Armstrong proposed a system wherein the transmission bandwidth was spread beyond the information bandwidth in an FM system. He also proposed the use of an amplitude limiter at the receiver to eliminate the effects of amplitude variation that result from bandwidth spreading and mobile channel distortion. He collaborated with the Radio Corporation of America to verify his idea. This idea was followed by Gustav Guanella's invention in 1938 of continuous-wave radar, which described a transmission method over several different frequencies of smaller power compared to the total signal power. This work was performed for Brown, Boveri and Co. in Switzerland.

With the advent of World War II, the need for secure communication brought about an even greater examination of the benefits of wideband transmission. In particular, a definitive need existed for transmission methods that were inherently resistant to narrowband interference most likely caused by hostile parties. In addition to this "anti-jamming" capability, there was also a need to develop communications systems that were difficult to intercept. In response to this need, what was probably the first public disclosure of a spread spectrum communication system came about in 1941 with a patent by Hedy Keisler Markey (a.k.a. "Hedy Lamarr" of film) and George Anthiel. This patent described a frequency hopped guidance system for torpedoes.

The motivation for spread spectrum systems as a means of multiple-access communications is found in Claude E. Shannon's pioneering work in information theory. In particular, in 1948, Shannon derived the maximum channel capacity of a bandlimited communications system as

$$C = B\log_2\left(1+\frac{S}{N}\right) \tag{2-1}$$

In (2-1), C is the channel capacity (bits/second), B is the transmission bandwidth (in Hz), S is the received signal power (in watts), and N is the total noise power at the receiver. This contribution is important in that it provides a rigorous justification for increasing the transmission bandwidth in a communications system, as the capacity is directly related to the transmission bandwidth for a given signal-to-noise ratio. This idea gave impetus to the use of spread spectrum as a means of increasing the available capacity for wireless systems.

Spread spectrum remained a primarily military technology until the 1970s. In the late 1970s, Prof. George Cooper at Purdue University postulated that spread spectrum CDMA systems would be able to provide greater capacity than the AMPS systems that were under development at the time [3]-[4]. Shortly afterward, the Equitorial Communications Company developed a spread spectrum satellite receiver for commercial use [1].

Perhaps the most significant venture to date in the area of the commercialization of CDMA began in 1985 with the founding of Qualcomm Incorporated in San Diego, California, by Prof. Andrew Viterbi of the University of California at San Diego and Dr. Irwin Jacobs. Some of the most significant innovations at Qualcomm include the development of fast power control mechanisms for interference reduction in CDMA systems, and the use of soft handoff to improve network performance. Although Qualcomm's first products were spread spectrum communications systems for the transportation industry, by the mid-1990s Qualcomm technology was deployed in the United States and Asia for public cellular communications.

As was alluded to in the previous chapter, more recent work on third-generation systems has yielded new innovations for CDMA. These will be discussed in more detail in Chapters 4, 5, and 6.

2. PROCESSING GAIN

CDMA systems are inherently interference limited. This is due to the fact that several users occupy the same bandwidth simultaneously. As a result, the concept of *processing gain* is extremely important to CDMA systems, as this value provides a measure for the interference sensitivity for CDMA receivers.

The processing gain, with respect to the variables defined in (2-1), is

$$P_g = \frac{B}{R} \tag{2-2}$$

where R is the user bit rate. The processing gain is also sometimes known as the *spreading gain*. The processing gain is usually measured in decibels. For instance, in the IS-95 CDMA system, the user data rate for a single voice service is 9600 bits/second. Since the system occupies a 1.2288 MHz bandwidth in this case, the processing gain is $10 \log_{10}(1.2288 \times 10^6/9600) =$ 21.1 dB.

The processing gain may be used to determine the received bit signal-to-noise ratio. For instance, cellular systems normally suffer from two forms of interference: the interference from within the cell and the interference from outside the cell. If the transmitted signals appear to be random (more precisely, Gaussian) in nature, then the interference caused by these users starts to resemble thermal (background) noise. Assume that the interference to the received signal can be grouped into the term I_0. Then the receiver bit SNR (i.e., the ratio of the bit energy E_b to overall noise power N_0) may be shown to be

$$\frac{E_b}{N_0} = \frac{P_g P_u}{I_0} \tag{2-3}$$

where P_u is the user-transmitted in-band signal power. This relationship shows how the received bit SNR increases linearly with the processing gain.

The larger the processing gain of a CDMA system is, the larger the possible coverage of each cell in the CDMA system. This is due to the fact that the transmitted signal power P_u necessary to achieve a desired E_b/N_o becomes less as the processing gain increases. However, increasing the processing gain in a bandlimited system results in reducing the user information bit rate.

3. PSEUDORANDOM SEQUENCES

As mentioned in Chapter 1, direct-sequence CDMA systems make use of a spreading sequence known at the transmitter and receiver to spread the user information. It is desirable in a communications system to have this sequence resemble a random sequence as much as possible. This not only provides security to the transmission, but also results in the interference appearing more like thermal noise. The generation of such a sequence that can be reconstructed at the receiver can be performed through the use of *pseudorandom* sequence generators. These sequences are also known as *pseudonoise* (PN) sequences, due to their similarity to quantized background noise.

Pseudorandom sequences used for CDMA systems are normally binary to reduce complexity. These sequences are periodic in nature. Proper design of a pseudorandom sequence entails ensuring that the sequence has the following properties [4]:

1. The sequence has an approximately equal number of 1's and 0's in a given period.

2. Consecutive occurrences of 1's or consecutive occurrences of 0's occur with the probability of multiple coin flips. In other words, the probability of three 0's occurring in a row would appear with the probability of 1/8 (e.g., 3 consecutive coin flips of "heads").
3. The sequence is orthogonal with a time-shifted version of itself over one period. In other words, if one compares the sequence with the circular shift over one period, there would be equal numbers of bit positions in which the binary values agree and disagree.

It should be noted that criteria 1 and 3 above imply that the sequence is even-length over a period. In general, a pseudorandom sequence can also have odd length. In such a case, the relative probabilities of 1's and 0's in a period of the sequence aren't exactly equal to 1/2 but are very close. In addition, the sequence is not orthogonal to its shift over one period, but the number of bit positions in which the sequences disagree is almost equal to the number of positions in which they agree.

Pseudorandom binary sequences are typically generated through the use of shift registers with feedback. A shift register consists of several memory stages linked together consecutively along with feedback logic. The shift register operates with a master clock, and with each clock cycle a new output from the shift register is generated. A generic N-stage shift register is shown in Figure 2-1.

In Figure 2-1, the coefficients $\{d_i\}$ are also binary and simply act as switches: if the coefficient is '1', then the input is passed through. The content of memory stage i is represented by x_i. The state of the shift register is the collective values of the memory stage contents, i.e., $x_1x_2 \ldots x_N$. The feedback logic in this case consists of a modulo-2 addition of all the stages, depending on the associated values of d_i. As a result, for the shift register sequence to operate properly, the shift register state must be initialized to a state other than the all-zero state; otherwise the sequence produced will be all zeros as well.

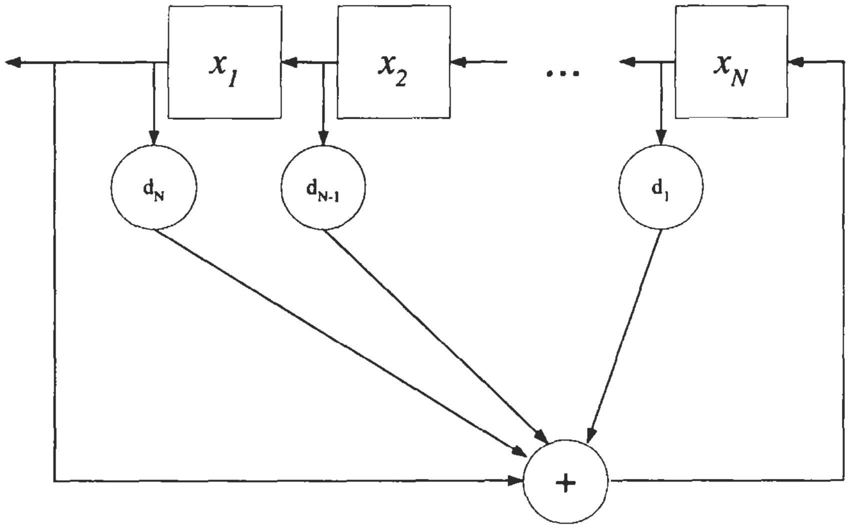

Figure 2-1: Shift Register with Feedback

If the output of the shift register sequence at clock sample index n is denoted as c_n, then this value can be represented as the modulo-2 sum of a combination of the future output values and the coefficients $\{d_i\}$ as

$$c_n = \sum_{i=1}^{N} d_{N-i+1} c_{n+i} \tag{2-4}$$

This representation leads to shift register usually being described in terms of a polynomial function of a delay variable X. Thus, the generator polynomial that describes the shift register sequence may also be represented as

$$G = \sum_{i=1}^{N} d_i X^i \tag{2-5}$$

where X^i represents a delay of i clock pulses. This polynomial is also known as the *characteristic polynomial* of the shift register sequence.

The sequence produced by the feedback shift register is periodic and will repeat after a certain number of samples from the register. If the sequence period is equal to $2^N - 1$ samples, then the sequence is of *maximal length*.

Pseudorandom binary sequences are normally of maximal length and are generated through the use of *maximal length shift registers* (MLSRs). These sequences have the properties previously mentioned for pseudorandom sequences, but also have the additional property that the modulo-2 addition of the output sequence with a shifted version of itself yields another shifted version of itself.

As mentioned previously, an important property of binary PN sequences generated from MLSRs is that they are orthogonal with shifts of themselves. If it can be assumed that the binary output values of the PN generator "0" and "1" can be mapped to -1 and +1, respectively, then the following holds true:

$$R(\tau)=\sum_{i=1}^{2^N-1} c_i c_{i-\tau} = \begin{cases} +1 \quad \text{or} \quad -1, \tau \neq 0 \\ 2^N-1, \tau = 0 \end{cases} \tag{2-6}$$

where $R(\tau)$ is sometimes referred to as the autocorrelation function. It should be noted that due to the odd-length of an MLSR PN sequence, it is not perfectly orthogonal with its own shifted version; if it were orthogonal, the autocorrelation function would be 0 for all $\tau \neq 0$. A correlator can easily be implemented in digital hardware using an XNOR gate with a summer as is shown in Figure 2-2.

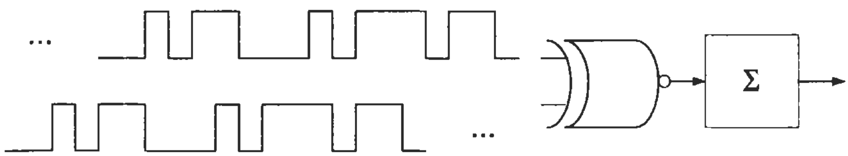

Figure 2-2: PN Correlator

An important property of PN sequences is that a time-shifted version of any PN sequence can be generated by forming a modulo-2 addition of the PN sequence with a delayed version of itself, i.e., the "delay-and-add" property [5].

Another important aspect of PN sequences is their *partial correlation* properties. In many instances, it may be desirable to correlate a PN sequence with a shifted version of itself over a duration shorter than the period of the sequence. In this case, PN sequences usually display behavior in which the partial correlation looks like noise unless the shift is 0.

The MLSR characteristic polynomial must be *irreducible*, meaning that it cannot be factored into lower-order polynomials. In fact, it can be shown that the number of irreducible polynomials of degree N is [5]

$$K_p(N) = \frac{2^N - 1}{N} \prod_{i=1}^{J} P_i \tag{2-7}$$

where J is derived from $\{P_i\}$, which represents the prime decomposition of 2^N-1, i.e.:

$$2^N - 1 = \prod_{i=1}^{J} P_i^{e_i} \tag{2-8}$$

In (2-8), e_i is a positive integer and P_i is a prime number. As a result, there is a limit on the number of PN sequences that can be generated of each length. The number of polynomials for some values of N is given in Table 2-1.

N	Length (2^N-1)	$K_p(N)$
2	3	1
3	7	2
4	15	2
5	31	6

Table 2-1: Number of Irreducible Polynomials for Different Orders

PN sequences with better cross-correlation can be generated by starting with two distinct MLSR-generated sequences of the same length. Then by taking the modulo-2 sum of one sequence with a cyclically shifted version of the other, a new code can be generated. Thus if the length of each sequence is 2^N-1, then 2^N-1 distinct sequences can be created in this way (one for each possible cyclic shift of the second sequence). These types of sequences are called *Gold sequences*, or Gold codes.

3.1 Shifting the PN-Sequence: The Use of PN-Masks

The delay-and-add property of PN sequences implies that temporal shifting of the PN sequence can be formed with simple logic operations involving the MLSR. This leads to the concept of a PN *mask*. A mask is a binary value with the same number of bits as there are delay elements in the shift register used to generate the PN sequence. The mask is AND'ed bitwise with the contents of each corresponding delay element, and a modulo-2 addition is performed on the results of these AND operations. The resulting sequence is a delayed version of the original PN sequence. An example of a PN-masking circuit is shown in Figure 2-3. The mask in Figure 2-3 can be represented as a binary number $m_1 m_2 \ldots m_N$, where m_i corresponds to the multiplicative weighting of the x_{N-i+1} delay element in the final masked output.

4. ORTHOGONAL CODES

While PN sequences are useful as spreading sequences, there are a limited number of them for any given code length as seen by Table 2-1. As a way to accommodate several different users simultaneously in a CDMA system, assigning each user a unique PN spreading code has limitations. Therefore, in order to increase the bandwidth efficiency in CDMA systems, different users are modulated using their own unique codes derived from orthogonal functions combined with PN spreading sequences.

A set of functions $\{f_n\}$ all of length N are mutually orthogonal if the following relationship holds:

$$\sum_{i=1}^{N} f_k(i) f_l(i) = \begin{cases} N, k = l \\ 0, k \neq l \end{cases} = N\delta_{k-l} \tag{2-9}$$

In (2-9), δ_r is known as the *dirac-delta* operator. This operator is 1 when $r =$ 0, or zero otherwise. If the summation in (2-9) is simply δ_{k-l} rather than $N\delta_{k-l}$, then the functions are *orthonormal*. Taking advantage of mutual orthogonality, a user can suppress signals intended for other users in a CDMA system while capturing his own.

Although there has been considerable work in the area of mutually orthogonal functions, one family of orthogonal functions which are particularly attractive for digital communications systems are *Walsh* functions. Walsh functions are derived from the rows of Walsh matrices (also known as a *Walsh-Hadamard* matrices). These matrices are square matrices whose dimensions are 2^r by 2^r, where r is a non-negative integer.

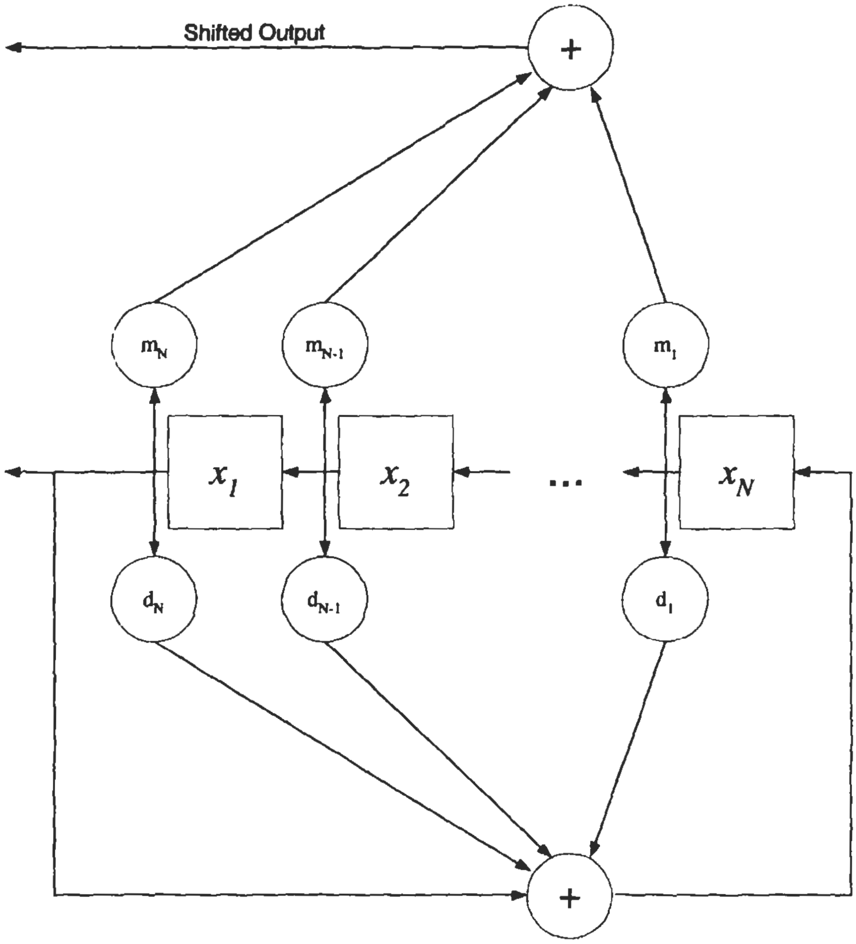

Figure 2-3: PN-Mask Circuit

Assume that the N by N Walsh matrix is denoted as $\mathbf{H}_N$. Then the first two Walsh matrices in the series $\{\mathbf{H}_1, \mathbf{H}_2, \mathbf{H}_4, \ldots\}$ are

$$\mathbf{H}_1 = [1]$$
$$\mathbf{H}_2 = \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \tag{2-10}$$

In fact, each consecutive Walsh matrix in the series can be generated recursively from the previous Walsh matrix in the series, i.e.,

$$\mathbf{H}_{2N} = \begin{bmatrix} \mathbf{H}_N & \mathbf{H}_N \\ \mathbf{H}_N & -\mathbf{H}_N \end{bmatrix} \tag{2-11}$$

5. SPREADING

Now that the concepts of PN sequences and orthogonal functions have been introduced, these areas may be used to develop the idea of spreading a signal. Recall that in any direct-sequence spread spectrum system, the objective is to transform narrowband information into a wideband signal for transmission. This is achieved by modulating the narrowband signal with a waveform with a higher bandwidth. For a spreading sequence in a CDMA system, its bandwidth is a function of the duration of one of its spreading sequence samples. This duration is known as a *chip*.

Recall that PN sequence samples are generated according to a clock signal supplied to the associated shift register. This clock frequency is essentially the *chip rate* (also "chipping rate") of the spreading sequence. If the chipping rate is designated as C and the information bit rate is R, then C must be greater than R for bandwidth expansion to occur.

The basic spreading operation is depicted in Figure 2-4.

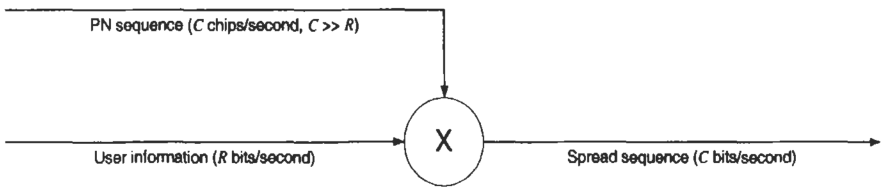

Figure 2-4: Spreading Operation

The multiplication operation in Figure 2-4 can also be accomplished through the use of an exclusive-NOR gate if the input signals are binary.

5.1 Quadrature Spreading

A common method for spreading in CDMA systems involves the use of two spreading sequences to modulate user information. User information in this case is usually split into two information streams, one corresponding to an "in-phase" or "I" component and one corresponding to a "quadrature" or "Q" component. If the two spreading sequences are denoted as PN_I and PN_Q, then the basic *quadrature spreading* operation is as depicted in Figure 2-5.

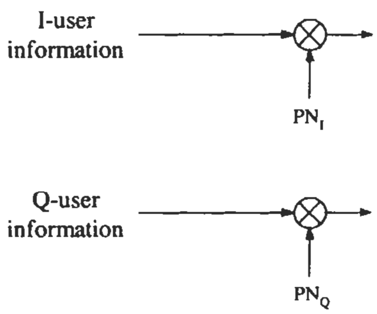

Figure 2-5: Quadrature Spreading

The use of multiple spreading sequences can actually result in *diversity*, i.e., the transmission of multiple signal copies over different transmission paths. Recall that the partial correlation properties of a spreading sequence are important for proper reception of the CDMA signal. The partial correlation at any given instant in time on the in-phase and quadrature received signals is very likely different.

Another form of quadrature spreading is *hybrid* quadrature spreading, which involves representation of the two PN sequences and the I and Q input signals as complex quantities and accomplishes spreading by modulating the input signal with the complex PN sequence. This operation is shown in Figure 2-6.

Hybrid quadrature spreading still provides diversity as with quadrature spreading, but it also reduces the output signal variation, which is important in the design of CDMA radios.

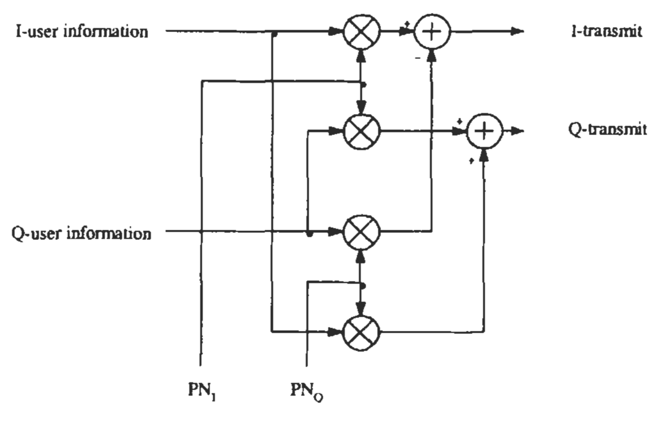

Figure 2-6: Hybrid Quadrature Spreading

5.2 Orthogonal Modulation

Orthogonal modulation is used in CDMA systems as a means of distinguishing either: (a) one user's information from another on the downlink, or (b) one type of information from another type from the same user on the uplink.

Orthogonal modulation operates by assigning a user a Walsh code, which is drawn from the Walsh matrix specified by (2-10) and (2-11). The Walsh code assigned to a user on the downlink is unique for that user. Moreover, dedicated Walsh codes can be used for common information streams (or "channels") that are to be monitored by all mobiles. On the uplink, however, different Walsh codes transmitted from a single user can indicate different types of information. For instance, a user may transmit data over one Walsh code and voice over another.

Each information bit to be transmitted is modulated by the user's Walsh code. The information bit is assumed to occur at a lower rate than the chipping rate, so each information bit is modulated by the entire Walsh code assigned to that user. This is also sometimes referred to as "Walsh spreading." Walsh spreading operates identically to the spreading operation depicted in Figure 2-4, except that a Walsh code is used rather than a PN sequence. The Walsh code repeats at the input information bit rate, and sometimes each Walsh code entry is referred to as a Walsh chip. An example of a waveform spread by a Walsh code is given in Figure 2-7.

The Walsh code assigned to a single user does not have to be the same length as the Walsh code assigned to another user. For instance, the Walsh code "-1 +1" is orthogonal to the Walsh code "+1 -1 +1 -1". However, Walsh code "-1 +1" would correspond to a higher information rate than "+1 -1 +1 -1".

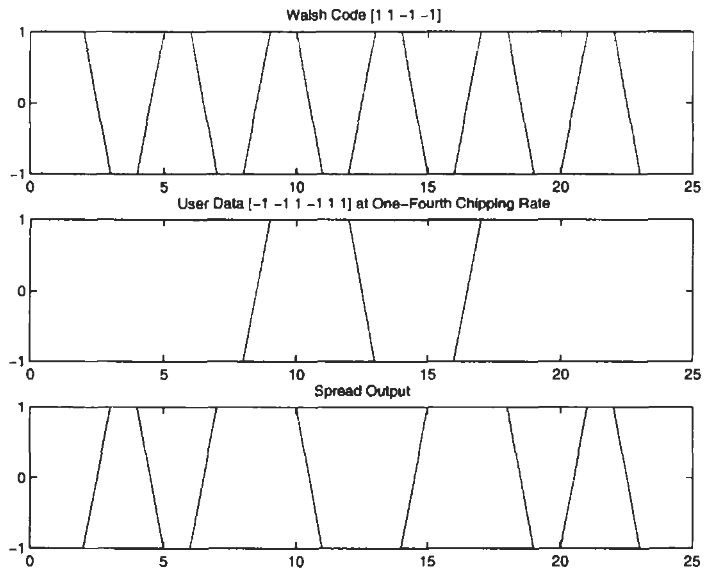

Figure 2-7: Walsh Spreading

In practice, several Walsh codes may be assigned on the downlink, each for a particular mobile (see Figure 2-8). Similarly, on the uplink, an individual mobile may transmit different types of information over different Walsh codes. For instance, a mobile may use one Walsh code to transmit voice information and another to transmit video.

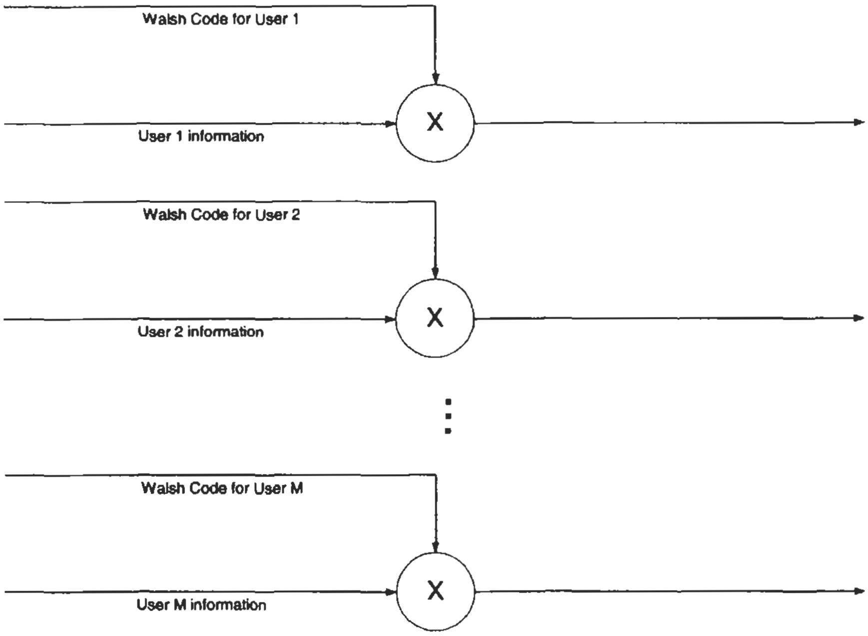

Figure 2-8: Walsh Codes for Multiple Users

5.3 Sequential Spreading Process

The spreading process in a typical CDMA system involves first the orthogonal modulation described in Section 5.2 followed by quadrature spreading described in Section 5.1. If the user information has been separated into in-phase and quadrature components, then the orthogonal modulation is usually performed prior to the quadrature spreading in order to reduce the number of computations necessary to carry out the spreading operation. For instance, examine the two-user quadrature spreading example in Figure 2-9.

By preceding the quadrature spreading operation with the orthogonal modulation, only one quadrature spreading operation is necessary for all the users. If the orthogonal modulation is performed *after* the quadrature spreading operation, two quadrature spreading operations are required — one per user. Otherwise there would be no way of separating all the users' information for orthogonal modulation purposes.

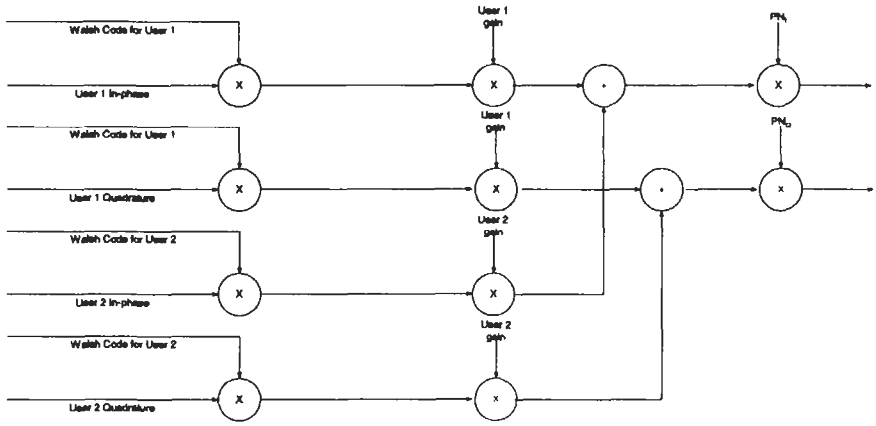

Figure 2-9: 2-user Orthogonal Modulation and Quadrature Spreading

It should be noted that because the PN sequence period is usually much longer than the duration of an entire Walsh code in most CDMA systems, the processing gain is directly dependent on the length of the Walsh code used. Clearly, the trade-off becomes lower processing gain when information rates increase for a given user.

6. MODULATION CONSTELLATIONS

In Section 5, the concept of separating user information into in-phase and quadrature components was introduced. The actual method of extracting in-phase or quadrature components from a continuous stream of user information bits usually involves the use of a *signal constellation*. A signal constellation maps a group of bits to a complex number. Each complex number has a real and imaginary part; the real part corresponds to the in-phase information signal and the imaginary part to the quadrature information signal. The complex number is also known as a *modulation symbol*. The total number of symbols in a constellation is referred to as the *constellation order*. This basic operation is depicted in Figure 2-10.

The in-phase and quadrature signal sequences resulting from the constellation mapper in Figure 2-10 can be provided as input to a spreading circuit such as the ones in Figure 2-5 and Figure 2-6.

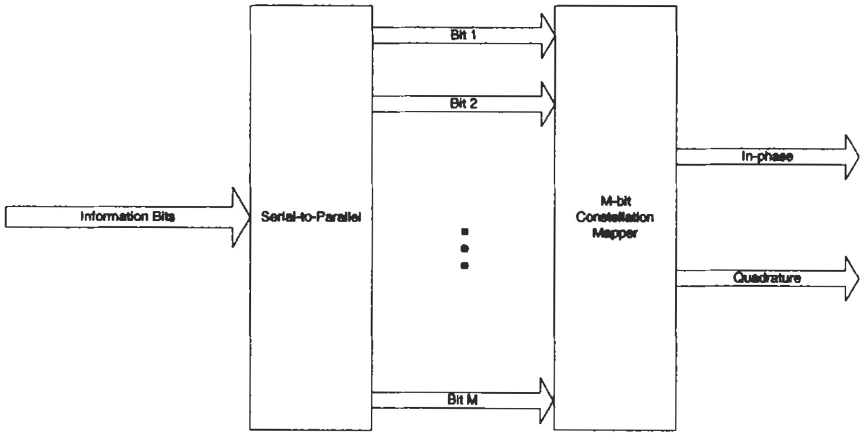

Figure 2-10: Constellation Mapping Operation

Modulation constellations encompass two primary types of modulation: *phase-shift keying* (PSK) and *quadrature amplitude modulation* (QAM). These two types of modulations will now be examined.

PSK modulation schemes map information bits to phases on the unit circle in the complex plane. The two most commonly used forms of PSK in CDMA systems are *binary phase shift keying* (BPSK) and *quadrature phase shift keying* (QPSK). BPSK involves simply mapping a single bit to one of two values on the real axis of the complex plane; it is sometimes referred to as *antipodal* signaling. A sample BPSK constellation is pictured in Figure 2-11. This example shows a simple mapping of a binary "1" to the in-phase value of +1 (0 degree phase) and a binary "0" to the in-phase value of –1 (180 degree phase).

Both mappings have a quadrature value of 0; as a result, BPSK is more interference-resilient than other modulation constellations that use non-zero quadrature values.

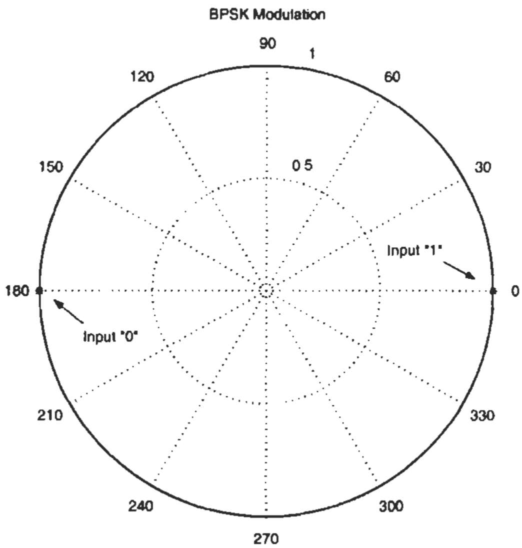

Figure 2-11: BPSK Constellation

Quadrature phase shift keying involves the mapping of two input bits to one of four possible constellation values. An example QPSK constellation is shown in Figure 2-12. The input mapping depicted in Figure 2-12 shows an example of *Gray encoding*. Gray encoding involves mapping groups of input bits to modulation constellation symbols in such a way that adjacent modulation symbols differ only in one bit. This is beneficial when the symbols are detected at a receiver, as common receiver errors result in a modulation symbol being detected as one of the adjacent symbols [6].

Quadrature amplitude modulation involves mapping to points in the complex plane wherein the magnitude (the square root of the sum of the squares of the complex and real parts) of the modulation symbol can be different for different symbols. This differs from PSK constellations that actually restrict the magnitudes of modulation symbols to map to the unit circle. The 16-QAM constellation with Gray encoding is pictured in Figure 2-13.

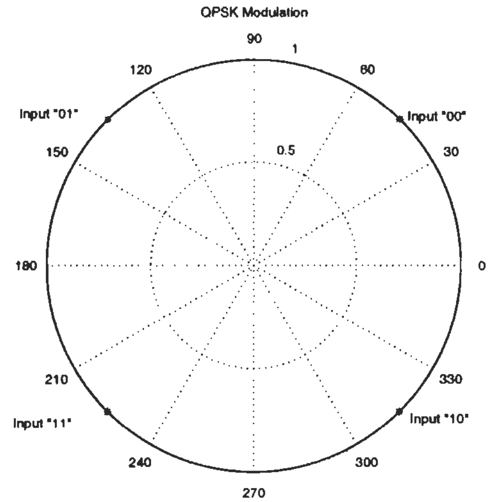

Figure 2-12: QPSK Modulation

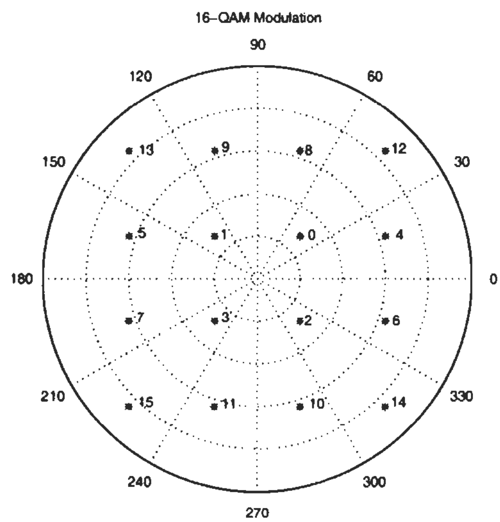

Figure 2-13: 16-QAM Modulation

QAM constellations require an amplitude reference at the receiver (unlike PSK), which sometimes makes them undesirable for communications systems that undergo amplitude distortion during transmission (such as in wireless). However, this drawback is balanced by the increased throughput QAM systems offer versus PSK systems when the signal-to-interference ratio is sufficiently large. This is due to the fact that the error performance for PSK or QAM constellations is bounded by the *minimum Euclidean distance* between constellation points [6]. The Euclidean distance between two constellation points represented by the complex scalar quantities $A + Bj$ and $C + Dj$ (where j is the square root of —1 and A, B, C and D are real constants) is

$$d = \sqrt{(A-C)^2 + (B-D)^2} \tag{2-12}$$

Given that the noise power in a communications system is represented by the quantity N_0, for an M-point constellation, the symbol error is bounded as a function of the minimum distance between any two constellation points d_{min}:

$$P_M < (M-1)Q\left(\sqrt{\frac{d_{\min}^{\;2}}{2N_0}}\right) \tag{2-13}$$

In (2-13), the function $Q(\,)$ is defined as

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-\frac{t^2}{2}} \mathrm{d}t \tag{2-14}$$

The *Q-function* describes a portion of the area under a curve with a Gaussian distribution, and is useful in characterizing the performance of communications systems with background noise.

Equation (2-13) implies that the larger d_{min} is for a given constellation with a fixed number of points, the better the performance. As the number of symbols in the constellation increases, the restriction in PSK systems of equal magnitude symbols results in a smaller d_{min} than a QAM system with an identical number of symbols. As a result, PSK systems are usually sufficient for constellation orders up to 8, but larger-order constellations require QAM constellations.

It should also be noted that (2-13) also implies that larger order constellations suffer in performance when compared to smaller order constellations, due to the fact that d_{min} decreases as more and more symbols are packed into the complex plane.

7. PULSE SHAPING

The spreading operation results in a bandwidth expansion of the signal to be transmitted. While this is certainly a desired effect for a spread spectrum system, other considerations exist where strict bandwidth requirements need to be met. For instance, a CDMA system deployed in a spectrum allocated for public wireless systems such as the PCS band in North America will have to operate around several other types of systems (e.g., TDMA, other CDMA systems). If strict bandwidth limitations are not placed on public wireless systems, *spectral emissions* can become problematic. Spectral emissions are a measurement of the amount of power leaked outside the allowed bandwidth for a wireless system. If acceptable emissions limits are exceeded, the system may be in violation of national or international regulations. For instance, the International Telecommunications Union (ITU) has emissions limits in place for various spectral bands around the world. In addition, in the United States, the Federal Communications Commission (FCC) has placed emissions limitations for *Part 15* equipment, which encompasses commercial wireless devices.

Pulse shaping is the process of filtering a signal for transmission for the purposes of bandlimiting the signal. A filter is normally described by its *impulse response function*, $h(t)$. Given an input signal $x(t)$, the output of the filtering operation may be represented as

$$y(t) = \int_{-\infty}^{t} x(\tau)h(t-\tau)\mathrm{d}\tau \tag{2-15}$$

The operation depicted in (2-15) is also known as a linear *convolution* operation. Filters whose output is determined by a linear convolution operation are also known as *linear filters*. The convolution of two signals in the time domain results in the multiplication of those signals' spectrums. Therefore, if the filter spectrum is bandlimited, the output of the filter will also be bandlimited.

Normally, for communications systems it is desirable that the pulse-shaping filter satisfies the *Nyquist criterion*. If the input symbol duration is T, then the Nyquist criterion is

(2-16)

$$h(nT) = \begin{cases} 1, n = 0 \\ 0, n \neq 0 \end{cases}$$

This criterion ensures that no *intersymbol interference* results from pulse-shaping. Intersymbol interference (ISI) is a result of distortion of a communications signal in such a way that consecutive transmitted symbols overlap in time.

It should be noted that meeting the Nyquist criterion and meeting emissions requirements simultaneously is not always possible. Sometimes in CDMA systems, the pulse-shaping may not meet the Nyquist criterion if the amount of ISI introduced by the transmit filter is at acceptable levels. However, the trade-off in meeting emissions requirements and the Nyquist criterion becomes even more significant if the processing gain is reduced, as the signal under such conditions will be even more sensitive to ISI.

With the pulse-shaping filter, a basic CDMA transmit chain can now be derived as in Figure 2-14.

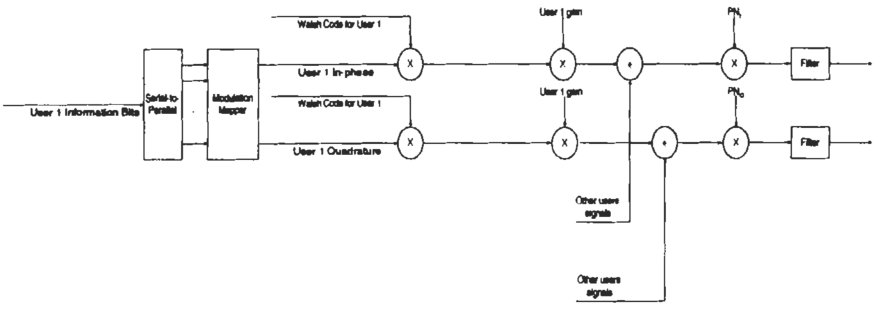

Figure 2-14: CDMA Transmit Chain

8. CHANNEL CODING

Many times, the distortion present in a wireless communications system requires error protection. Channel coding is a widely used means of providing error protection by adding redundancy to the user information that the receiver can use to correct errors that may have occurred during transmission.

Usually two types of channel coding are present in CDMA systems: *convolutional* coding and *turbo* coding. Convolutional coding is exclusively used in second-generation CDMA systems. However, turbo coding has been adopted for third-generation CDMA systems to provide error resiliency, primarily for data applications.

8.1 Convolutional Coding

Convolutional coding is usually implemented by passing input information bits into a shift register of some predetermined length. The outputs of each delay element in the shift register are then combined through modulo-2 addition to yield several redundancy bits. These bits are processed by the receiver to retrieve the original information sequence. For instance, the IS-95 convolutional code is depicted in Figure 2-15.

A convolutional code *rate* is the ratio of the number of input bits k to the number of output bits n, i.e., the rate is k/n. For the code in Figure 2-15, the code rate is 1/2. The *constraint length* K of a convolutional code is the number of delay elements in the shift register used for the code plus one. The IS-95 code has a constraint length $K = 9$.

Convolutional code performance increases with decreasing code rate and with increasing constraint length. However, this performance increase comes at a considerable cost. In decreasing the code rate, more redundancy is necessary for each input bit to the convolutional encoder. If the system is bandwidth-limited, this restricts the maximum bit rate of the system.

Increasing the constraint length does not significantly impact the complexity of the encoding process. However, the decoder used at the receiver is impacted heavily by an increase in the constraint length. This is due to the fact that typical communications systems receivers use *Viterbi decoders* to decode a convolutional code. A Viterbi decoder takes advantage of a *trellis* representation of the convolutional code and finds the optimal path through the trellis based on the received symbols.

The number of convolutional encoder shift register states is equal to 2^{K-1}, resulting in an exponential increase with constraint length. The trellis representation takes advantage of the fact that, based on all possible inputs to the convolutional encoder, all of the number state transitions are characterized recursively. The trellis relationship for the convolutional code of Figure 2-15 is shown in Figure 2-16. In Figure 2-16, the states of the shift register are represented as functions of the value J, which is an integer ranging from 0 to 127. These values are referred to as *nodes* of the trellis. M represents a *branch metric*, which is the Euclidean distance of each received information symbol from each expected symbol associated with that particular state transition.

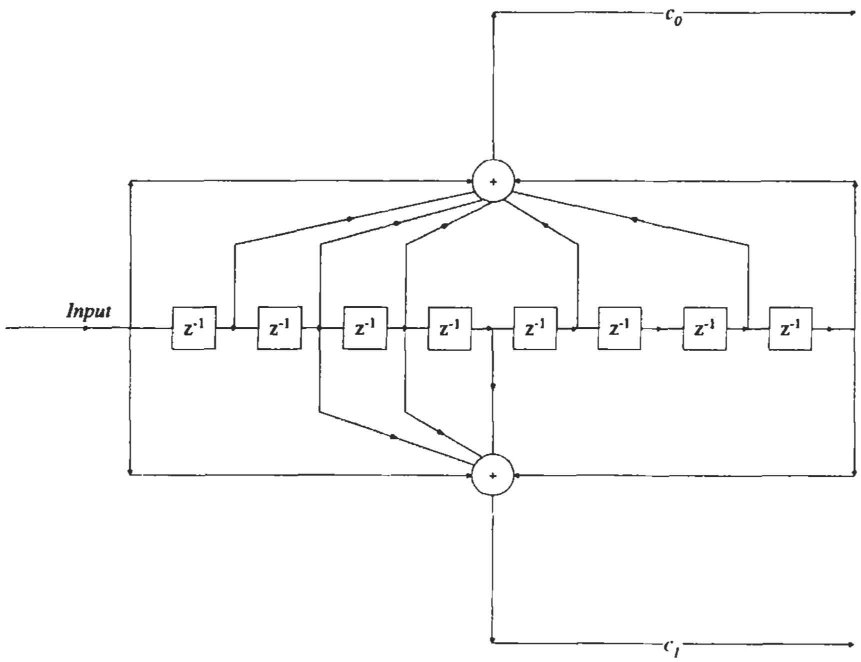

Figure 2-15: IS-95 Convolutional Code

The trellis represents the state transitions from all the nodes of a given *stage* of the trellis to the next stage of the trellis. A stage is associated with each bit input to the convolutional encoder. The Viterbi algorithm compares the cumulative metrics of all the paths entering each node in the trellis and selects a *survivor* path. All other paths are eliminated and the algorithm is repeated for the next stage of the trellis.

The trellis representation increases exponentially with constraint length, thereby making the decoding process more complicated. A more detailed description of the trellis representation and Viterbi decoding algorithm may be found in [7].

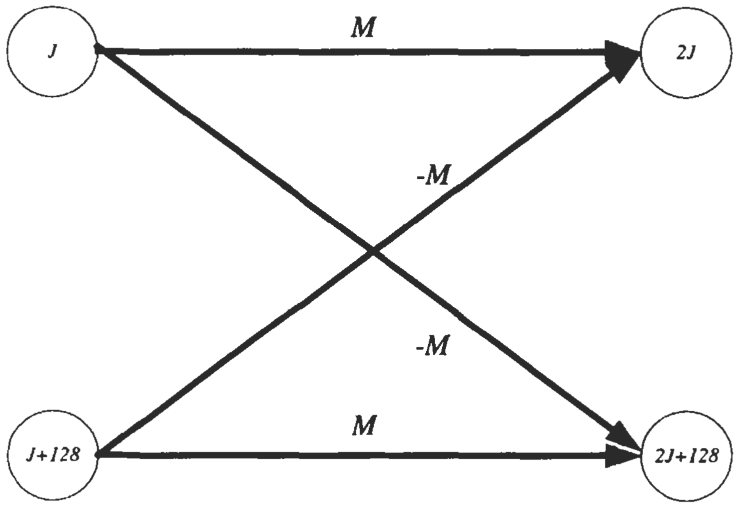

Figure 2-16: Trellis for IS-95 Convolutional Code

8.2 Turbo Encoding

Turbo encoding was incorporated into third-generation CDMA systems as a means of improving the performance for data traffic in wireless systems. Turbo encoding involves the use of several channel coders (typically convolutional encoders) interworked with one other to produce output symbols. Each channel code used for a particular turbo encoder is known as a *constituent encoder*. A general turbo code structure is given in Figure 2-17.

In Figure 2-17, block interleavers are used at the input of all except one of the constituent encoders. A block interleaver formats the data into a two-dimensional array by reading bits into the interleaver memory sequentially and reading out from the memory in a different fashion. One simple means is to read into the interleaver from left-to-right and top-to-bottom, but read out top-to-bottom and left-to-right. Interleaving is used in communications systems to provide robustness against bursty errors, but introduces a delay in the system in that the entire block of data must be read into the interleaver memory before it is read out. This is one of the reasons turbo codes are primarily used for data applications that are delay-tolerant, but not for applications such as voice telephony, which is sensitive to delay.

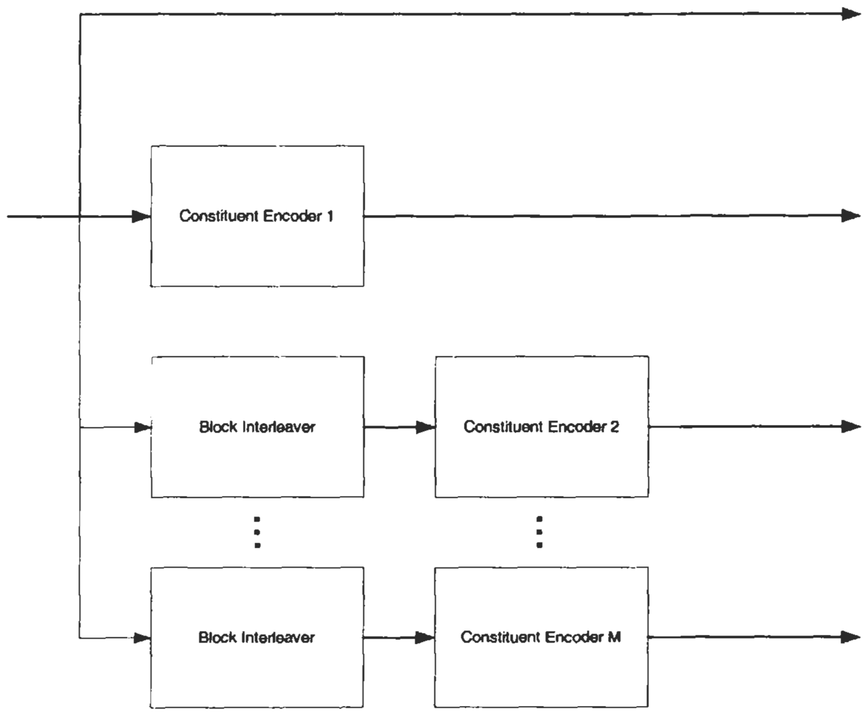

Figure 2-17: Turbo Code

Also note that the input bits are sent unencoded in the turbo code. These bits are sometimes called *systematic* bits and are important for the turbo decoding process.

The turbo code pictured in Figure 2-17 is sometimes also referred to as a *parallel-concatenated turbo code*, as each of the constituent encoders works in parallel with each other. A *serially concatenated* turbo code would involve constituent encoder outputs driving other constituent encoder inputs. Generally, parallel-concatenated turbo codes are used in third-generation CDMA systems.

Turbo decoding is a complicated process for which there are several solutions. Among them, MAP (maximum a posteriori) decoding and soft-output Viterbi algorithm (SOVA) decoding are prominently used. MAP decoders generally provide better performance than SOVA decoders, but SOVA decoders are relatively simple to implement. For more information on turbo decoding methods, see [8]-[10].

9. CONCLUSIONS

A basic overview of direct-sequence spread spectrum systems was provided in this chapter. This serves as a basis for the next chapter, which discusses the mobile wireless channel and diversity receivers for CDMA systems.

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Chapter 3

The Mobile Channel and Diversity Reception in CDMA Systems

Key Topics: rake receiver, diversity, fading

Overview: The mobile radio channel in wireless systems is described and concepts such as multipath, coherence bandwidth, fading, and diversity are introduced. The wireless channel is exploited in CDMA systems to improve performance under diversity conditions by using the *rake receiver*. In addition, an overview on the power control mechanism in CDMA systems is given in relation to interference issues specific to CDMA systems.

1. INTRODUCTION

Mobile wireless transmissions undergo more distortion in the mobile channel than do wireline systems. These effects result in distortion of the transmitted signal and are a result of several factors, including reflectors in the transmission path, mobile velocity, the transmission frequency (i.e., *carrier frequency*), and other parties' attempts to utilize the transmission band.

Generally, the signal to be transmitted is pulse-shaped to restrict its bandwidth. Receivers are also designed using bandpass filters to isolate the signal of interest. As a result, the effects of the mobile channel are usually considered only in the band of interest, as effects outside the transmission band are assumed to be suppressed through receiver filtering. In addition, if a perfect frequency reference is available at the receiver such that the signal of interest may be demodulated from its carrier frequency properly, then the channel may be analyzed using a *lowpass* model.

Reflections in the transmission path result in multiple signal copies arriving at the receiver. These multiple signal copies are a result of *multipath*, i.e., multiple propagation paths between the transmitter and the receiver. A simple example of two-path multipath is given in Figure 3-1.

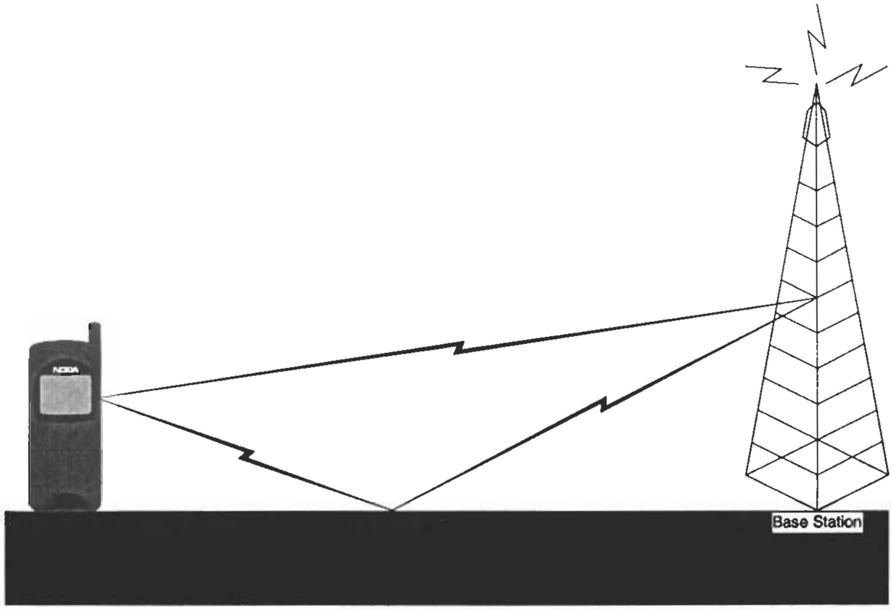

Figure 3- 1: Mobile Multipath Channel

These multiple signal copies usually do not arrive at the receiver simultaneously. As a result, intersymbol interference generally is a consequence of the addition of these multipaths at the receiver.

The *delay spread* [1] of the mobile wireless channel is defined as the difference in arrival times at the receiver between the earliest arriving and the latest arriving multipath. This quantity will be denoted as T_m. Thus the channel profile at any given instant in time may be modeled based on the arriving multipaths and their relative complex amplitudes. A sample channel profile is given in Figure 3-2.

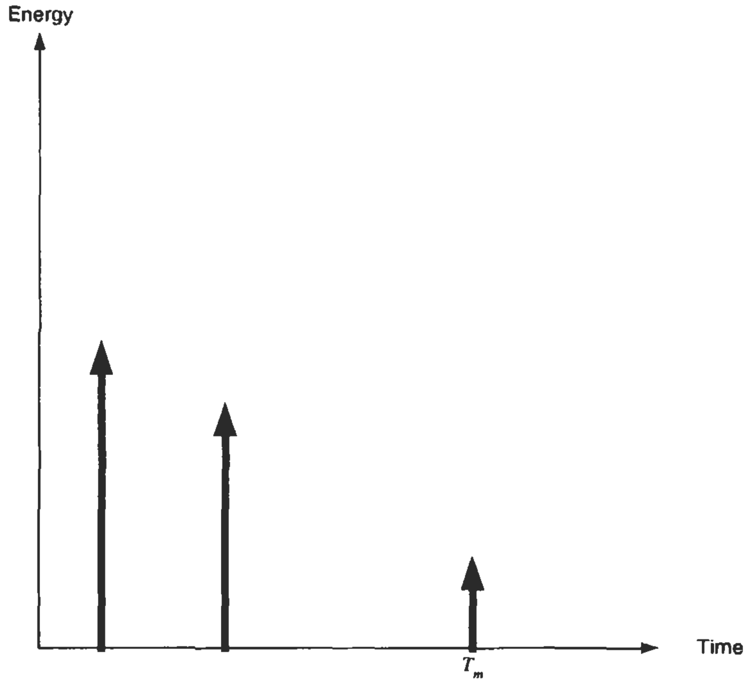

Figure 3-2: Channel Profile

The channel profile shown in Figure 3-2 is also the *channel impulse response* as it provides an energy distribution for each of the multipaths that can be used to model the channel as a linear filter.

The *coherence bandwidth* of the channel is the minimum frequency separation required between two sinusoids in the transmission band such that the channel-induced distortion for each of the sinusoids is uncorrelated. From the delay spread, the coherence bandwidth of the channel B_c may be approximated as:

$$B_c \approx \frac{1}{T_m} \tag{3-1}$$

Any motion of a mobile results in time variations in the channel. This results in a frequency shift of the signal spectrum, also known as *Doppler shift*. This quantity is defined as

$$f_d = \frac{v}{\lambda} = \frac{v}{c / f_c} \tag{3-2}$$

In (3-2), v is the mobile velocity in meters/second, λ is the carrier wavelength in meters, c is the speed of light (3×10^8 m/s), and f_c is the carrier frequency in Hertz. As the mobile moves toward the base station, the Doppler shift is positive in frequency; as it moves away, the shift is negative.

The impact of Doppler is usually modeled as a multiplicative effect. As a result, the Doppler spectrum has a bandwidth that covers the total range of the Doppler shift:

$$B_d = 2 f_d \tag{3-3}$$

The Doppler spectrum usually takes an "inverse horseshoe" shape as in Figure 3-3. The amplitude variation in the received signal resulting from the mobile velocity and carrier frequency is referred to as *fading*.

The *coherence time* of the channel (Δt_c) is the duration over which the channel is correlated in time. It is roughly equivalent to the inverse of the Doppler bandwidth B_d.

Given the concepts of fading and multipath, the mobile channel may be generally represented as a time-varying impulse response. If there are L multipaths present, and each multipath i has a propagation delay of $\tau_i(t)$ and complex amplitude of $\alpha_i(t)$ at time t, then the channel impulse response is

$$c(t) = \sum_{i=1}^{L} \alpha_i(t) e^{-j2\pi f_c \tau_i(t)} \delta\left(t - \tau_i(t)\right) \tag{3-4}$$

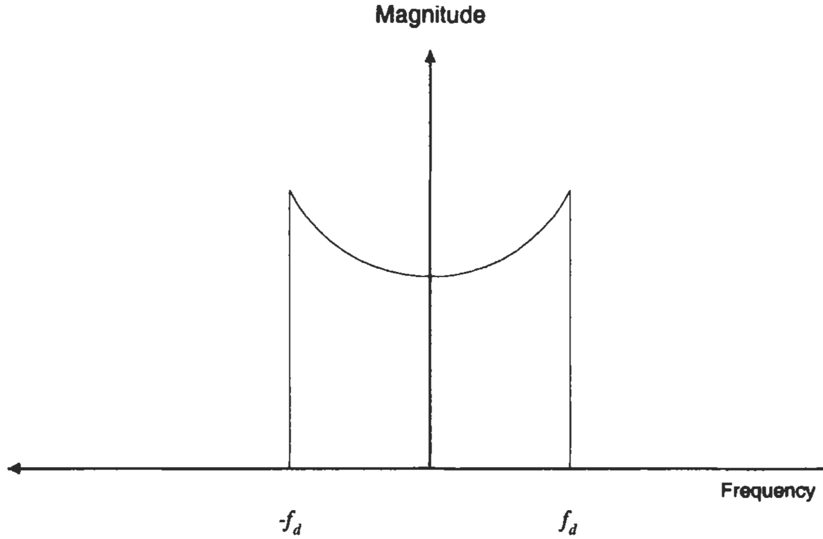

Figure 3-3: Doppler Spectrum

Thus the received signal $r(t)$ is simply a convolution of the channel impulse response and the transmitted signal $s(t)$:

(3-5)

$$r(t) = \int_{-\infty}^{\infty} c(\tau)s(t-\tau)\mathrm{d}\tau$$

Given the frequency responses of the channel and transmitted signal ($C(f)$ and $S(f)$ respectively), this can also be represented as

(3-6)

$$r(t) = \int_{-\infty}^{\infty} C(f)S(f)e^{j2\pi ft}df$$

It should also be noted that the received signal is also modeled with a background noise term, $z(t)$, which is usually assumed to be Gaussian. This term is additive, and is therefore often referred to as *additive white Gaussian noise* (AWGN), i.e.:

$$r(t) = \int_{-\infty}^{\infty} c(\tau)s(t-\tau)\mathrm{d}\tau + z(t) \tag{3-7}$$

2. FADING CHANNEL ANALYSIS

Let us assume that the signaling interval (i.e., the transmit bit duration) is T and a signal bandwidth of W. The equivalent spectral distortion due to channel effects is a function of the following:

1. The bandwidth of the channel
2. The bandwidth of the transmitted signal
3. The spectral response of channel within signal bandwidth

If $W > B_c$ (B_c being the coherence bandwidth of the channel), then the channel is said to be *frequency-selective*. The frequency selectivity of the channel is a function of the multipath spread.

If $W << 1/T_m$ (T_m being the delay spread of the channel) or if $W << 1/B_c$, then channel is *frequency-nonselective*. In this case, the signal bandwidth can be assumed to be concentrated around the zero frequency ("DC").

If multiplicative channel amplitude is a zero-mean complex valued Gaussian random process, then for any t, the received signal envelope $\alpha(t)$ is Rayleigh-distributed (square-root of the sum of the squares of two Gaussian random variables) and $\phi(t)$ is uniformly distributed from $(-\pi,\pi)$. In this case, the Doppler bandwidth is sufficient to give an idea about the fade duration.

If it is possible to select W such that $W << B_c$, and T such that $T << (\Delta t_c)$ ((Δt_c) being the coherence time of the channel), then the signal is subject to a *slow-fading channel*. An example of a slow-fading channel could occur in a mobile radio communications system where only a single transmission path exists (guaranteeing a large coherence bandwidth) and a relatively low velocity between the transmitter and receiver (leading to a large coherence time).

The product $T_m B_d$ is the *spread* factor. If $T_m B_d << 1$, the channel is underspread, otherwise it is overspread. Most low-frequency communications systems (e.g., VHF) will be underspread, but high-frequency communications systems (e.g., satellite) have higher spreading factors. There are some ways to combat high spread factors:

1. Interleaver design (making interleaver depth greater than coherence time of channel "decorrelates" the fade; see also Chapter 2, Section 8.2)
2. Induced diversity. Adding multiple transmission paths that are separable at the receiver (e.g., multiple transmission antennas).
3. Automatic gain control. This is a control loop at the receiver that adjusts the incoming received signal power as presented to various components in the receiver chain. In particular, if analog-to-digital (A/D) converters are used, it is important that the power of the signal be adjusted with respect to the maximum input voltage levels (i.e., the "rail voltages") of the A/D converters. The AGC loop must be capable of responding to changes in received signal power so that the A/D converter does not get overloaded (or underloaded) for long periods of time relative to the signaling interval. However, the AGC loop must not be so sensitive to normalize the received signal energy very quickly with respect to fade duration, as information about channel characteristics relative to the current received symbol may be lost with respect to its context. This affects the effectiveness of any forward error correcting codes.

2.1 Frequency-Nonselective, Slowly Fading Channel

Let us assume that the random process characterizing the channel distortion of the slow-fading channel is considered to be constant over a signaling interval:

$$r(t) = \alpha e^{-j\phi} s(t) + z(t) \qquad 0 \le t \le T \tag{3-8}$$

If we assume that the channel-induced phase shift is perfectly estimated, then *coherent* detection is possible. Coherent detection is possible using matched filtering for PSK or QAM systems. Since we are assuming fixed attenuation for the signaling interval, the probability of error for BPSK (P_2, where the '2' signifies bi-level signaling) in AWGN is

$$P_2(\gamma_b) = Q\left(\sqrt{2\gamma_b}\right) \tag{3-9}$$

where $\gamma_b = \alpha^2 E_b/N_0$ is the SNR-per-bit.

2.2 Rayleigh Fading

If α^2 is the sum of the squares of two Gaussian random variables, then it is *Rayleigh*-distributed. As a result, γ_b, which is simply a scaling of α^2, is also chi-square. Thus the distribution results:

(3-10)

$$p(\gamma_b) = \frac{1}{\bar{\gamma}_b} e^{-\gamma_b / \bar{\gamma}_b}$$

where

(3-11)

$$\bar{\gamma}_b = \frac{E_b}{N_0} E(\alpha^2)$$

For BPSK, the probability of error now becomes

(3-12)

$$P_2 = \frac{1}{\sqrt{2\pi\bar{\gamma}_b}} \int_0^\infty e^{-\frac{\gamma_b}{\bar{\gamma}_b}} \int_{\sqrt{2\gamma_b}}^\infty e^{-\frac{t^2}{2}} dt\, d\gamma_b$$

which simplifies to

(3-13)

$$P_2 = \frac{1}{2}\left(1 - \sqrt{\frac{\bar{\gamma}_b}{1+\bar{\gamma}_b}}\right)$$

2.3 Nakagami Fading

The Nakagami-m distribution is the following distribution for γ_b:

(3-14)

$$p(\gamma_b) = \frac{m^m}{\Gamma(m)\bar{\gamma}_b^{\,m}} \gamma_b^{\,m-1} e^{-m\gamma_b / \bar{\gamma}_b}$$

$m = 1$ is simply Rayleigh fading.

Nakagami fading has been suggested as an alternative to Rayleigh fading for communications systems that do not undergo the severe fading patterns modeled by Rayleigh fading.

3. DIVERSITY TECHNIQUES FOR FADING MULTIPATH CHANNELS

Diversity techniques involve transmission of several copies of the information-bearing signal over multiple transmission paths to ensure recovery of the signal when one (or more) of these transition paths is in a deep fade (provided at least one of the remaining transmission paths is not). Given more and more transmission paths, it is highly unlikely that all will be in a deep fade simultaneously.

Frequency diversity is one method, wherein the same information is transmitted on multiple carriers separated in frequency by some interval greater than the coherence bandwidth. This method has the expense of reducing spectral efficiency in the Gaussian channel.

Temporal diversity involves multiple copies of the signal separated in time by durations greater than the coherence time of the channel. Block interleaving is a means of achieving temporal diversity. Once again, this method has the expense of reducing spectral efficiency in the Gaussian channel.

Both methods mentioned above have the additional drawback of sometimes not being sufficiently robust to changing coherence times or bandwidths — this means that such systems are often designed with the worst case of these conditions in mind, or sometimes other constraints may come into play. For instance, block interleaving induces delay into the system — a real-time voice service may not be tolerant of excessive delay before noticeable degradation to the end listener results.

Finally, spatial diversity, employing a combination of multiple transmit and/or multiple receive antennas, can be used to avoid sensitivity to the coherence time and bandwidth of the channel. However, sufficient separation between antennas is necessary to ensure the transmission paths of the system are uncorrelated. Once again, this may not be possible. For instance, pole-mounted base stations have strict space limitations, ensuring that a maximum number of similarly polarized antennas may be deployed.

If the signal bandwidth W is much greater than the coherence time of the channel, then the multipath components are resolvable. This provides the receiver with several independently fading transmission paths. These paths must be separated in time by at least $1/W$. There are thus $T_m W$ resolvable multipaths, i.e., W/B_c resolvable multipaths. The optimum receiver for such a signal is known as a *rake* receiver.

The rake receiver works by having multiple demodulator elements (i.e., "fingers") tuned to each transmission path. It is the most commonly used receiver for a CDMA system. The rake fingers are combined based on their respective signal strengths; this process is known as *maximal ratio combining* (MRC) (see Figure 3-4).

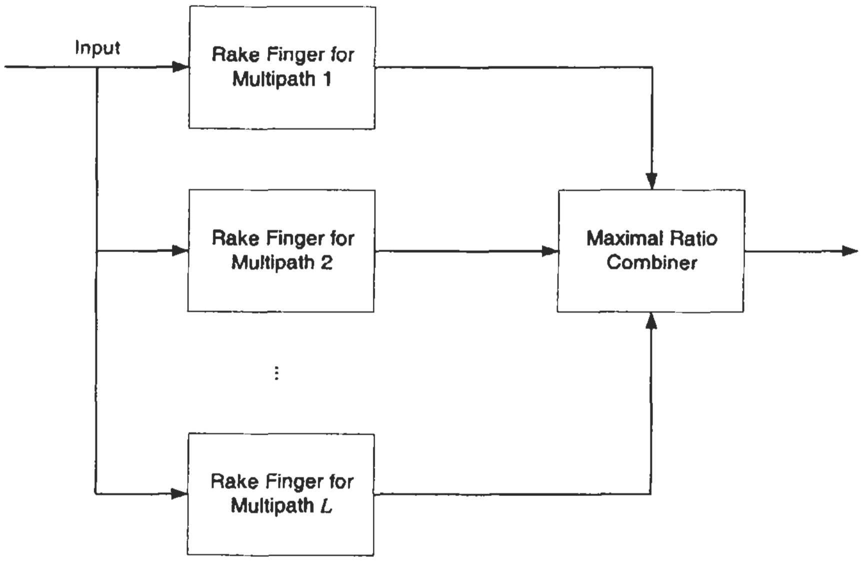

Figure 3-4: Rake Receiver with MRC

3.1 Binary Signals

If we assume L statistically independent transmission paths to the receiver in a binary signaling system, then the possible received signals for each transmission path may be represented as

$$r_k(t) = \alpha_k e^{j\phi_k} s_{km}(t) + z_k(t) \quad k = 1,2,\ldots,L \quad m = 1,2 \tag{3-15}$$

In (3-15), m denotes each of the possible binary signals that may be transmitted at any given time. Each demodulation element is composed of a matched filter for each possible binary signal. Following the matched filters is the maximal ratio combiner which forms the decision variables for the two possible transmitted signals; each filter output is weighted by the conjugate channel gain and phase distortion $\alpha_k e^{j\phi_k}$. This not only corrects for phase distortion but also weights the demodulator output by the received signal strength over each particular transmission path. In order to estimate the channel amplitude, a *pilot* signal is usually required. In CDMA systems, the pilot signal is usually in the form of a constant information sequence (e.g., all binary 1's) transmitted over a known Walsh code. Since the information sequence is known at the receiver, the pilot channel can be acquired at the receiver.

The channel estimation and phase distortion may be estimated at the receiver using a filter. One common approach for the estimation filter is a simple rectangular window filter, which just entails a sum over a fixed sliding window of received samples. An example of the entire despreading and channel estimation process for a single rake finger for a BPSK system with hybrid quadrature spreading is given in Figure 3-5.

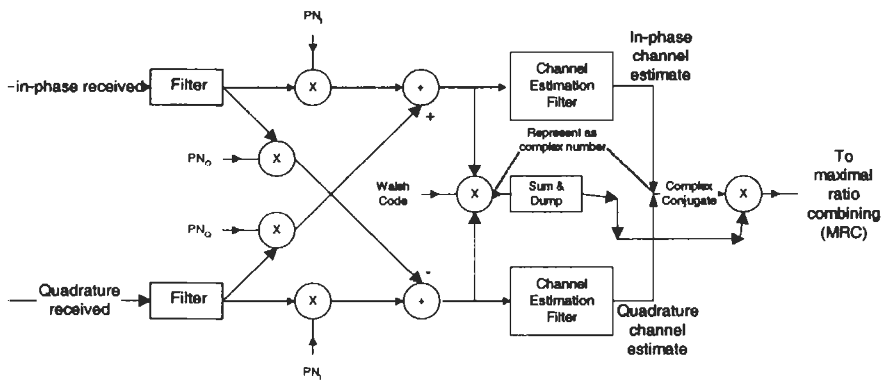

Figure 3-5: Rake Finger Processing

A more detailed description of rake receiver processing may be found in [2].

4. POWER CONTROL

Power control refers to a method of combating the *near-far effect* often found in mobile communications systems. In a typical mobile communications system, the base station must serve a variety of users (i.e., other handsets) simultaneously. Under such conditions, a handset very close to the base station transmitting at an excessively high power may have a tendency to "swamp out" users further away from the base station due to the differences in propagation losses. This effect is particularly troublesome in spread spectrum wireless systems, where many users occupy the same bandwidth continuously. In order to mitigate these effects, the base station uses power control whereby it commands handsets to increase or decrease their individual radiated power levels depending on the signal-to-noise ratio of the individual users' signals as detected at the base station.

The power control system is modeled as a linear time-invariant system. As a result, a typical power control system employs both an open and a closed loop controller (see Figure 3-6).

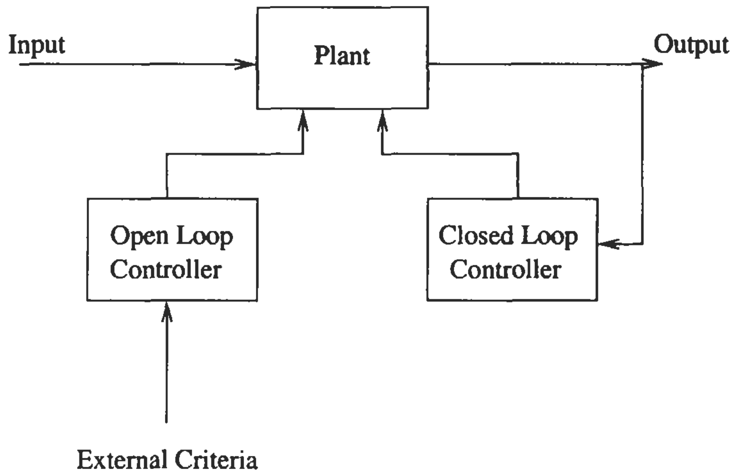

Figure 3-6: Linear Control System

The open loop control system is a power adjustment the mobile makes independent of the system response (which in this case is the received SNR at the base station). In the IS-95 spread spectrum wireless system, for instance, this is simply the received in-band power estimate at the mobile station; this value provides a rough estimate of the propagation path loss between the mobile and base station. However, this is not sufficient for power control because the mobile transmits and receives on different frequencies. Therefore, a closed loop is necessary for power control. This is a mechanism in which the power adjustment is based on the response of the system, in this case the received SNR at the base station. Closed loop power control is comprised of two mechanisms: *inner loop* power control and *outer loop* power control

In the inner loop power control mechanism, the base station simply sends back commands to tell the mobile to increase or decrease its radiated power appropriately. This is sometimes known as *bang-bang* control. As an example, the IS-95 system only uses 1-bit commands to accomplish this; this results in a limited range of adjustment.

The *outer loop* power control mechanism entails the process of determining an accurate SNR threshold for a given user. An outer loop control mechanism in its strictest sense seeks to find the necessary decision thresholds for proper open and closed loop controller operation in a linear control system. The outer loop power control algorithm adjusts the threshold by which the received SNR is measured for formulation of closed loop commands. These thresholds are usually based on a desired setpoint, such as a desired frame error rate (FER). It is well known that there exists a one-to-one correspondence between signal SNR and FER for fixed channel conditions. Existing outer loop power control methods entail adjustment of the threshold at a rate based on the desired FER, with adjustments occurring only after a decision is made at the receiver whether or not the current received frame is in error. The adjustment should be larger in response to a received frame in error than for a frame received correctly. For instance, an algorithm described in [3] provides a method for adjusting the target SNR threshold, denoted by $(E_b/N_t)_T$, by the following algorithms:

1. For frame j, j=0,1,..., adjust $(E_b/N_t)_T$, using fixed step-size Δ
2. If frame in error, $(E_b/N_t)_T(j+1) = (E_b/N_t)_T(j) + K\Delta$
3. If frame not in error, $(E_b/N_t)_T(j+1) = (E_b/N_t)_T(j) - \Delta$

K is an integer, which may be chosen by the rule [1 / (required frame error rate) — 1]. This algorithm tries to ensure that for every $K + 1$ frames, one frame will be in error.

4.1 SNR Estimation at the Receiver

Estimation of the SNR at the receiver is an important step in the proper functioning of the feedback power control system in CDMA. Following the notation from [4], the received signal for a reference (pilot) channel on the uplink may be represented as:

(3-16)

$$r_{pi}(k) = \alpha_i(k)\sqrt{E_c}\,c(k)A_p a_p(n) + z_i(k)$$

In (3-15), $\alpha_i(k)$ is the complex channel gain of the i-th demodulated multipath (assuming a rake receiver is used), E_c is the energy per chip, $c(k)$ is the complex value of the k-th chip, A_p is the pilot channel amplitude scaling, $a_p(n)$ is the n-th pilot symbol, and $z_i(k)$ is the complex wideband noise (assumed to be Gaussian zero-mean and having variance I_{oi}). Although the only requirement for $a_p(n)$ is that it be known at the receiver, for analysis' sake, it is hereafter assumed to be unity. As a result, signal quality estimates are taken from this pilot channel, due to the fact that it is not data-modulated (since the pilot symbols are known at the receiver). Since it is assumed that the receiver knows the relative powers of the pilot and traffic channels, the signal quality estimate obtained from the pilot channel may be scaled to result in a signal quality estimate for the traffic channel.

In forming the signal quality estimate, several de-spread chips are summed together under the assumption that $\alpha_i(k)$ stays constant over the integration period. It is assumed that since the noise term $z_c(k)$ is zero-mean, the remaining quantity will provide a scaled estimate of the complex signal amplitude. The signal estimate over N chips for the i-th multipath is

(3-17)

$$s_i(n) = \sum_{k=0}^{N-1} c^*(n-k)r_{pi}(n-k) = N\alpha_i(n)\sqrt{E_c}A_p + z_i'(n)$$

The noise term in (3-17), denoted by $z_i'(n)$, may be assumed to be the sum of several independent, identically distributed random variables, and therefore may be assumed to be Gaussian with variance NI_{oi}. The frequency of $s_i(n)$ may be represented in terms of the chipping frequency f_c as

(3-18)

$$f_{s_i} = \frac{f_c}{N}$$

The noise power may be reduced by passing the signal estimate $s_i(n)$ through a lowpass filter $h(n)$ of one-sided noise bandwidth B_n. This will reduce the variance of $z_i'(n)$ by a factor of $f_{s_i}/(2B_n)$. The final resultant signal estimation term after filtering may be denoted by

(3-19)

$$\hat{s}_i(n) = h(n) * s_i(n) = N\alpha_i(n)\sqrt{E_c}A_p + z_i''(n)$$

Returning back to (3-19), if it is assumed that $I_{oi} >> |\alpha_i(k)|^2 E_c A_p{}^2$, then it can be assumed that the second-order moment of $r_{pi}(k)$ is $E\{|r_{pi}(k)|^2\} \cong I_{oi}$. Therefore, if a sample variance is formed over $\left\lfloor \frac{f_c}{B_n} \right\rfloor$ samples (where $\lfloor \phi \rfloor$ denotes the nearest integer value to ϕ which is not greater than ϕ), in other words matching the rate at which signal amplitude estimates are formed, then the (unbiased) estimator would be defined as

(3-20)

$$\hat{I}_{oi} = \frac{B_n}{f_c - B_n} \sum_{k=0}^{\lfloor f_c / B_n \rfloor - 1} |r_{pi}(k)|^2$$

Finally, an SNR estimate may be formed as

(3-21)

$$SNR_i(n) = \frac{|\hat{s}_i(n)|^2}{\hat{I}_{oi}(n)}$$

This SNR estimate can be scaled to a E_b/N_t measurement based on the processing gain.

5. CONCLUSIONS

An introduction to some of the fundamental aspects of accurate reception of a CDMA signal was presented in this chapter. This not only includes diversity reception through the use of a rake receiver but also the application of feedback power control to CDMA systems to combat the near-far effect. Along with the material presented in Chapter 2, the reader should have a sufficient background for understanding the more detailed descriptions of existing CDMA cellular systems in the next three chapters.

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Chapter 4

An Overview of IS-95 and cdma2000

Key Topics: IS-95, cdma2000

Overview: The cdma2000 system is the third-generation extension of the existing IS-95 CDMA system currently deployed worldwide. This chapter presents an overview of the cdma2000 air interface, starting from the original derivation of IS-95 and leading into the enhancements that make up cdma2000.

1. INTRODUCTION

Qualcomm Incorporated, beginning in the mid-1980s, developed the first CDMA system deployed for commercial use in the cellular band. This system was considered an attractive alternative to the existing FDMA technologies (AMPS, primarily) and TDMA systems (IS-54, IS-136, GSM) that were in use during this era. One of most interesting potential benefits of CDMA was the potentially enhanced voice capacity when compared to competing wireless cellular systems.

However, previous work had already been performed in the area of spread spectrum communications as applied to cellular systems. For instance, Prof. George Cooper's work at Purdue University in the late 1970s also covered many of the same issues Qualcomm had to overcome in making CDMA viable for mass commercial deployment (for instance, [4]). In particular, Cooper recognized the need for some type of power control system for mobile users to overcome the near-far effect prevalent in CDMA systems.

While Qualcomm had the benefit of Cooper's previous work when laying the foundation for a commercially viable CDMA system, the need for a practical power control system and a means for mobility were still critical. In AMPS and TDMA systems, mobiles could roam from one base station to another based on the fact that each neighboring base station occupied different frequencies on both the transmit and receive bands. As a result, such systems are hampered by the *frequency reuse factor*, which places a practical limit on the minimum number of neighboring base stations in a hexagonal grid that may use different frequencies. CDMA, on the other hand, was supposed to avoid these types of problems due to the fact that all base stations within a network could use the same frequency and any interference that one CDMA base station imposed on another could be suppressed in despreading the signal. However, this posed a problem: how does a mobile move from one base station to another without breaking the connection? Recall that mobiles in a CDMA system in general are in continuous communication with the base station. However, in a TDMA system, mobiles may monitor or initiate contact with other base stations prior to a handover during the periods of time when the mobile is not directly communicating with the base station (i.e., "break before make"). CDMA systems solve this problem with the concept of soft handoff, which entails the mobile being in simultaneous communication with several base stations at once. Once the mobile roams from one base station to another, there is no need to suspend communication with either base station in the process. This feature also leads to some diversity benefits and potentially lowers the probability of dropped calls as a result of mobility.

1.1 Standardization History

Qualcomm's innovations in the area of CDMA for cellular systems resulted in the Telecommunications Industry Association (TIA) developing the IS-95 standard [2]. This standard formed the basis for the first CDMA systems deployed in the cellular band (from 800 to 900 MHz) in North America. This development eventually led to the TIA working with the T1P1 to develop the J-STD-008 [3] standard for the PCS band (from 1800 to 1900 MHz). Since then, there has been some effort to enhance symmetric data rates for IS-95, resulting in the formation of a new standard in 1998, IS-95-B.

Since then, the focus has been on developing the third-generation version of IS-95, which was entitled cdma2000. This effort was initiated in response to the International Telecommunications Union's IMT-2000 effort, which was designed to arrive at a global third-generation radio system. The initial development resulted in the submission of the candidate "radio technology text" (RTT) for cdma2000 by the TIA to the ITU in the middle of 1998.

In response to the growing need to address the interests of cdma2000 operators worldwide, the *Third-Generation Partnership Project 2* (3GPP2) was created at the end of 1998. The standardization bodies involved in the formation of the 3GPP2 were the TIA, the Association of Radio Industries and Businesses (Japan), the Telecommunication Technology Committee (Japan), the Telecommunication Technology Association (Korea), and the China Wireless Telecommunications Standards Group (People's Republic of China). The 3GPP2 was placed in charge of developing a global cdma2000-based standard.

In the spring of 1999, a group of cellular operators and vendors worldwide known as the Operators Harmonization Group (OHG) agreed on a global CDMA standard that encompassed cdma2000 and the standard being developed by the Third-Generation Partnership Project (3GPP) for Wideband CDMA (WCDMA). This agreement called for two types of systems: a direct-spread (DS) system and a multicarrier (MC) system. The DS system was based on WCDMA, while the MC system was based on cdma2000.

The 3GPP2 has continued in developing the multicarrier version of cdma2000, not only defining the air interface but also defining the core network and network interfaces. In 1999, work was initiated on the first packet-switched core network for an IS-95–based system in the 3GPP2. In addition, a global A-interface was developed for cdma2000 (Interoperability Specification, or IoS) that allowed for multivendor radio access networks.

The spectral deployment for IS-95 and cdma2000 systems is primarily in the cellular and PCS bands in North America. Both parts of the spectrum are deployed in *paired bands*, meaning that reverse link (uplink) and forward link (downlink) transmission is performed over separate parts of the spectrum that are associated with one another. The cellular band reverse link spectrum allocation is between 824 and 849 MHz, and the associated forward link allocation is between 869 and 894 MHz. These frequencies are also applicable to South Korea, the other major area of IS-95 deployment. Similarly, the PCS band deployment is between 1850 and 1910 MHz for the reverse link, and 1930 and 1990 MHz for the forward link. The PCS frequencies in South Korea and Japan are nearly the same.

2. OVERVIEW OF IS-95 AIR INTERFACE

The IS-95 system was originally designed to work in the cellular band in North America, leading to a necessity for a 1.25 MHz system. This was a result of the 30 kHz center frequency spacing in the cellular band. As a result, certain modulation and spreading parameters were chosen with this limitation in mind.

The IS-95 system achieved channelization in the forward link and reverse link using different means. Channelization in the forward link was accomplished through the use of orthogonal Walsh codes, while channelization in the reverse link was achieved using temporal offsets of the spreading sequence.

Different base stations are identified on the downlink based on unique time offsets utilized in the spreading process. Therefore, all base stations must be tightly coupled to a common time reference. In practice, this is accomplished through the use of the *Global Positioning System* (GPS), a satellite broadcast system that provides information on Greenwich Mean Time and can be used to extract location information about the receiver. This common time reference is known as *system time*.

There are two types of PN spreading sequences used in IS-95: the *long* code and the *short* codes. Both the PN sequences are clocked at 1.2288 MHz, which is the chipping rate. Two short code PN sequences are used because IS-95 employs quadrature spreading. These two codes are the in-phase sequence $P_I(x) = x^{15} + x^{13} + x^9 + x^8 + x^7 + x^5 + 1$ and the quadrature sequence $P_Q(x) = x^{15} + x^{12} + x^{11} + x^{10} + x^6 + x^5 + x^4 + x^3 + 1$. These two sequences are generated length-15 shift register sequences; although they are nominally $2^{15} - 1 = 32767$ chips, a binary '0' is inserted in each sequence after a string of fourteen consecutive 0's appears in either sequence to make the final length of the spreading sequence an even 32768 chips.

The long code is given by the polynomial $p(x) = x^{42} + x^{35} + x^{33} + x^{31} + x^{27} + x^{26} + x^{25} + x^{22} + x^{21} + x^{19} + x^{18} + x^{17} + x^{16} + x^{10} + x^7 + x^6 + x^5 + x^3 + x^2 + x^1 + 1$. It is of length $2^{42} - 1$ chips as it is generated by a 42-length shift register. It is primarily used for privacy, as each user of the mobile network may be assigned a unique temporal offset for the long code with reference to system time. Since the long code has a period of 41 1/2 days, it is nearly impossible to blindly detect a user's temporal offset. The offset is accomplished with the use of a long code mask, which is a 42-bit value that is combined with the shift register state using a logical AND operation. The modulo-2 sum of the 42 bits which result from this AND operation provides a time-shifted version of the long code sequence.

The particulars of the IS-95 air interface will be presented in the ensuing sections. In order to facilitate description of the air interface, the forward link and reverse link descriptions will be presented separately.

2.1 IS-95 Forward Link

The IS-95 forward link is designed to take advantage of the inherent ability of CDMA systems to use a frequency reuse factor of 1. Moreover, the IS-95 forward link is also designed to achieve coherent reception at mobile receivers by means of a pilot signal.

Channelization by means of code multiplexing is a fundamental feature of IS-95 systems. In particular, channelization is accomplished using length-64 Walsh codes, which are assigned to different channels. Recall from Chapter 2 that Walsh codes can be generated recursively; using this approach, all 64 length-64 Walsh codes can easily be generated and indexed according to their row number in the 64 by 64 Walsh matrix.

The types of channels used can be grouped into *common* channels and *dedicated* channels. Common channels are *broadcast* to all the users in the cell served by the base station. Dedicated channels are meant to be heard by only one user.

2.1.1 Common Channels

The three types of common channels used in IS-95 are the Pilot, Sync and Paging channels. Each has a unique Walsh code associated with it, and serves a particular purpose in the IS-95 forward link.

2.1.1.1 Pilot Channel

The mobile uses the pilot channel for the following purposes:

a. Multipath channel amplitude estimation for coherent detection

b. Timing recovery for synchronization to network time reference (GPS-based)
c. Frequency offset correction for the mobile receiver
d. Pilot strength measurements for soft and hard handoff decisions

There are also several other possible uses for the pilot at the mobile receiver, such as interference correction and interfrequency handoff measurements.

The pilot channel must be a known sequence to be useful at the mobile station. In this case, the pilot channel is simply all binary 0's (see Figure 4-1).

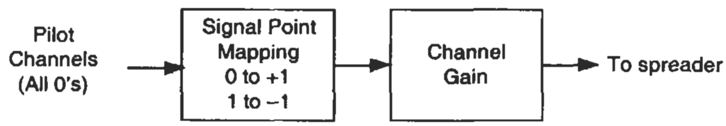


Figure 4-1: IS-95 Pilot Channel

The pilot channel undergoes orthogonal modulation with Walsh code 0, which is the first row of the Walsh-64 matrix and is the all binary 0's code. Since the orthogonal modulation of a binary 0 with a binary 0 is a binary 0, this actual operation does not have to be carried out in the spreading process. Instead, a stream of binary 0's is transmitted at the chipping rate for the pilot channel.

The pilot channel must be transmitted at a sufficiently high power so that mobiles at the cell boundaries can still receive it. As a result, the pilot must occupy a significant amount of base station transmitter power (typically 20% of the total power).

2.1.1.2 Sync Channel

The sync channel is primarily used by the mobile to acquire a timing reference. The mobile station, when it acquires the pilot channel, knows the PN timing of that particular base station. However, the mobile does not know how the timing of this base station relates to other base stations in the network.

Recall that an IS-95 system requires base stations to transmit at fixed time offsets from GPS-based time. This synchronization to system time ensures that one base station's signal does not interfere with another, as the partial correlation properties of the PN sequences used will allow the mobile to despread the desired base station and suppress other base station signals.

The Sync Channel Message appears on the sync channel to let the mobile know timing parameters such as the PN timing offset of the base station relative to system time. The bit rate of this message is 1.2 kbps. This message is then convolutionally encoded, repeated, and block interleaved. The block interleaver depth is a function of the sync channel frame duration, which is 26 2/3 ms.

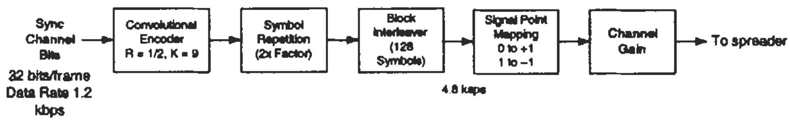


Figure 4-2: IS-95 Sync Channel

The sync channel undergoes orthogonal modulation via the length-64 Walsh code with row index 32.

2.1.1.3 Paging Channel

Up to 7 paging channels, each with its own unique Walsh code, may be used by the IS-95 base station. This channel provides system parameters, voice pages, short message services, and any other broadcast messaging to users in the cell. The paging channel can take two bit rates, 4800 bps or 9600 bps. The rate is given in the Sync Channel Message. These paging channels bits are then convolutionally encoded, repeated, and block interleaved. The block interleaver depth is a function of the paging channel frame duration, which is 20 ms.

The paging channel interleaver output bits are then scrambled. "Scrambling" entails a modulo-2 addition of the input bit and a bit from a predetermined sequence. In this case, the predetermined sequence is the long code generator sequence with a mask unique to the particular paging channel being used. This sequence is decimated from the nominal 1.2288 MHz rate to the necessary 19.2 kHz rate for scrambling by simply taking every 64th bit from the masked long code generator output.

The basic modulation is shown in Figure 4-3.

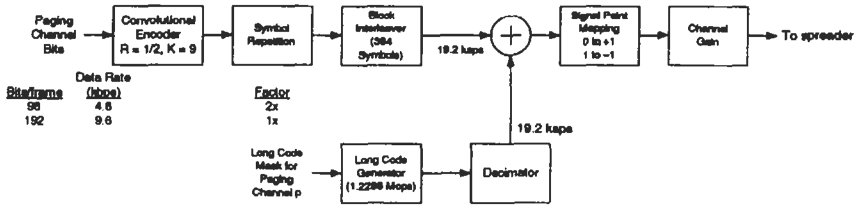


Figure 4-3: IS-95 Paging Channel

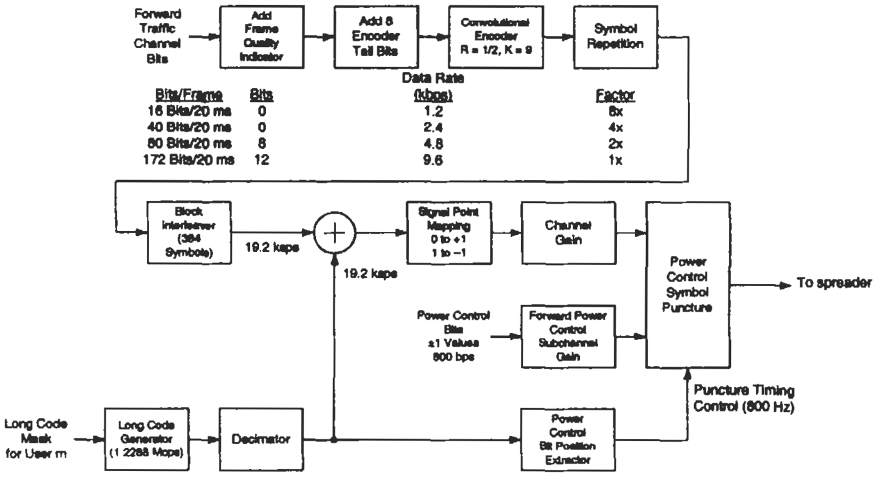
2.1.2 Dedicated Channels

Dedicated channels deliver user traffic and user-specific signaling. There are two types of dedicated channels used in IS-95: the forward fundamental channel and the forward supplemental code channel. The forward fundamental channel was simply called the forward traffic channel in IS-95-A, as it was the only channel capable of delivering dedicated traffic. In IS-95-B [4], the forward supplemental code channel was introduced as a means of improving data rates to individual users.

Voice always goes over a fundamental channel and can never go over a supplemental code channel. However, data may travel over both types of channels.

The fundamental channel is variable rate. This is to take advantage of periods of time when the voice activity is low and therefore the voice codec (i.e., *co*der/*dec*oder) rate may be reduced. In IS-95 systems, voice codecs generally take four rates, sometimes denoted as full rate, half-rate, quarter-rate, and eighth-rate. The first IS-95 systems used source rates of 9.6 kbps, 4.8 kbps, 2.4 kbps, and 1.2 kbps; this set of data rates is known as *Rate Set 1*. These data rates were necessary for the first 8-kbps IS-95 vocoder (i.e., *vo*ice en*coder*), known as QCELP8. *Rate Set 2* (14.4 kbps, 7.2 kbps, 3.6 kbps, 1.8 kbps) was introduced to accommodate a 13-kbps vocoder known as QCELP13.

The supplemental code channel is not variable-rate, yet can take either 9.6 or 14.4 kbps forms. This channel is primarily used for providing higher data rates to individual users through the use of *code channel aggregation*, where an individual user is assigned several supplemental code channels (up to 7) to increase data throughput. The modulation streams for the two rates sets are shown in Figure 4-4 and Figure 4-5.



Power control bits are not punctured in for Forward Supplemental Code Channels.

Figure 4-4: Rate Set 1 Traffic Channel

Note that each frame is appended with a frame quality indicator, which is a cyclic redundancy check (CRC) that can be used by the receiver for error detection. In addition, each frame is appended with 8 "tail bits," which are binary 0's. The purpose of these bits is to flush the convolutional encoder and return it to the all-0 state at the end of each frame. This is helpful in the decoding process, as each frame can be decoded individually at the receiver.

Note also that a combination of symbol repetition and puncturing is used to keep the input to the block interleaver always at 384 symbols at a time. However, for lower data rates, the corresponding channel gain reduces as the data rate reduces. For instance, if the block interleaver output for full-rate symbols is transmitted with power E_s, the transmit power for a half-rate frame is $E_s/2$, quarter-rate is $E_s/4$, and eighth-rate is $E_s/8$.

The power control subchannel is punctured into the transmitted frame on the fundamental channel only. These are one-bit power control commands punctured at an 800 Hz rate. The puncturing location is randomized based on the long code state. For the previous power control interval (known as a *power control group*), its duration is 1.25 ms.

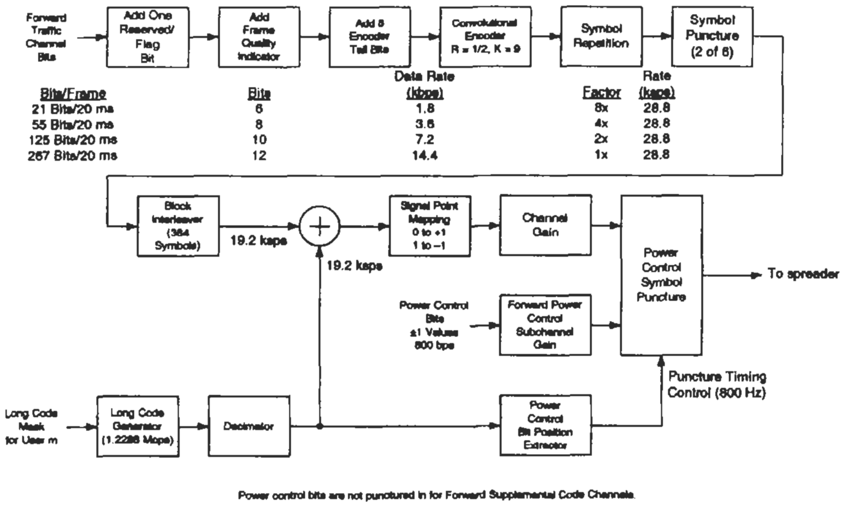


Figure 4-5: Rate Set 2 Traffic Channel

2.2 IS-95 Reverse Link

The IS-95 reverse link channels may also be grouped into common and dedicated channels. The common channels in the IS-95 reverse link are meant primarily for tasks such as call origination, registration and authentication, page responses, and delivery of SMS.

Channelization of users in the reverse link is accomplished by the use of long code masks. Recall that each mobile must acquire a system time reference based on the pilot signal it receives from the base station and the associated sync channel information. Therefore, each mobile can utilize a unique long code mask assuming that the mobile's long code generator is synchronized with the long code generator being used by the base station. As a result, the mobile may transmit with a unique long code mask known only to the base station. In reality, due to propagation delays and imperfect timing references at the mobile stations, the base station must also examine other timing offsets near what the mask value indicates when acquiring an individual user. However, this process is still far less complex than if the base station had to blindly acquire the mobile's timing offset.

2.2.1 Common Channels: Reverse Access Channel

The reverse access channel (R-ACH) is the reverse link common channel in IS-95. The basic transmission operations are depicted in Figure 4-6.

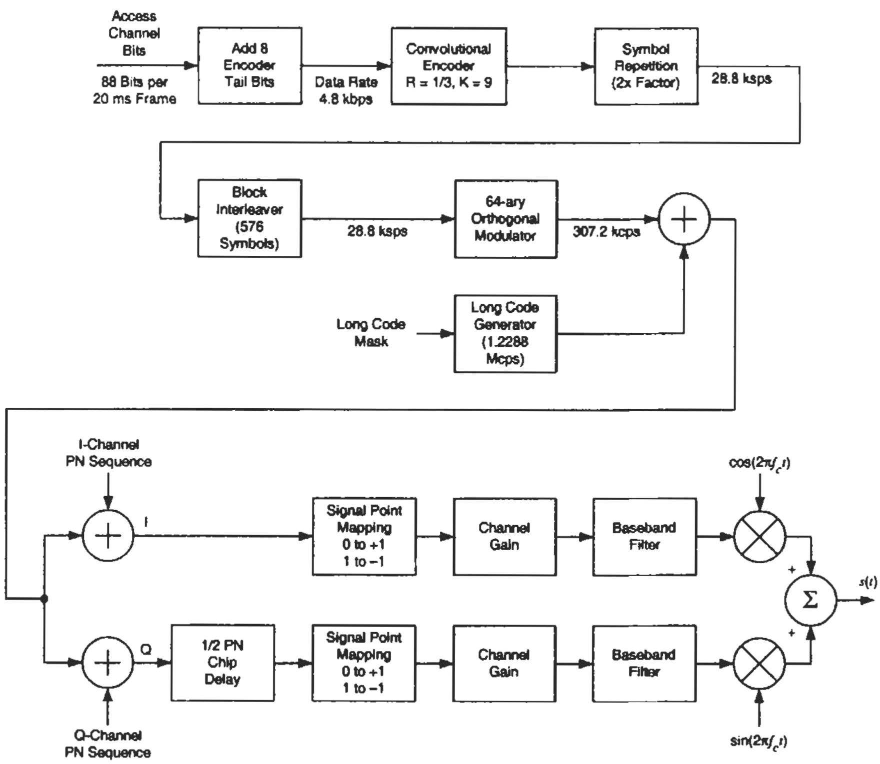


Figure 4-6: IS-95 Reverse Access Channel

The R-ACH messaging is at 4.8 kbps. It is convolutionally encoded, repeated, and interleaved over 576 symbols. Note that the next step is *64-ary orthogonal modulation*. This step entails grouping each set of 6 consecutive bits output from the interleaver into a row address to a memory that contains the 64 by 64 Walsh matrix. Once a row is selected, all 64 bits that make up the row entry are output at a rate of 307.2 kHz. Since the mobile is not transmitting a pilot signal on the reverse link (unlike the forward link), coherent detection is not possible. As a result, the base station receiver may correlate the received signal with all of the 64 possible Walsh codes and determine a peak correlation to determine which row was sent. This operation does not require an estimate of the channel amplitude, but receiver performance is worse than if a pilot signal was used.

After orthogonal modulation, the sequence is spread to 1.2288 MHz by the long code; the long code generator state should be synchronized with the base station long code generator based on the information the mobile has received from the sync channel. The signal may now be quadrature spread; however, note the 1/2-chip delay in the Q-branch of the quadrature spreader. This results in *offset-QPSK* modulation. Offset QPSK modulation reduces the *peak-to-average ratio* (PAR) in the signal the mobile must transmit. Reducing PAR reduces the dynamic range one must design a mobile transmitter over, which generally results in simpler design.

2.2.1.1 R-ACH Timing

R-ACH timing is critical, as several mobiles may try to send R-ACH messages simultaneously. As a result, R-ACH messaging is sent in the form of *access probes*. The mobile sends an access probe aligned with system time and waits for a response from the base station on the forward paging channel. If it does not get a response before a timer expires, it sends another probe at a power greater than the previous probe. The power difference between the probes is a fixed step size measured in decibel units.

In order to reduce the probability that mobiles send probes simultaneously (i.e., a "collision" occurs), the access probe timing is aligned with system time and a *random backoff*. Based on a set sequence, the mobile transmits an access probe aligned with 20 ms increments of system time but backed off by a time offset based on the results of the algorithm. Since each mobile's algorithm is based on input parameters unique to the mobile, the chances of collision are reduced. The access probe power over time is shown in Figure 4-7.

Sixteen probes are allowed in a sequence before the mobile must "give up" and start the process again at the original power levels.

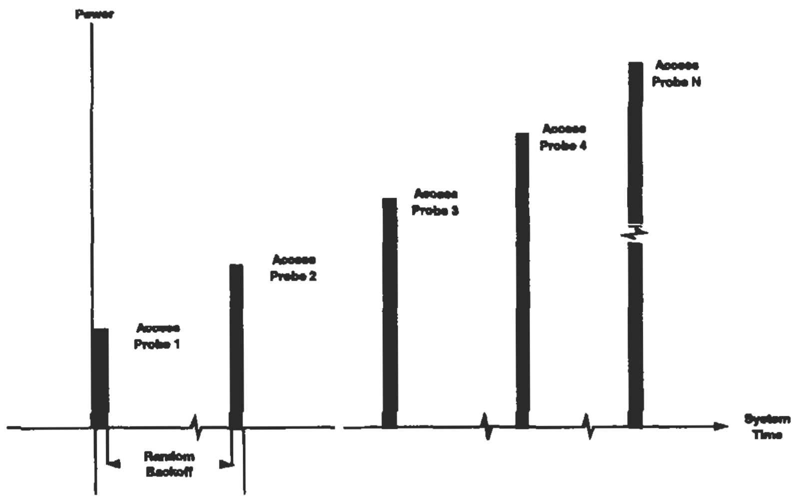


Figure 4-7: Access Probe Timing

2.2.2 Dedicated Channels

As in the forward link, reverse fundamental and supplemental channels are still applicable. The reverse fundamental channel must be able to deliver variable rate data at Rate Sets 1 and 2 for voice services, while the supplemental channels deliver data at full rate. The basic transmission sequences are depicted in Figure 4-8 and Figure 4-9.

The FEC used in the reverse link is rate 1/3 for Rate Set 1, which is different from the forward link that uses rate 1/2. Note also the presence of the *data burst randomizer* for both rate sets. Recall that in the forward link for half-rate, quarter-rate, and eighth-rate frames, symbol repetition was used with power reduction for each transmitted symbol. Although symbol repetition is depicted for the reverse link at the input to the interleaver, in fact only one symbol repetition is actually transmitted. The data burst randomizer actually turns off the transmitter ("gating off" the transmitter) during periods when repetitions are transmitted to ensure that only one symbol repetition is ever actually sent. The pattern with which symbols are eliminated from the transmission sequence is pseudorandom, determined by the state of the long code generator at each power control group. As a result, the base station receiver must be able to detect these on-off transitions, and the mobile must ignore power control commands sent by the base station in response to a gated-off period.

Note also that the sinusoids used to modulate the spread signal to the carrier frequency have an associated phase offset ϕ. This phase offset is unique to each supplemental channel transmitted by the mobile, and can be determined by an index based on the additional number of supplemental channels used (see Table 4-1).

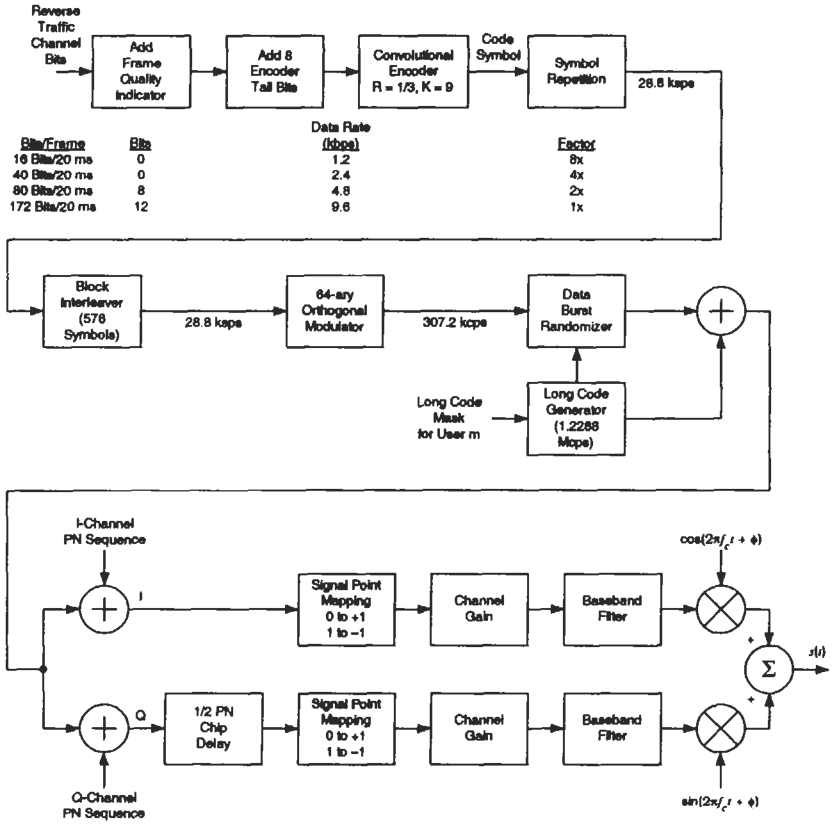


Figure 4-8: Rate Set 1 Reverse Traffic Channels

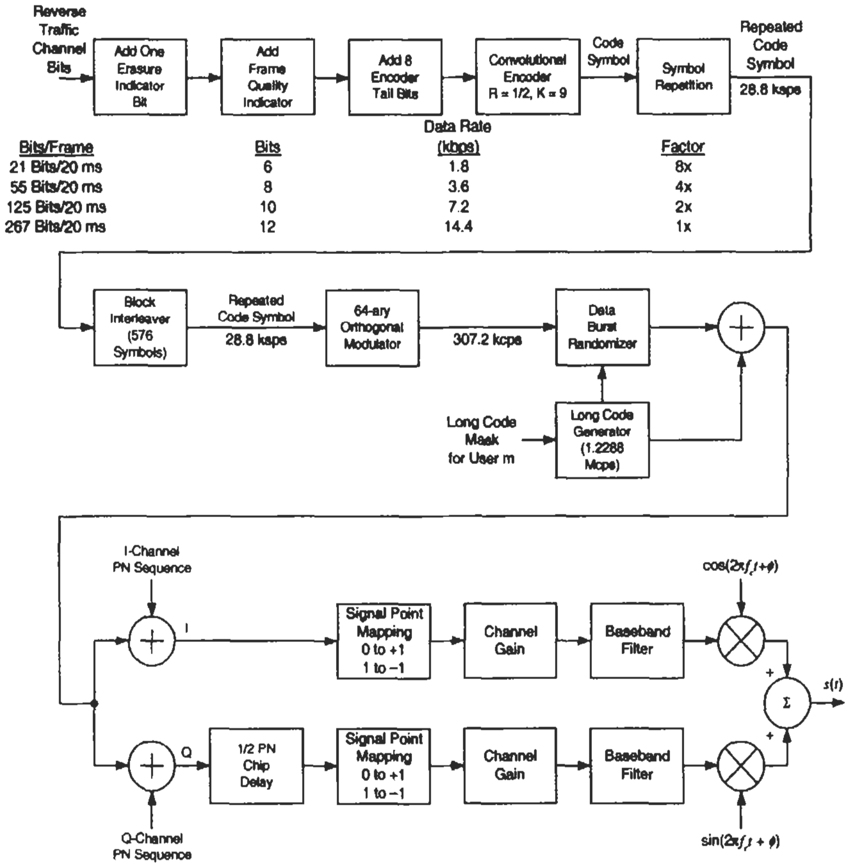


Figure 4-9: Rate Set 2 Reverse Traffic Channels

Additional Channel	Phase Offset from Fundamental Channel
1	$\pi/2$
2	$\pi/4$
3	$3\pi/4$
4	0
5	$\pi/2$
6	$\pi/4$
7	$3\pi/4$

Table 4-1: Reverse Supplemental Channel Phase Offsets

2.3 Baseband Pulse Shaping

The pulse-shaping filter specified in IS-95 is depicted in Figure 4-10. It is nominally a 48-tap finite impulse response (FIR) filter that *does not* satisfy the Nyquist criterion for zero-ISI pulse shaping. However, it does have a spectral characteristic (see Figure 4-11) that provides isolation for the 1.25 MHz transmitted signal. Given that the processing gain in an IS-95 system is 128, the interchip interference introduced by the filter is considered to be negligible.

2.4 Power Control

In IS-95, power control exists for the reverse link and the forward link. Forward link power control generally operates at a slower rate than reverse link power control.

2.4.1 Reverse Link Power Control

Recall from Chapter 3 that power control in CDMA systems normally requires open loop, closed loop, and outer loop control.

Open loop power control is based on measuring the total in-band received power at the transmitter over a sufficiently long duration of time to attain an accurate mean received power estimate. This power estimate is added to a "turnaround constant" to form a base power level at the transmitter. In fact, due to the need for access probes to establish initial contact with the base station, the final open loop estimate also includes the power increments for each consecutive access probe.

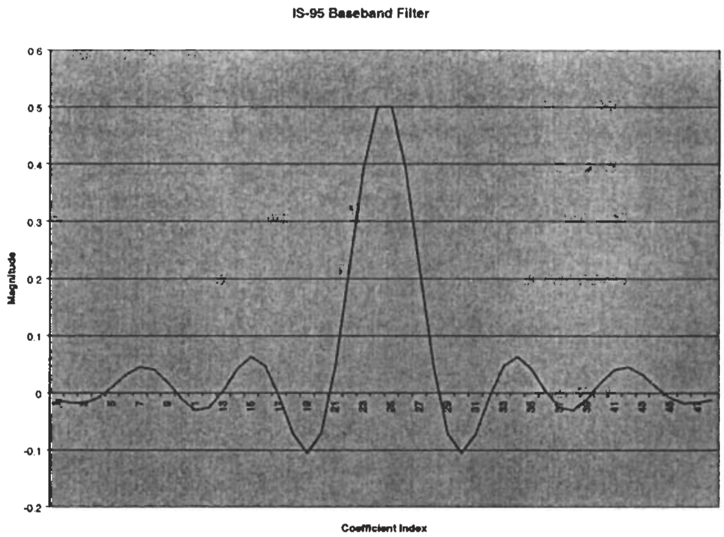


Figure 4-10: IS-95 Baseband Filter

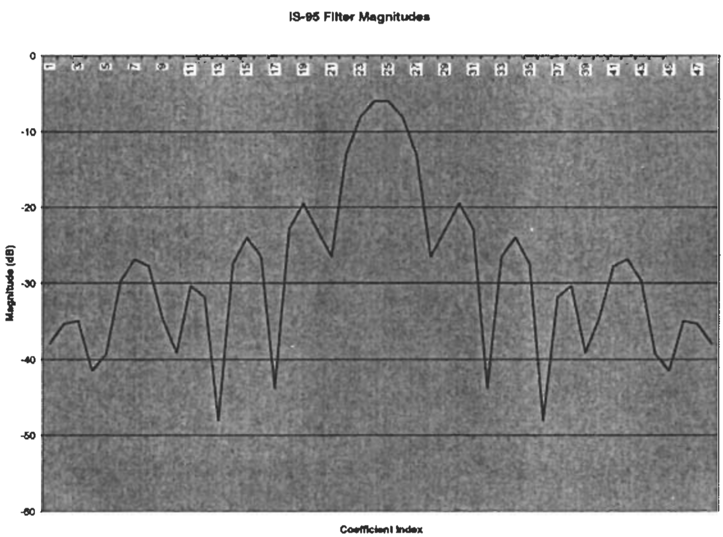


Figure 4-11: Baseband Filter Magnitude Response

Closed loop power control in IS-95 works through the use of power control commands inserted at an 800 Hz rate into the forward fundamental channel (see Figure 4-4 and Figure 4-5). These one-bit commands instruct the mobile to increase or decrease power in 1 dB increments. It should be noted that there are three consequences of this type of power control mechanism:

a. The puncturing of traffic symbols with power control bits actually weakens the error-correcting performance of the convolutional code.
b. The 800 Hz rate at which commands are sent and the 1 dB step size place a restriction as to how small the coherence time (and therefore how large the coherence bandwidth) of the wireless transmission channel can be for power control to be effective.
c. The power control commands themselves are not protected by an error-correcting code. This allows for quick demodulation of power control commands at the handset, but certain channel conditions could lead to very high error rates in the power control commands themselves.

Finally, outer loop power control works in the same fashion as described in Chapter 3. Outer loop algorithms are not specified in wireless standards normally, as they are considered implementation-dependent.

2.4.2 Forward Link Power Control

Forward link power control works at a much slower rate than reverse link power control in IS-95. For Rate Set 1 applications, the only means of implementing forward link power control is through the *Power Measurement Report Message* (PMRM). This is a message that the mobile may be triggered to send if the frame erasure rate it measures on the forward link traffic it receives exceeds a threshold. A mobile classifies a received frame as an erasure when it cannot detect the frame rate or fails to correctly decode the frame quality indicator. Upon receipt of a PMRM, the base station may choose to increase transmitted power to the mobile in question.

The PMRM-based power control mechanism normally works slowly in typical networks (about 3-4 Hz typically). Therefore, if Rate Set 2 is used, in the outgoing fundamental channel frame, the mobile may set an *Erasure Indicator Bit* (EIB). This tells the base station whether the most recently received forward link frame was in error. Since the frame rate is 50 Hz, the forward link power control mechanism under Rate Set 2 is also 50 Hz.

2.5 Soft Handoff

Soft handoff in IS-95 requires the mobile to constantly search for multipaths from different base stations even while actively receiving and sending traffic. Soft handoff can be costly for network capacity if not implemented properly, as each base station that the mobile is in simultaneous contact with must allocate resources for the communications link.

In IS-95, the *active set* is the set of base stations (normally identified by their pilot channels, or simply pilots) with which the mobile may be in simultaneous contact. This set is controlled by the base station based on measurement information relayed by the mobile (in the form of the *pilot strength measurement message* PSMM). The *candidate set* entails pilots that are not in the active set but can be successfully demodulated by the mobile. The maximum number of pilots in either the active or candidate sets is 6. The *neighbor set* is based on predetermined pilots that are in the vicinity of the base station or base stations the mobile is currently in contact with. This may have as many as 20 pilots in it. The *remaining set* covers all other pilots.

A base station is added to the active set not simply because a mobile determines it is of sufficient strength. The mobile alerts the base station with a PSMM when it detects that a pilot has sufficient strength to be added to the active set. The mobile determines sufficient strength by comparing the pilot E_c/I_{or} (chip energy to total in-band received power ratio) to a network-determined threshold, T_ADD. This may cause the pilot to be promoted to the candidate set. In addition, the mobile may compare a candidate pilot with members of the active set to see if it exceeds any of the active set pilots by a threshold (T_COMP); this can result in promotion of the pilot from the candidate set to the active set.

Similarly, the base station must also drop from the active set pilots whose strength has dropped below a certain threshold, T_DROP. This is another event that can trigger the mobile to send a PSMM. Threshold comparisons to drop a pilot are timer based, meaning that the pilot must be below T-DROP for a certain amount of time (given by T_TDROP) before the pilot is demoted.

In addition, the mobile may search for multipaths at certain PN offsets given the network deployment. For instance, if the base station knows the PN offsets of neighboring base stations, it may provide the mobile with search windows. The search windows define a range of time offsets with respect to PN offsets for pilots in the four handoff sets used in IS-95.

2.6 Coding

The type of coding used in IS-95 is convolutional coding (see Section 8.1, Chapter 2). In the forward link, the rate ½ convolutional code discussed in Section 8.1 of Chapter 2 is used (see Figure 2-15, Chapter 2). In the reverse link, a rate 1/3 convolutional code is used (see Figure 4-12).

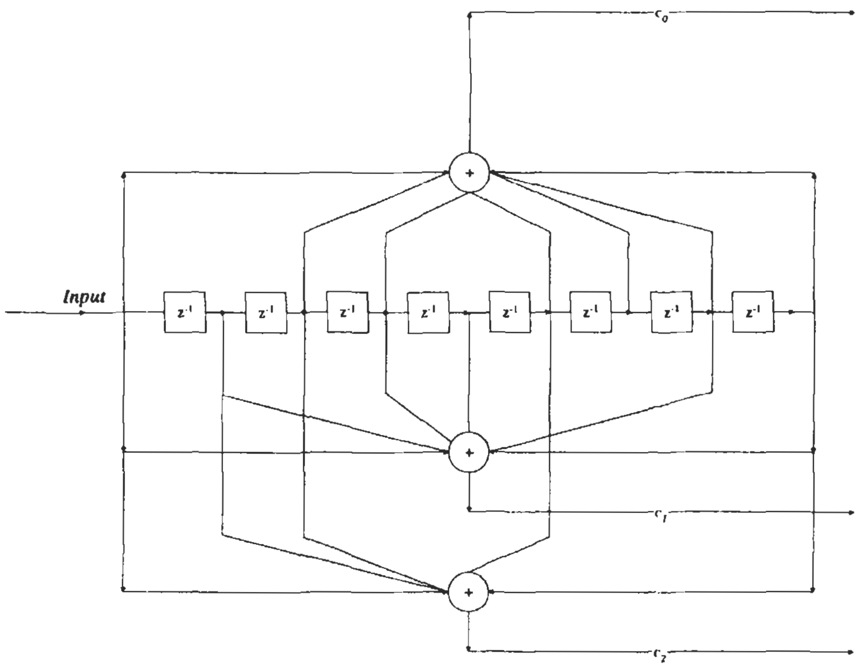


Figure 4-12: Rate 1/3 Convolutional Code

3. CDMA2000

cdma2000 came about as a response to the ITU's IMT-2000 effort for developing global third-generation wireless services (see Chapter 1). The cdma2000 system developed as a result of the TIA's efforts to evolve TIA/EIA-95B. Although the underlying motivation for evolution of TIA/EIA-95-B was to provide the types of services mandated by the ITU for third-generation systems, cdma2000 was developed with the second-generation system in mind. Therefore, an explicit requirement for cdma2000 was backwards compatibility. This requirement ensures that second-generation products can be easily evolved to meet third-generation requirements.

3.1 System Design Issues

3.1.1 Bandwidth

An important design goal for all third-generation proposals is to limit spectral emissions to a 5 MHz dual-sided passband. There are several reasons for choosing this bandwidth. First, data rates of 144 and 384 kbps, the main targets of third-generation systems, are achievable within 5 MHz bandwidth with reasonable coverage. Second, lack of spectrum availability calls for limited spectrum allocation, especially if the system has to be deployed within the existing frequency bands already occupied by the second-generation systems. Third, the 5 MHz bandwidth improves the receiver's ability to resolve multipath when compared to narrower bandwidths, increasing diversity and improving performance.

3.1.2 Chip Rate

Given the bandwidth, the choice of chip rate depends on spectrum deployment scenarios, pulse shaping, desired maximum data rate, and dual-mode terminal implementation. Figure 4-13 shows the relation between chip rate (CR), pulse-shaping filter roll-off factor (α), and channel separation (Δf). If raised cosine filtering is used, spectrum is zero (in theory) after $CR/2*(1+\alpha)$. In Figure 4-13, channel separation is selected such that two adjacent channel spectra do not overlap. Channel separation should be selected this way, if there can be high power level differences between the adjacent carrier. If channel separation is selected in such a way that the spectrum of two adjacent channel signals overlaps, some power leaks from one carrier to another. Partly overlapping carrier spacing can be used, for example in micro cells, where the same antenna masts are used for both carriers.

cdma2000 continues to employ the linear-phase pulse-shaping filter introduced in TIA/EIA-95. This filter complies with electromagnetic compatibility requirements of the United States Federal Communications Commission (FCC).

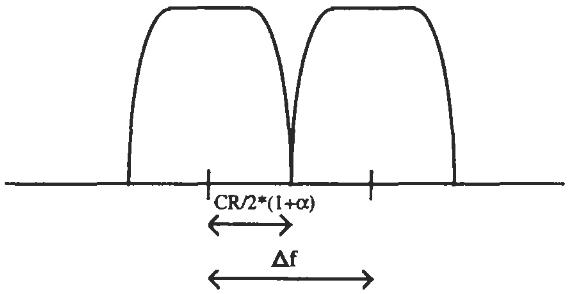


Figure 4-13: Relationship Between Chip Rate (CR), Roll-off Factor (α), and Channel Separation (Δf)

3.1.3 Multirate

Multirate design means multiplexing different connections with different quality of service requirements in a flexible and spectrum-efficient way. The provision for flexible data rates with different quality of service requirements can be divided into three subtopics: how to map different bit rates into the allocated bandwidth, how to provide the desired quality of service, and how to inform the receiver about the characteristics of the received signal. The first problem concerns issues such as multicode transmission and variable spreading. The second problem concerns coding schemes. The third problem concerns control channel multiplexing and coding.

Multiple services belonging to the same session can be either time or code multiplexed as depicted in Figure 4-14. The time multiplexing avoids multicode transmissions, thus reducing peak-to-average power of the transmission. A second alternative for service multiplexing is to treat parallel services completely separately with separate channel coding/interleaving and map them to separate physical data channels in a multicode fashion as illustrated in the lower part of Figure 4-14. With this alternative scheme, the power, and consequently the quality of each service, can be controlled independently.

Time multiplexing

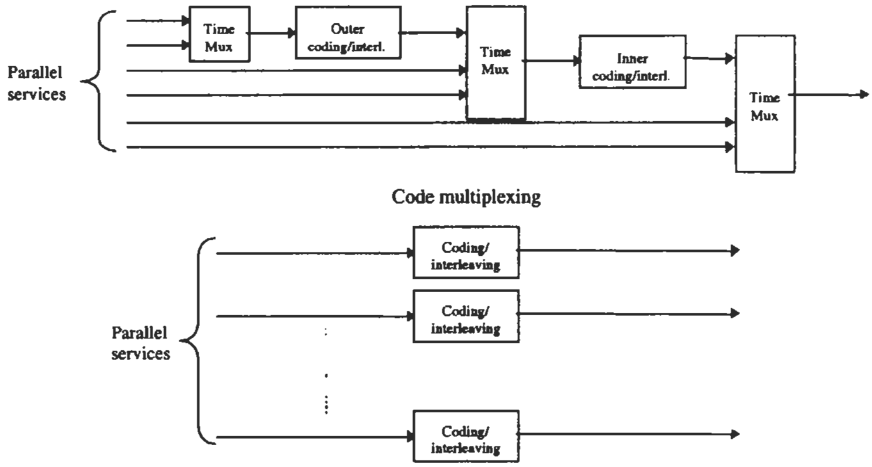


Figure 4-14: Time and Code Multiplexing Principles

The new system, cdma2000, continues to support time multiplexing of services, as introduced in TIA/EIA-95-B in the form of primary and secondary traffic. In addition, multicode transmission is also supported. Although its time-multiplexing capability may be expanded, time multiplexing of service instances on a single physical channel is currently not possible. However, cdma2000 evolution systems (see Chapter 5) try to address this issue.

3.1.4 Spreading and Modulation Solutions

A complex spreading as shown in Figure 4-15, helps to reduce the peak-to-average power and thus improves power efficiency. It is essentially the HQPSK modulation method described in Chapter 2.

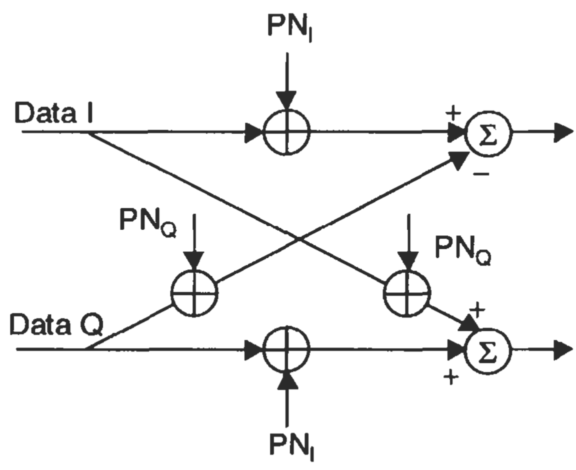


Figure 4-15: Complex Spreading

The spreading modulation can be either balanced or dual channel QPSK. In the balanced QPSK spreading, the same data signal is split into I and Q channels. In dual channel QPSK spreading the symbol streams on the I and Q channels are independent of each other. In the forward link, QPSK data modulation is used in order to save code channels and allow the use of the same orthogonal sequence for I and Q channels. In the reverse link, each mobile station uses the same orthogonal codes; this allows for efficient use of BPSK data modulation and balanced QPSK spreading.

3.1.5 Coherent Detection in the Reverse Link

Coherent detection can improve the performance of the reverse link up to 3 dB compared to the noncoherent reception used by the second-generation CDMA system. To facilitate coherent detection a pilot signal is required. The actual performance improvement depends on the proportion of the pilot signal power to the data signal power and the fading environment.

3.1.6 Fast Power Control in Forward Link

To improve the forward link performance fast power control is used. The impact of the fast power control in the forward link is twofold. First, it improves the performance in a fading multipath channel. Second, it increases the multiuser interference variance within the cell since orthogonality between users is not perfect due to multipath channel. The net effect, however, is improved performance at low speeds.

3.1.7 Soft Handoff

Soft handoff was to remain essentially the same in operation as in IS-95. This was possible due to the fact that cdma2000 is backwards compatible, and therefore existing cells did not need to be redeployed. As a result, handoff mechanisms did not need to change either.

3.1.8 Additional Pilot Channel in the Forward Link for Beamforming

An additional pilot channel on the forward link that can be assigned to a single mobile or to a group of mobiles enables deployment of adaptive antennas for beamforming since the pilot signal used for channel estimation needs to go through the same path as the data signal. Therefore, a pilot signal transmitted through an omnicell antenna cannot be used for the channel estimation of a data signal transmitted through an adaptive antenna.

3.1.9 Transmit Diversity

The forward link performance can be improved in many cases by using transmit diversity. For direct spread CDMA schemes, this can be performed by splitting the data stream and spreading the two streams using orthogonal sequences or switching the entire data stream between two antennas. For multicarrier CDMA, the different carriers can be mapped into different antennas.

3.1.10 Layering

In an effort to create a more layered approach to the protocol stack in cdma2000, functionality was aligned according to the OSI modeling put forth by the International Standards Organization (ISO) [5]. The OSI layers may be described as:

- Application: user access to OSI environment
- Presentation: translates application information to syntax for data communications

- Session: provides control for application-level peer-to-peer communications; responsible for establishing, maintaining, and terminating connections between peer entities
- Transport: provides sufficiently reliable transportation of data between peer entities
- Network: corresponds to underlying network (e.g., wireless network) over which packets from transport layer are transmitted; has own connection establishment and maintenance procedures
- Data link: provides sufficiently reliable transportation of network-level packets over physical layer
- Physical: actual transmission pipe over which raw data flows (e.g., microwave link in wireless network)

This type of layering model is difficult to implement in wireless networks due to functions requiring several layers to coordinate with each other. For instance, in a wireless system, the network layer may be in charge of radio resource control; however, that functionality cannot be totally independent from the physical layer due to functions such as soft handoff. Nevertheless, cdma2000 was structured to include layering for the network layer and below. The basic cdma2000 layers [6] and their corresponding OSI layers are:

- Layer 3 signaling, packet data services, voice data — OSI network layer and up
- Link access control (LAC) — OSI data link layer
- Medium access control (MAC) — OSI data link layer
- Physical layer — OSI physical layer

This has also led to the development of four standards related to these layers [7]-[10]:

1. IS-2000.2: cdma2000 Physical layer
2. IS-2000.3: cdma2000 Medium access control layer
3. IS-2000.4: cdma2000 Link access control layer (Layer 2)
4. IS-2000.5: cdma2000 Upper layer signaling (Layer 3)

4. CDMA2000 PHYSICAL LAYER

The cdma2000 physical layer retains backwards compatibility not only to leverage IS-95 equipment development but also to provide a smooth upgrade path for cellular operators. In this way, cdma2000 systems could be gradually phased into existing IS-95 networks without disrupting service. As a result, many mechanisms such as reverse link power control and soft handoff remain essentially the same from the physical layer standpoint.

The cdma2000 physical layer classifies different modes of operation into *radio configurations* (RCs) for both the forward and reverse links. For instance, Radio Configurations 1 and 2 (RC1 and RC2) are the Rate Set 1 and Rate Set 2 modes of operation respectively in IS-95. However, radio configurations greater than 2 define new modes of operation in cdma2000.

In addition, the cdma2000 radio configurations encompass two modes of operation: 1X and 3X. 1X refers to the mode that is bandwidth-compatible with IS-95, i.e., its bandwidth is 1.25 MHz. 3X refers to the *multicarrier* option, which involves the use of 3 1X carriers to increase the data rate to the mobile user on the forward link. The data rates on the reverse link in the multicarrier version increase data rates via direct spreading up to three times the 1X chip rate of 1.2288 MHz. More recently, modes that involve 3X forward link and 1X reverse link have been adopted to allow for asymmetric high-speed data services.

In this chapter, the 1X option for cdma2000 will be examined. This option is what is supported in the first cdma2000 products, deployed in South Korea in the spring of 2001.

4.1 Forward Link

In the cdma2000, several new code channels are introduced to improve data services. However, backwards compatibility is an issue. In particular, the cdma2000 forward link includes pilot, paging, and sync channels that are identical in operation to their IS-95 counterparts. However, the dedicated channels for RC3 and higher have a different structure.

The basic spreading structure now involves hybrid quadrature spreading (see Chapter 2) with the same baseband filter as in IS-95. This is depicted in Figure 4-16.

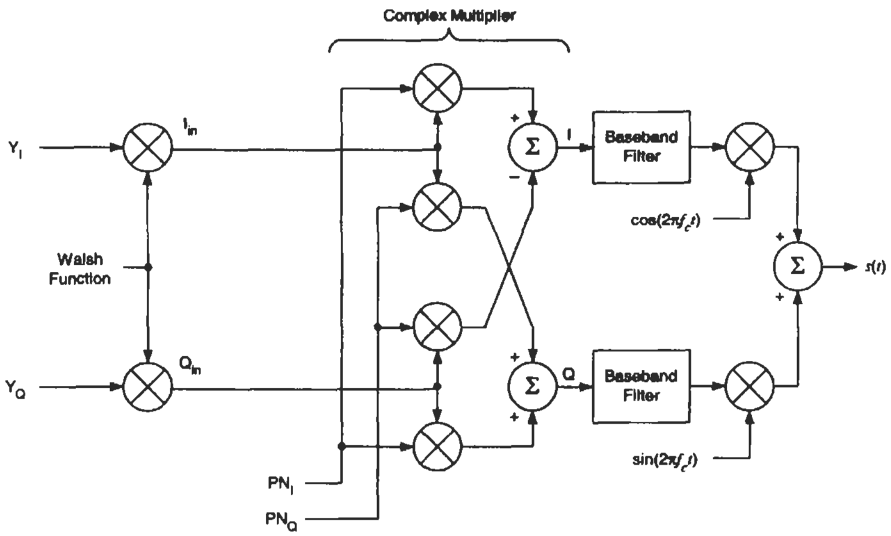


Figure 4-16: cdma2000 Spreading

The spreading function takes as inputs the appropriate Walsh function, and in-phase and quadrature inputs Y_I and Y_Q, respectively. These inputs are derived from a demultiplexing operation (see Figure 4-17). With reference to Figure 4-17, when the input is X, then the output is split into the two streams Y_I and Y_Q by mapping each bit in the input stream to one of the two output streams alternately (starting from Y_I, i.e., a "top-to-bottom" approach). This is a pure QPSK system, as two bits of information are transmitted for each (I,Q) pair of inputs. However, two inputs X_I and X_Q may be mapped directly to Y_I and Y_Q as well. This feature is useful for backwards-compatible channels, as will be seen in the next section.

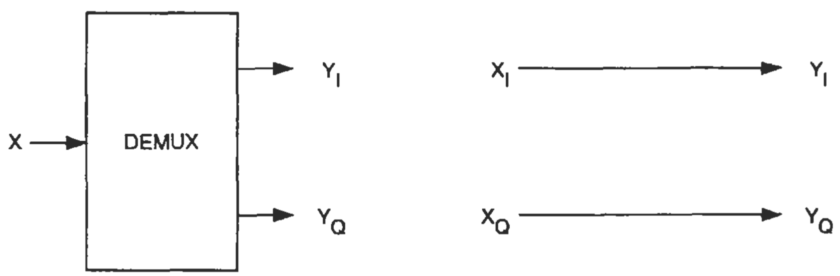


Figure 4-17: Demultiplexing Operation for 1X

4.1.1 Backwards-compatible Common Channels

The IS-95 pilot, paging, and sync channels are still applicable to cdma2000 1X. As seen in Figure 4-18, the basic transmission sequence is the same. However, note that the output after the signal point mapping for each of the channels is mapped only to X_I and zero is mapped to X_Q. When taken as input to the demultiplexer in Figure 4-17, X_I and X_Q are mapped directly to Y_I and Y_Q, meaning 0 is mapped to Y_Q. Therefore, when Y_I and Y_Q are provided as inputs to the spreader in Figure 4-16, then the hybrid quadrature spreading simplifies to basic quadrature spreading. As a result, both IS-95 and cdma2000 mobile stations can read these channels. Similar mappings occur for RC1 and RC2 to provide a means for cdma2000 base stations to support IS-95 mobiles.

Note that in cdma2000 nomenclature, abbreviations are given to designate channels. Thus the forward pilot, sync, and paging channels can also be referred to as FPICH, FSYNC, and FPCH respectively.

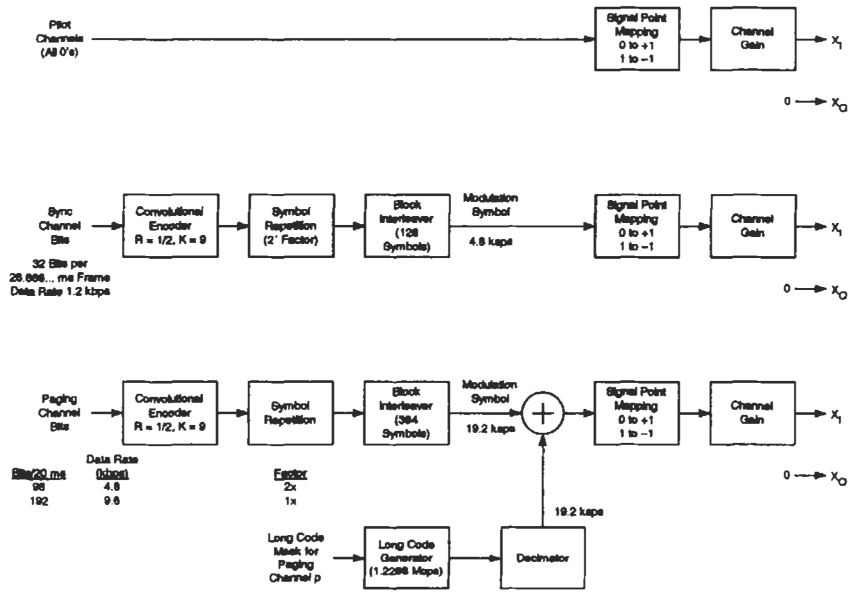


Figure 4-18: cdma2000 Backwards-Compatible Common Channels

4.1.2 New cdma2000 1X Common Channels

Several new common channels exist for cdma2000. The forward common control channel (FCCCH) and forward broadcast control channel (FBCCH) may be used for carrying common signaling much like the paging channel. These channels may be used specifically for cdma2000 mobiles, thus relieving some of the overhead on the paging channel.

The forward common auxiliary pilot channel (F-CAPICH) is used to assist in sport coverage and transmit diversity applications. The Walsh code assignment for an auxiliary pilot can be of length 128, 256, or 512.

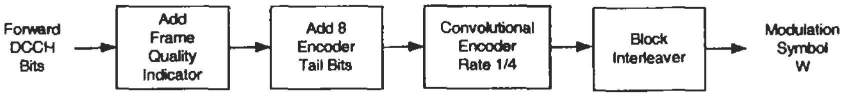
The forward quick paging channel (FQPCH) is an on-off keyed (OOK) indicator to the mobile to wake up for a message on the paging channel. This is useful for conserving battery life in the mobile when it is idling.

In addition, for new access modes on the reverse link, a forward common power control channel (FCPCH) exists in which users' power control bits may be slotted in time.

4.1.3 New cdma2000 1X Dedicated Channels

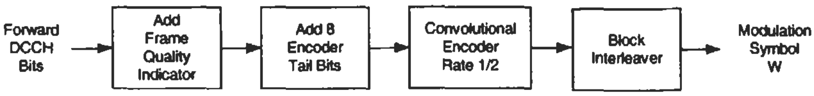
cdma2000 1X introduces the forward dedicated common control channel (FDCCH) and forward supplemental channel (FSCH). The FDCCH may be used primarily for signaling and can be used for other high priority (non-voice) traffic such as retransmissions for data protocols. The FSCH is strictly for data traffic, and can take much higher rates than an IS-95 compatible forward supplemental code channel can. The forward fundamental channel (FFCH) is still primarily for voice, although the lowest data rates for Rate Set 1–compatible data rates changed slightly from 1.2 and 2.4 kbps to 1.5 and 2.7 kbps, respectively.

RC3 and RC4 in general are Rate Set 1–compatible, while RC5 is Rate Set 2–compatible. Note two notable exceptions regarding framing; now the dedicated channels can carry 5 ms frames for short messages, and the FFCH and FSCH can have longer framing (40 and 80 ms) in addition to 20 ms. The transmission sequences for the FDCCH are shown in Figure 4-19 through Figure 4-21. The transmission sequences for the FFCH and FSCH are shown in Figure 4-22 through Figure 4-24.



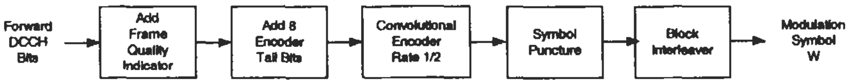
Bits/Frame	Bits	Data Rate (kbps)	Symbols	Rate
24 Bits/5 ms	16	9.6	192	38.4
172 Bits/20 ms	12	9.6	768	38.4

Figure 4-19: FDCCH for RC3



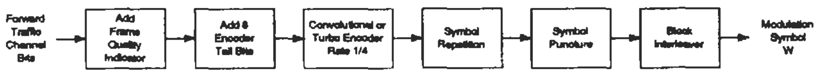
Bits/Frame	Bits	Data Rate (kbps)	Symbols	Rate
24 Bits/5 ms	16	9.6	96	19.2
172 Bits/20 ms	12	9.6	384	19.2

Figure 4-20: FDCCH for RC4



Bits/Frame	Bits	Data Rate (kbps)	Deletion	Symbols	Rate
24 Bits/5 ms	16	9.6	None	192	38.4
267 Bits/20 ms	12	14.4	4 of 12	768	38.4

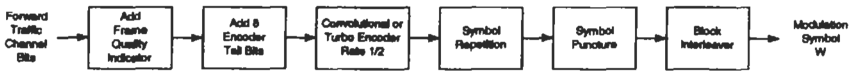
Figure 4-21: FDCCH for RC5



Bits/Frame	Bits	Data Rate (kbps)	Factor	Deletion	Symbols	Rate (ksps)
24 Bits/5 ms	16	9.6	1x	None	192	38.4
16 Bits/20 ms	6	1.5	8x	1 of 5	768	38.4
40 Bits/20n ms	6	2.7/n	4x	1 of 9	768	38.4/n
80 Bits/20n ms	8	4.8/n	2x	None	768	38.4/n
172 Bits/20n ms	12	9.6/n	1x	None	768	38.4/n
360 Bits/20n ms	16	19.2/n	1x	None	1536	76.8/n
744 Bits/20n ms	16	38.4/n	1x	None	3072	153.6/n
1512 Bits/20n ms	16	76.8/n	1x	None	6144	307.2/n
3048 Bits/20n ms	16	153.6/n	1x	None	12288	614.4/n

n=length of frame in multiples of 20 ms: n can equal 1, 2 or 4

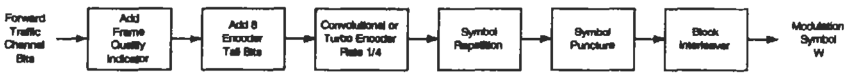
Figure 4-22: FFCH and FSCH for RC3



Bits/Frame	Bits	Data Rate (kbps)	Factor	Deletion	Symbols	Rate (ksps)
24 Bits/5 ms	16	9.6	1x	None	96	19.2
16 Bits/20 ms	6	1.5	8x	1 of 5	384	19.2
40 Bits/20n ms	6	2.7/n	4x	1 of 9	384	19.2/n
80 Bits/20n ms	8	4.8/n	2x	None	384	19.2/n
172 Bits/20n ms	12	9.6/n	1x	None	384	19.2/n
360 Bits/20n ms	16	19.2/n	1x	None	768	38.4/n
744 Bits/20n ms	16	38.4/n	1x	None	1536	76.8/n
1512 Bits/20n ms	16	76.8/n	1x	None	3072	153.6/n
3048 Bits/20n ms	16	153.6/n	1x	None	6144	307.2/n
6120 Bits/20n ms	16	307.2/n	1x	None	12288	614.4/n

n=length of frame in multiples of 20 ms; n can equal 1, 2 or 4

Figure 4-23: FFCH and FSCH for RC4



Bits/Frame	Bits	Data Rate (kbps)	Factor	Deletion	Symbols	Rate (ksps)
24 Bits/5 ms	16	9.6	1x	None	192	38.4
21 Bits/20 ms	6	1.8	8x	4 of 12	768	38.4
55 Bits/20n ms	8	3.6/n	4x	4 of 12	768	38.4/n
125 Bits/20n ms	10	7.2/n	2x	4 of 12	768	38.4/n
267 Bits/20n ms	12	14.4/n	1x	4 of 12	768	38.4/n
552 Bits/20n ms	16	28.8/n	1x	4 of 12	1536	76.8/n
1128 Bits/20n ms	16	57.6/n	1x	4 of 12	3072	153.6/n
2280 Bits/20n ms	16	115.2/n	1x	4 of 12	6144	307.2/n
4584 Bits/20n ms	16	230.4/n	1x	4 of 12	12288	614.4/n

n=length of frame in multiples of 20 ms, n can equal 1, 2 or 4

Figure 4-24: FFCH and FSCH for RC5

The transmission sequences may now be scrambled and inserted with power control commands. Referring to Figure 4-19 through Figure 4-24, the transmission sequences all result in a stream of modulation symbols W. These symbols are then input into the multiplexing structure in Figure 4-25. Note that the mobile may receive power control commands either on an FFCH or FDCCH (however never on both). With reference to Figure 4-25, the output symbol X may now be used as input to the demultiplexer structure of Figure 4-17.

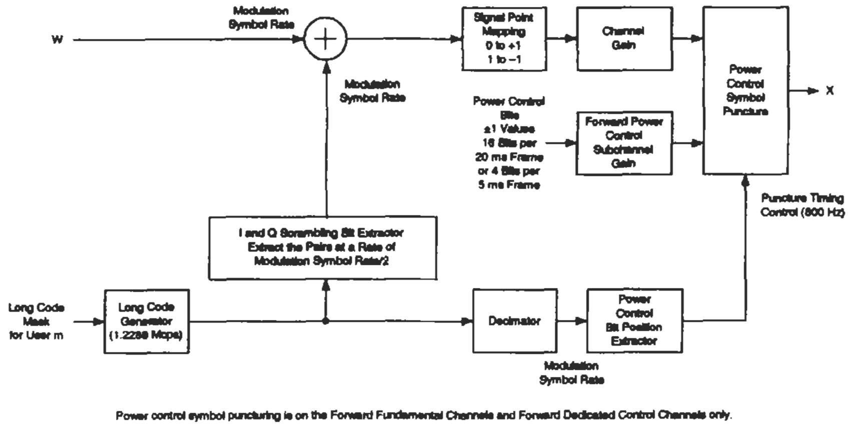


Figure 4-25: Scrambling and Power Control Multiplexing

4.1.4 Transmit Diversity

A new feature in cdma2000 that was not present in IS-95 is the capability to transmit over 2 antennas in the forward link, also known as *transmit diversity*. All channels except for pilot channels, FPCH, and FSYN may be transmitted in this fashion. Channels that are transmitted using the transmit diversity techniques of cdma2000 are demultiplexed differently from the demultiplexing shown in Figure 4-17; the transmit diversity demultiplexing procedure is shown in Figure 4-26. The demultiplexing operation is still top-to-bottom (see Section 4.1).

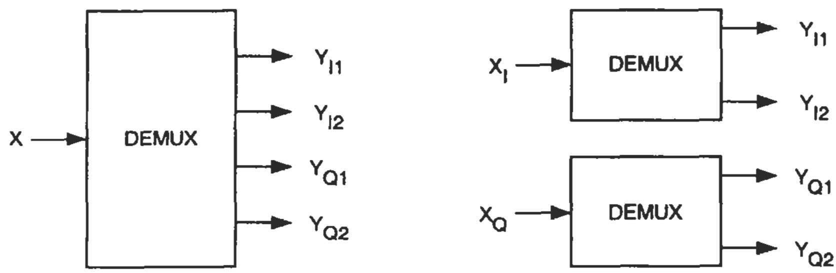


Figure 4-26: Transmit Diversity Demultiplexing

There are two methods of transmit diversity in cdma2000: *space-time spreading* (STS) and *orthogonal transmit diversity* (OTD). The STS spreading operation [11] is shown in Figure 4-27. Assuming that the receiver is perfectly synchronized to the transmitted signal and despreading is done properly, then the received symbols will have been subject to a complex channel gain associated with each of the two antenna; these two values will be denoted as $h_1(t)$ and $h_2(t)$. The receiver can find an estimate of these two values based on the pilot channel transmitted only over the primary antenna and an auxiliary pilot transmitted only over the second antenna. Using these values and the receiver structure in Figure 4-28, the symbols input to the space-time spreader may be recovered with additional diversity gain from the transmission paths from both antennas.

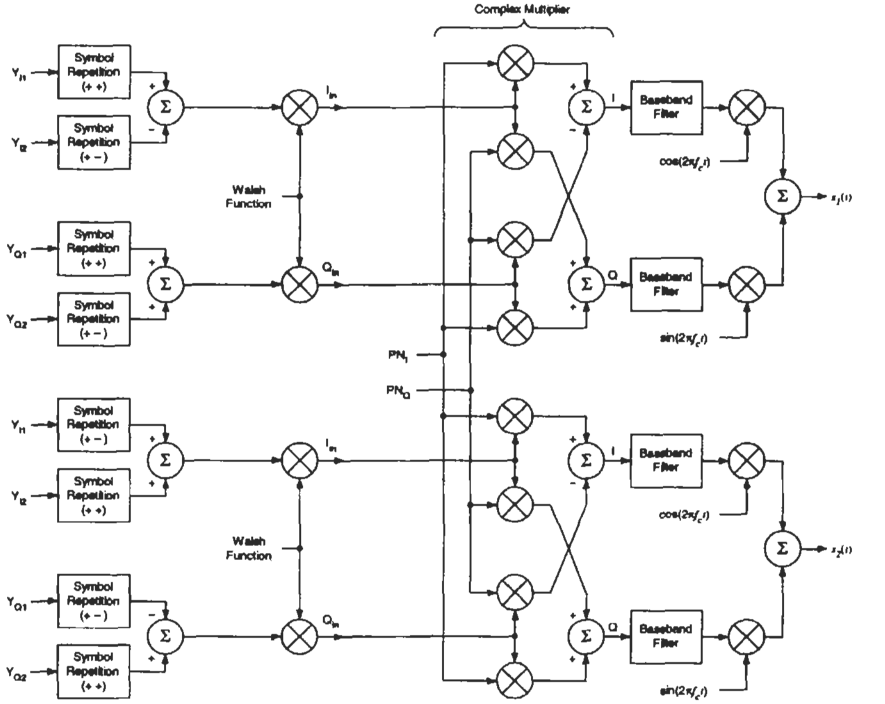


Figure 4-27: Space-Time Spreading

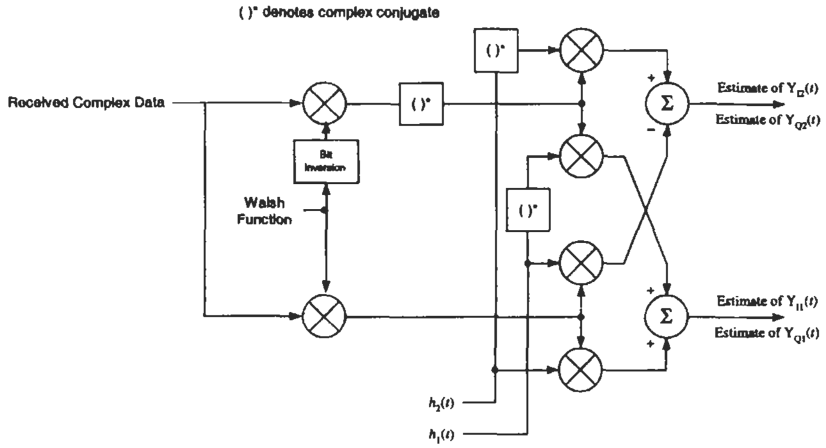


Figure 4-28: STS Receiver

The OTD spreader is given in Figure 4-29 and the receiver is given in Figure 4-30.

4.2 Reverse Link

The reverse link in cdma2000 was designed to introduce two important concepts that were not available in IS-95, namely coherent reverse link detection and fast forward link power control. As a result, a new physical layer was necessary for cdma2000.

The basic spreading structure is depicted in Figure 4-31. Note certain aspects of this spreading method which differ from the IS-95 reverse link:

- Inclusion of a pilot channel for coherent demodulation
- Use of Walsh code multiplexing for transmission of multiple data channels
- Long code spreading is slightly different to reduce peak-to-average ratio
- Use of hybrid quadrature spreading rather than offset-QPSK spreading

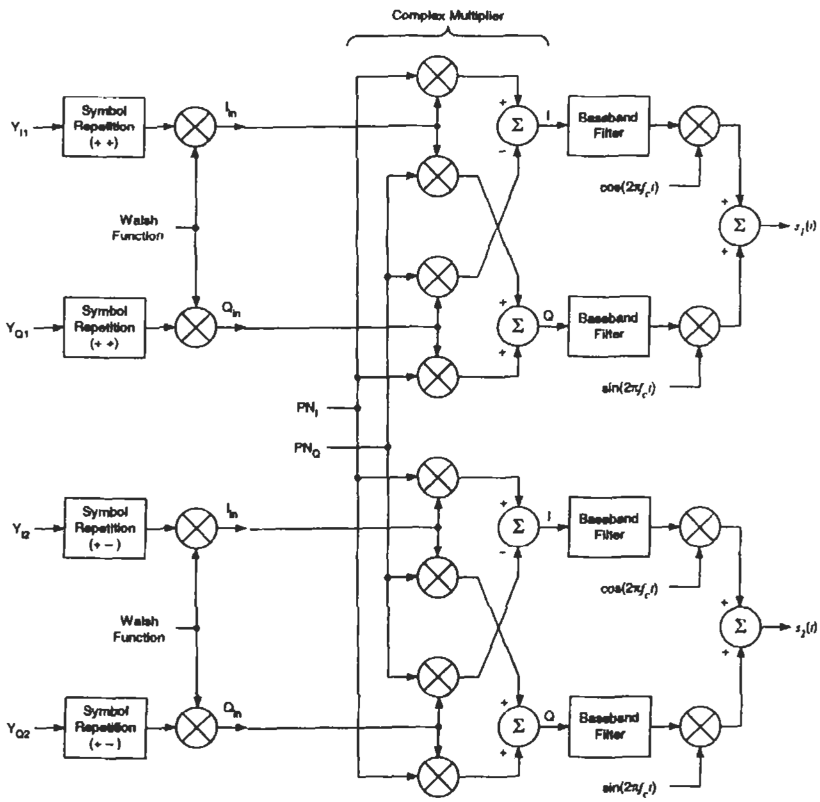


Figure 4-29: Orthogonal Transmit Diversity

( )* denotes complex conjugate

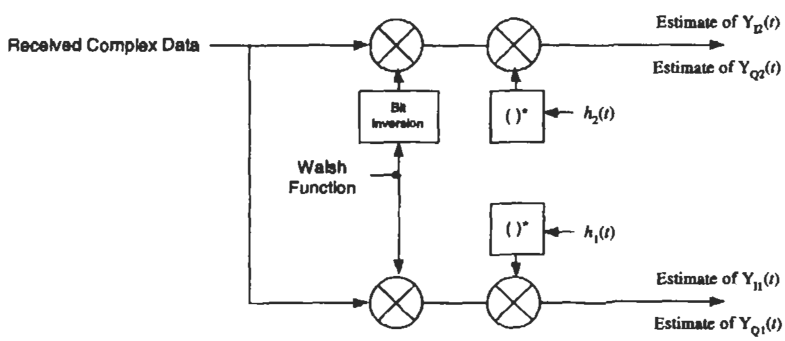


Figure 4-30: OTD Receiver

Several new code channels exist and will be discussed in the ensuing subsections. Each code channel will result in a modulation symbol indexed by A, B, or C. These indices are with respect to the complex spreading inputs in Figure 4-31.

4.2.1 Backwards-Compatible Common Channels

The main backwards-compatible feature that exists in the cdma2000 reverse link for common channels is the presence of a reverse access channel (R-ACH) identical to that used in IS-95.

4.2.2 New cdma2000 1X Common Channels

The reverse common control channel (RCCCH) and reverse enhanced access channel (REACH) were introduced to provide more reliable access by means of slotting different users before they transmit bursts on the reverse link and to provide a coherence form of reception for the base station access probe receiver. In addition, some burst data service that was not possible over the IS-95 access channel could be accommodated using these features.

For REACH operation, it is possible to transmit a burst in the form of an EACH header (see Figure 4-32). Otherwise, the transmission sequences for both the REACH and RCCCH are given in Figure 4-33. Note that both these channels are transmitted over the same Walsh code; their usage is different depending on the type of information being transmitted. The REACH is strictly for access. However, transmission on an RCCCH is possible after successful REACH transmission. In this case, the mobile is scheduled onto time slots (with respect to system time) for transmission of burst data. This type of operation is known as *reservation* mode. Moreover, operation on either the REACH or RCCCH can be *power controlled* by means of an assignment of the mobile to a power control slot on the *forward common power control channel* (FCPCH).

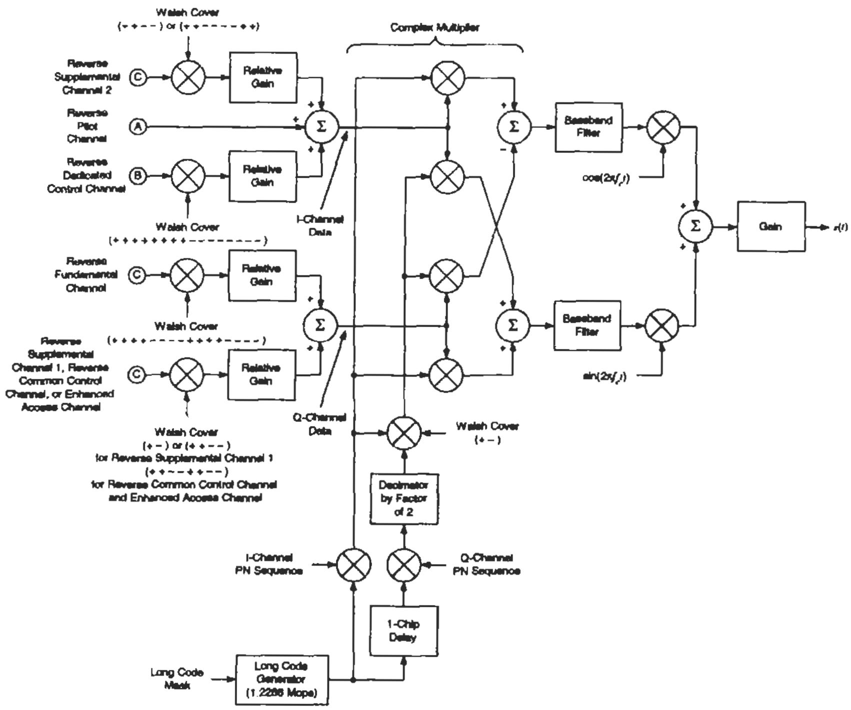


Figure 4-31: Reverse Link Spreading

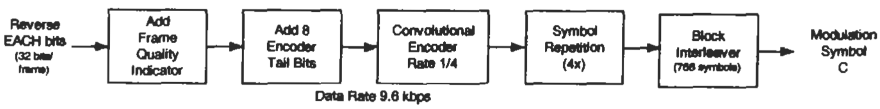
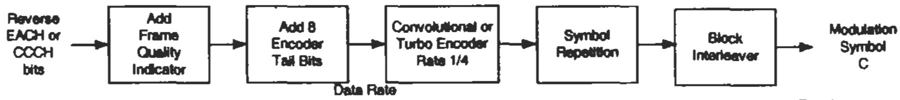


Figure 4-32: REACH Header



Bits/Frame	Bits	Data Rate (kbps)	Factor	Symbols	Rate(ksps)
172 Bits/5 ms	12	38.4	1x	768	153.6
360 Bits/10 ms	16	38.4	1x	1536	153.6
172 Bits/10 ms	12	19.2	2x	1536	153.6
744 Bits/20 ms	16	38.4	1x	3072	153.6
360 Bits/20 ms	16	19.2	2x	3072	153.6
172 Bits/20 ms	12	9.6	4x	1536	153.6

Figure 4-33: REACH and RCCCH

4.2.3 New cdma2000 1X Dedicated Channels

Several new dedicated control channels exist for the cdma2000 reverse link. Among them are the reverse pilot channel (RPICH), the reverse fundamental channel (RFCH), the reverse supplemental channel (RSCH), and the reverse dedicated common control channel (RDCCH). Except for the RPICH, the types of data associated with each channel are identical to their forward link counterparts.

The RPICH actually has power control commands for fast forward power control multiplexed into the constant stream of all 0's at an 800 Hz rate (see Figure 4-34).

Just as with the forward link dedicated channels, the other three reverse link dedicated channels are associated with radio configurations. In general RC3 corresponds to Rate Set 1–compatible data rates, while RC4 corresponds to Rate Set 2–compatible data rates. Higher RCs correspond to 3X spreading factors. The transmission sequences for the RDCCH, RFCH, and RSCH are given in Figure 4-35 through Figure 4-38.

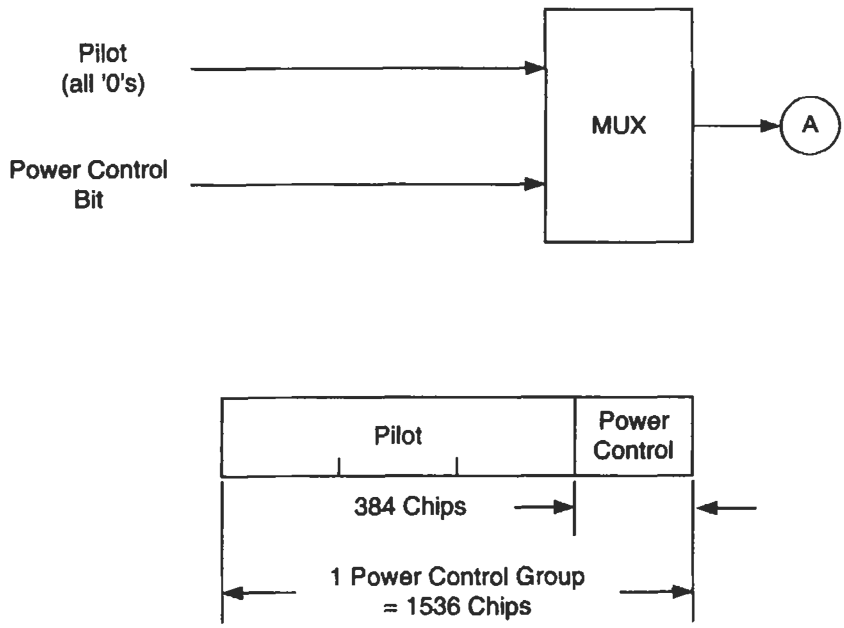


Figure 4-34: RPICH and Power Control Multiplexing

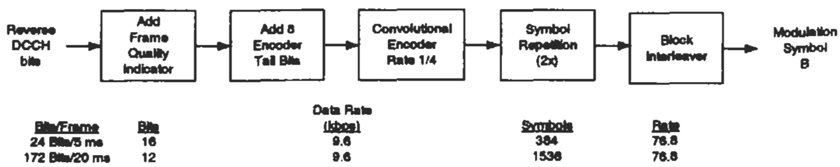


Figure 4-35: RDCCH for RC3

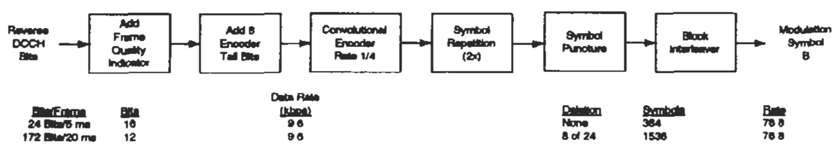
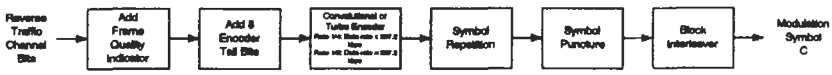


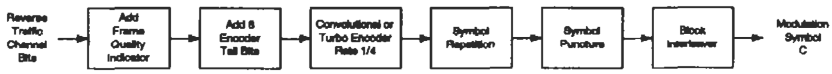
Figure 4-36: RDCCH for RC4



Bits/Frame	Bits	Data Rate (kbps)	Factor	Deletion	Symbols	Rate (ksps)
24 Bits/5 ms	16	9.6	1x	None	384	76.8
16 Bits/20 ms	6	1.5	16x	1 of 5	1536	76.8
40 Bits/20n ms	6	2.7/n	8x	1 of 9	1536	76.8/n
80 Bits/20n ms	8	4.8/n	4x	None	1536	76.8/n
172 Bits/20n ms	12	9.6/n	2x	None	1536	76.8/n
360 Bits/20n ms	16	19.2/n	1x	None	1536	76.8/n
744 Bits/20n ms	16	38.4/n	1x	None	3072	153.6/n
1512 Bits/20n ms	16	76.8/n	1x	None	6144	307.2/n
3048 Bits/20n ms	16	153.6/n	1x	None	12288	614.4/n
6120 Bits/20n ms	16	307.2/n	1x	None	12288	614.4/n

n=length of frame in multiples of 20 ms; n can equal 1, 2 or 4

Figure 4-37: RFCH and RSCH for RC3



Bits/Frame	Bits	Data Rate (kbps)	Factor	Deletion	Symbols	Rate (ksps)
24 Bits/5 ms	16	9.6	1x	None	384	76.8
21 Bits/20 ms	6	1.8	16x	8 of 24	1536	76.8
55 Bits/20n ms	8	3.6/n	8x	8 of 24	1536	76.8/n
125 Bits/20n ms	10	7.2/n	4x	8 of 24	1536	76.8/n
267 Bits/20n ms	12	14.4/n	2x	8 of 24	1536	76.8/n
552 Bits/20n ms	16	28.8/n	1x	4 of 12	1536	76.8/n
1128 Bits/20n ms	16	57.6/n	1x	4 of 12	3072	153.6/n
2280 Bits/20n ms	16	115.2/n	1x	4 of 12	6144	307.2/n
4584 Bits/20n ms	16	230.4/n	1x	4 of 12	12288	614.4/n

n=length of frame in multiples of 20 ms; n can equal 1, 2 or 4

Figure 4-38: RFCH and RSCH for RC4

4.3 Coding

The convolutional coding described in Section 2.6 is still applicable to cdma2000 (with a rate ¼ code added to the existing convolutional codes of IS-95). However, optional turbo coding (see Section 8.2, Chapter 2) is also included in cdma2000 for high-rate data services. The turbo code defined for cdma2000 is shown in Figure 4-39. With respect to the figure, by simply puncturing the output bits $\{X, Y_0, Y_1, X', Y'_0, Y'_1\}$, the rate of the turbo code can be modified to fit a variety of different information bit rates.

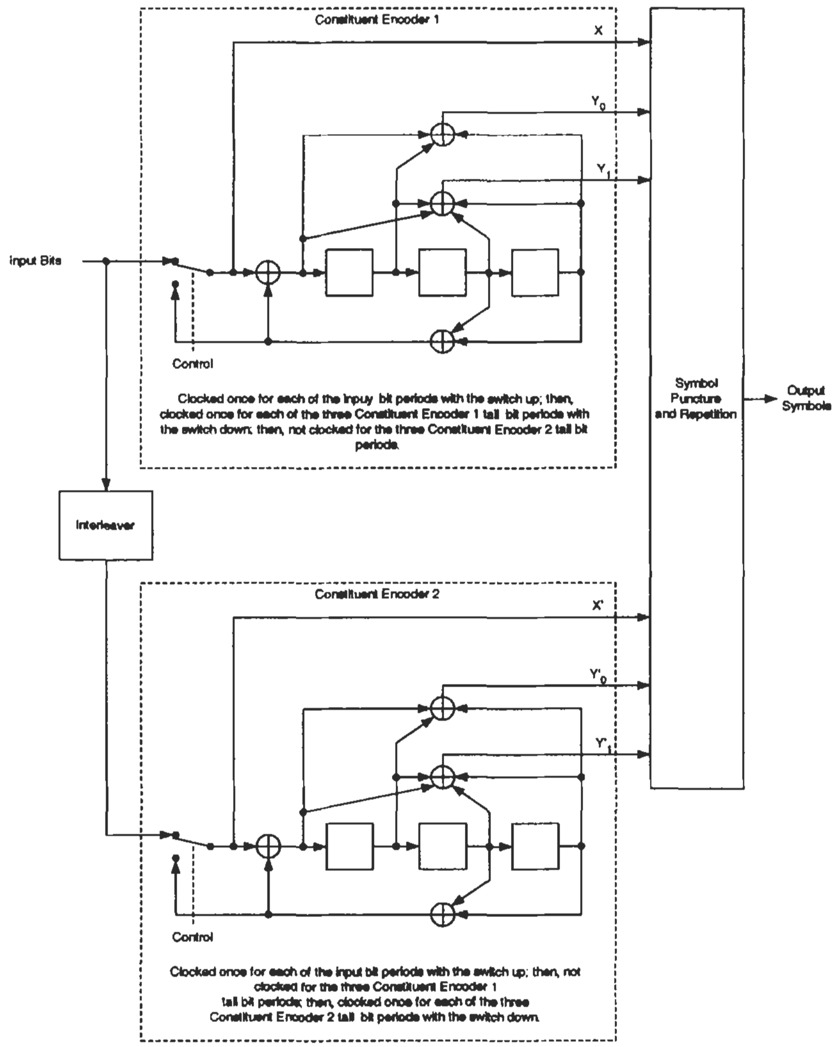


Figure 4-39: cdma2000 Turbo Code

4.4 Medium Access Control

cdma2000 introduced a MAC specification in an effort to ensure a correspondence between itself and the OSI layering model for data services. It receives its payload either from the link access control (LAC) layer or directly from one or more data services, provides the necessary functionality for multiplexing these different types of payload onto the physical layer, and also describes to an extent some aspects of the operation of the access modes for the cdma2000 reverse link. However, the new multiplex sublayer functionality introduced in the cdma2000 MAC will be focused on in this section, due to its criticality for cdma2000 operation.

4.4.1 Multiplex Sublayer

The two basic multiplex options introduced in IS-95-B were multiplex options 1 and 2, for fundamental channel (FCH) Rate Set 1 and 2 configurations, respectively. These two options are also referred to by their hexadecimal representations, 0x1 and 0x2. When supplemental code channels (SCCHs) are used, the new multiplex option (in decimal form) can be formed using the following equation:

$$m = n + 2r \tag{4-1}$$

where m is the multiplex option, n is the rate set (1 or 2), and r is the number of additional supplemental code channels (1 to 7). This formulation leads to a maximum multiplex option of 0x10. Each of these physical layer channels (FCHs or SCCHs) encompasses a single multiplex protocol data unit (i.e., MuxPDU), which is a basic information transmission unit for any type of payload.

These backwards-compatible multiplex options are still applicable to cdma2000 for radio configurations 1 and 2. However, for cdma2000's new radio configurations (radio configuration 3 and higher), these options are not directly applicable due to the following reasons:

a. The physical layer code channels in cdma2000's new radio configurations can carry a much larger payload than IS-95-B FCHs or SCCHs.
b. For multimedia services, it is desirable to have multiple MuxPDUs in a single physical layer code channel frame. This provides for transmission of different types of data simultaneously.

As a result, new multiplex options with relaxed restrictions on the number of MuxPDUs per frame were introduced in cdma2000. The types of MuxPDUs were further specified:

a. MuxPDU Type 1 — backwards-compatible Rate Set 1 MuxPDUs, i.e., FCHs and SCCHs. These MuxPDUs may also be used on supplemental channels (SCHs) and 20-ms–framed dedicated control channels (DCCHs) which are of the lowest rate (9600 or 14400 bps).
b. MuxPDU Type 2 — similar to MuxPDU Type 1, except applicable to Rate Set 2.
c. MuxPDU Type 3 — this type of MuxPDU may only be transmitted over high-rate SCHs. There are two allowable block sizes — single and double. The single block size is 170 bits for Rate Set 1 and 266 bits for Rate Set 2. Similarly, the double block size is 346 and 538 bits for Rate Sets 1 and 2 respectively.
d. MuxPDU Type 4 — carries 5-ms block of data (24 bits total of payload). It can be used on an FCH or DCCH.
e. MuxPDU Type 5 — this is a variable-length PDU that can only be used on an SCH. Its size is specified in the PDU header, and provides flexibility to the higher layers in forming the PDUs to be placed into the physical layer.
f. MuxPDU Type 6 — this is a variable-length PDU that can only be used on the FCH or DCCH, and is intended to provide total flexibility in forming the effective data rate on these channels.

Making use of these MuxPDU types, all multiplex options higher than 0x10 can be formed simply by a 16-bit binary number using the guidelines of Table 4-2.

It is possible to encapsulate MuxPDUs on the SCH into *logical transmission units* (LTUs), and to send multiple LTUs in a single 20-ms frame. An LTU can contain one or more MuxPDUs; however, a 16-bit cyclic redundancy check (CRC) must be appended at the end of each LTU to assure error detection within the LTU. The use of LTUs allows the receiver to recover parts of a frame when a frame in error has been detected.

Bit indices (0=most significant, 15=least significant)	Description	Options
0-3	Format	'0000'
4-5	MuxPDU Type	'00' — MuxPDU Types 1, 2, or 4 '10' — MuxPDU Type 3 '01' — MuxPDU Type 6 '11' — MuxPDU Type 5
5-6	Block Size	'00' — Single size '01' — Double size '11' — Variable size
7-13	Maximum No. Data Block per Frame	'000000' — No restriction '000001' — '001000'
14-15	Rate Set	'00' — Not Applicable '01' — Rate Set 1 '10' — Rate Set 2

Table 4-2: Multiplex Option Numbering

4.4.2 Interface to LAC

The LAC sublayer encapsulates layer 3 signaling into LAC service data units (LAC SDUs) with their own CRCs. Each LAC SDU is then segmented into MAC SDUs to be transmitted over the physical layer. The MAC SDUs are mapped to the physical layer through MuxPDUs to be routed on specific code channels.

The receiving function of the LAC sublayer entails reassembling the LAC SDU based on the SDUs received from the MAC. If a LAC SDU is not reassembled without detecting errors, the LAC may recover the SDU through an automatic repeat request (ARQ) protocol that requests a retransmission of part or all of the LAC SDU.

4.5 Data Services and Radio Link Protocol

The conventional internet protocol (IP) stack, which is used for wireline data communications, tends to suffer from lack of robustness when used over a wireless cellular environment. This is due to the fact that IP protocols generally rely on one layer of reliability, the *transmission control protocol* (TCP), which serves to recover lost octets based on ARQ methods. TCP is optimized for wireline networks; cellular networks, which generally suffer higher error rates and are subject to tighter bandwidth restrictions, cannot provide the basic performance needed for even TCP to be sufficient for error recovery. Moreover, some real-time applications that may be run at higher error rates but cannot sustain the delay of TCP octet recovery may instead use the *universal datagram protocol* (UDP). UDP has no means of recovering packets lost over the wireless network.

As a result, wireless systems use *radio link protocols* (RLPs) to provide an additional layer of reliability on top of TCP. The basic IS-95/cdma2000 RLP-based protocol stack is depicted in Figure 4-40. Recall that the cdma2000 MAC layer can directly accept PDUs from either the LAC or a data service. The lowest layer of an IS-95–based data service is the RLP layer.

RLP does not accept IP packets directly in IS-95/cdma2000. Instead, IP packets are passed through a *point-to-point protocol* (PPP) instance. PPP serves primarily to provide some security to the IP datagrams. RLP therefore must be able to segment PPP packets for transmission, and reassemble received RLP PDUs into whole packets for delivery to the PPP layer.

RLP provides reliability by means of ARQ, more specifically by *selective repeat* ARQ. This means that RLP PDUs are retransmitted only when the receiver specifically requests them. This type of ARQ is also sometimes referred to as a *NAK*-based protocol. NAKs (i.e., Negative AcKnowledgments) are sent by the receiver to the transmitter to request retransmissions of specific PDUs. Thus, the RLP receiver must keep track of correctly received PDUs through the use of embedded sequence numbers. It maintains two counters V(N) and V(R); the former keeps track of the next PDU to be received, and the latter keeps track of the next PDU expected to be received. When these two counters point to the same sequence number, then this signifies that there are no outstanding RLP PDUs to be delivered. However, when these two are not equal, then this means that an RLP PDU was not received correctly and should be NAK'ed. If the receiver sends a NAK for a missing RLP PDU, it must initiate a timer for that RLP PDU (the *retransmission timer*). This timer, which is only incremented upon the receipt of valid RLP PDUs, cannot exceed a certain pre-established number (based on the expected *round trip time* RTT for a retransmission). This timer is terminated once the retransmission is received correctly. Otherwise, once this timer has expired, the receiver may send another NAK and start another timer known as the *abort timer* (also based on the RTT). If the abort timer expires, then the RLP receiver "gives up" and passes whatever correct octets it has received in the PPP packet up to the PPP layer. In this manner, RLP can provide some limited error recovery for data packets in the wireless channel, but it does not provide excessive reliability so that upper layers may erroneously determine that the underlying connection has been lost.

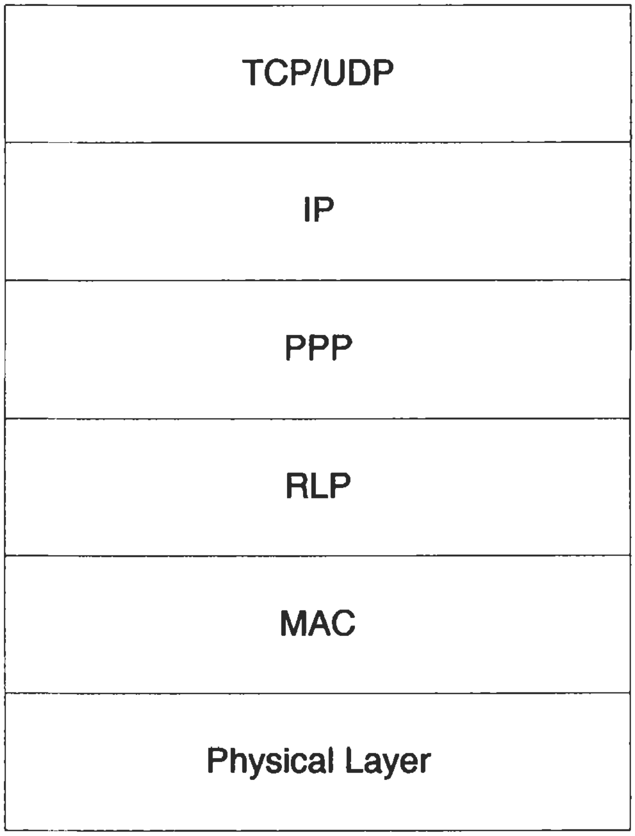


Figure 4-40: RLP-Based Protocol Stack

When the receiver has detected a missing RLP PDU, or a timer has expired so that a NAK must be re-sent, RLP may send more than one NAK as a result. The maximum number of timer events for the recovery of an RLP PDU is also sometimes referred to as the number of NAK rounds allowed. There are a maximum number of NAKs per round as well.

For instance, in IS-95-B, on detection of a missing RLP PDU, *two* NAKs are sent for the missing PDU and the retransmission timer is initialized for the PDU. Upon expiration of the retransmission timer, *three* NAKs are sent for the missing PDU and the abort timer is initialized. This is sometimes referred to as a 1,2,3 method of retransmission. Although 5 retransmissions might seem excessive, this feature is useful to provide resiliency in the presence of lost NAKs or lost retransmissions, and thus minimize overall packet delay.

cdma2000 has provided a means of negotiating the levels of reliability at the RLP layer during the initialization of the data service. More precisely, the NAKs per round and maximum number of rounds may be negotiated at the beginning of the call [12].

5. CONCLUSIONS

The IS-95 and cdma2000 systems were developed with the primary intention of increasing the voice capacity offered in cellular systems. As a result, these systems are also primarily optimized for circuit-switched data services. However, these systems have been extended (in the form of 1X-EV, as described in the next chapter) to make them more suitable for multiple-access packet-switched services.

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Chapter 5

1X-EV: Evolution of cdma2000

Key Topics: 1X-EV, 1X-EV-DV, 1X-EV-DO, best-effort data services

Overview: A description of the basic air interface of 1X-EV systems is given. The 1X EV standards development takes the form of two subsystems, 1X-EV-DO and 1X-EV-DV. The 1X-EV-DO system, based on Qualcomm's High Data Rate (HDR) proposal, is a system that enhances data services to cdma2000 users over a dedicated 1.25 MHz carrier. The 1X-EV-DV system is also described, which remains backwards compatible to cdma2000 yet provides enhanced data services over a 1.25 MHz carrier as well. Both of these systems achieve enhanced data rates through the use of bandwidth-efficient modulation and shared-channel services.

1. INTRODUCTION

Although cdma2000, particularly its 1X flavor, was able to provide enhancements in voice capacity over IS-95 through the use of a coherent reverse link and fast forward power control, many IS-95 (and cdma2000) operators looked forward to anticipated demand for wireless data services. Although wireless data services were deployed successfully in certain parts of the world (e.g., I-Mode in Japan by NTT DoCoMo) by 2001, enhanced wireless data services in North America were providing increasing revenues, but not sufficiently large to compromise voice revenues. As a result, an effort to enhance cdma2000 1X for wireless data services while retaining compatibility with cdma2000 voice services was initiated in early 2000.

In March 2000, Qualcomm Incorporated presented a proposal on their proprietary High Data Rate (HDR) system to the 3GPP2. This system provided enhanced data services over a 1X carrier, and was meant to provide compatibility with cdma2000 voice services in a deployment in a multicarrier system (i.e., HDR on one 1X carrier and cdma2000 on another 1X carrier). Also in March 2000, Nokia, Motorola and Texas Instruments presented to the 3GPP2 a proposal known as 1XTREME (1X Third-generation Enhanced Modulation and Encoding) that provided enhanced data services over a 1X carrier while retaining backwards compatibility to cdma2000 voice services over the same carrier.

The CDMA Developers Group (CDG), an industrial consortium of IS-95 vendors and operators, put forward a proposal for cdma2000 evolution based on the developments in the 3GPP2. In June 2000, the CDG submitted to the 3GPP2 a proposal for cdma2000 that would take two stages:

- 1X-EV-DO (1X EVolution for Data Only), based on the HDR proposal
- 1X-EV-DV (1X EVolution for Data and Voice), which would enhance data services while retaining backwards compatibility to cdma2000 on a 1X carrier

2. 1X-EV-DO

The 1X-EV-DO standard ([1],[2]) proposed a hybrid TDMA/CDMA approach to enhancement of data-only services over a dedicated 1.25 MHz carrier on the forward link. The reverse link retains many of the elements of the cdma2000 reverse link, with additional features to provide enhancements for data services on the forward link as well. One important aspect of 1X-EV-DO is that in order to support voice services, a dedicated cdma2000 carrier must be present. Therefore two 1.25 MHz carriers must be available to support voice and data operation in a 1X-EV-DO system.

One of the main features in 1X-EV systems is the use of *link adaptation*. This is a feature wherein different users are assigned different physical layer attributes depending on the channel conditions each user sees. For instance, a user near base station may be assigned a higher data rate than a user further away from the base station. The data rates on the forward link vary through the use of *adaptive modulation and coding* (AMC). AMC involves the use of different bandwidth-efficient modulation methods (e.g., BPSK, QPSK, QAM) along with changing the forward error correcting (FEC) code rate. Based on the user's forward link channel quality, there exists a modulation and coding scheme (MCS) that maximizes throughput to the user.

However, the base station does not know what the forward link channel quality is to an individual user without feedback information from the user. Based on fading conditions, the user can select lower data throughput MCSs when channel conditions are poor and higher throughput MCSs when channel quality conditions improve. However, there is a loop delay between the time it determines the best MCS and relays its request back to the base station and the time the base station actually implements the MCS request.

In addition, because data services (unlike voice) are bursty, 1X-EV-DO time-multiplexes users over a shared forward link channel. This utilizes radio resources effectively, as the forward link channels are not wasted on users who are not receiving data at the moment. As a result, some extra overhead is needed to schedule users on the forward link. Moreover, it is very difficult to coordinate base stations on the forward link for a true soft handoff scenario. This is because different base stations may have different users, and the times over which they schedule an individual user may differ. As a result, in 1X-EV-DO systems, the mobile may only receive data from one base station at a time.

The reverse link has to include new control information to provide data requests for the forward link, signal to the base station a new rate to be used on the reverse link, and provide indications to the network of which base station is the best base station to serve the user (since only one base station can deliver traffic to the user).

In addition, a fast ARQ method is implemented in 1X-EV-DO. ARQ (Automatic Repeat reQuest) involves the redelivery of a frame of information when the previous attempt at transmission has failed. The transmitter knows that the attempt has failed based on the failure to receiver an ACK (acknowledgement) indication from the receiver.

2.1 Forward Link

The forward link is one shared pipe over which all users may receive information. Forward link channels are all time-multiplexed, meaning that forward link power control is not applicable—the entire base station transmitter power is allocated to the channel being transmitted at any particular moment in time. The forward link spreading and channelization are depicted in Figure 5-1.

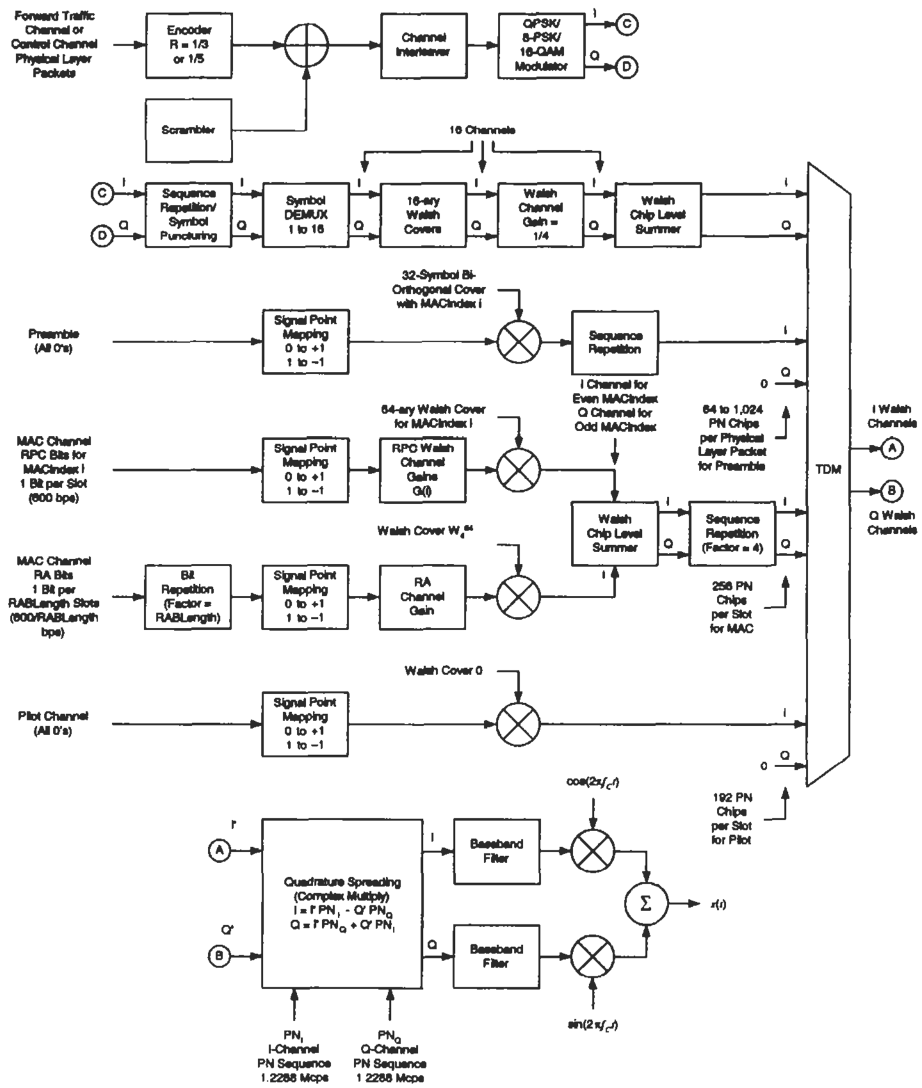


Figure 5- 1: 1X-EV-DO Forward Link Spreading

Note the presence of different channels all being time-multiplexed over the same modulation sequence. The time multiplexing is based on units of time called *slots*, where a slot is 1.67 ms (slots occur at a 600 Hz rate). The time-multiplexing sequence is given in Figure 5-2.

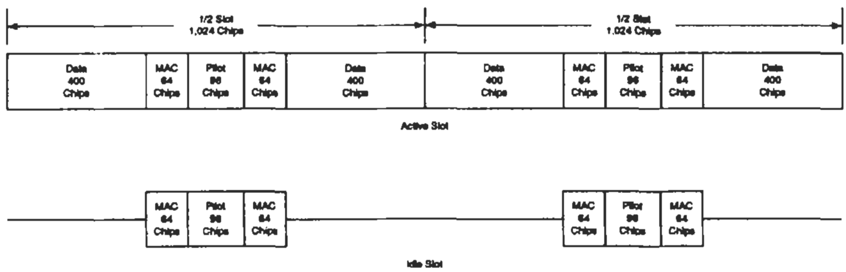


Figure 5-2: 1X-EV-DO Forward Link Time-Multiplexing Structure

The data, MAC (medium access control), and pilot channels are all multiplexed together to provide the necessary information to all users in the cell.

The 1X-EV-DO forward link still takes advantage of the short codes of IS-95/cdma2000 and uses PN offsets to distinguish between different base stations. In addition, the baseband filtering is identical to IS-95/cdma2000.

2.1.1 Pilot Channel

The pilot channel serves the same function as the IS-95/cdma2000 pilot channel, except it is time-multiplexed and occurs twice per slot. The total number of chips the pilot channel is spread over is 182, based on 96 chips per half-slot.

2.1.2 MAC Channel

The MAC channel in 1X-EV-DO provides for scheduling different users onto the forward link and also providing information to the different users about reverse link capabilities of the system. Length-64 Walsh codes are used to transmit different types of information to different users. Given Walsh codes W_{64}^{i}, where i is the Walsh code index ($0 \leq i < 64$), the even-indexed Walsh codes are transmitted over the in-phase input to the spreader while the odd indices are transmitted over the quadrature spreader input.

In addition to global control information, the MAC channel also indicated Reverse Activity (RA) to all users and Reverse Power Control (RPC) commands to individual users. RPC information must constantly be transmitted to users who are actively transmitting data. RA bit (RAB) is more of a broadcast indication to users that their data rates should be reduced (due to unacceptably high load levels on the reverse link) or may be increased.

The MAC channel is transmitted 4 times per slot in bursts of 64 chips each. The MAC information transmitted over each Walsh code in a burst is given in Table 5-1.

Walsh Code index i	MAC Channel Use
0 and 1	Not Used
2	Not Used
3	Not Used
4	RA Channel
5—63	Available for RPC Channel Transmissions

Table 5-1: MAC Channel Functionality for each Walsh Code

2.1.3 Traffic Channel Preamble

The preamble is important in that it indicates to the mobile what sort of MCS it will be receiving for the ensuing traffic bits. The preamble still uses *biorthogonal* modulation. A biorthogonal signal set with M signals can be constructed from an orthogonal signal set of $M/2$ signals by simply concatenating the negation of those $M/2$ signals. In the case of the 1X-EV-DO preamble, M = 32, i.e., the Walsh codes of length 32 are used for biorthogonal modulation. Therefore, for each code index i, where $0 \leq i < 64$, the Walsh codes are assigned as

$$
\begin{aligned}
&W_{i/2}{}^{32},\ i = 0,2,\mathrm{K}\ ,62 \\
&\overline{W_{(i-1)/2}{}^{32}},i = 1,3,\mathrm{K}\ ,63
\end{aligned}
\tag{5-1}
$$

where $\overline{W}$ is the bit-by-bit negation of W.

Biorthogonal sequences differ from orthogonal sequences in that any two distinct sequences from the set can have a cross-correlation of 0 or -1 (assuming each entry is mapped to $+/-$ 1 and the final result is normalized to 1). Two distinct orthogonal sequences drawn from the same set can only have a cross-correlation of 0.

Each of the 32-chip codes may be assigned to an individual user or to a control channel for broadcast use. In Table 5-2, the code assignments for different preambles are shown.

Walsh Code index i	Preamble Use
0 and 1	Not Used
2	76.8-kbps Control Channel
3	38.4-kbps Control Channel
4	Not Used
5—63	Available for Forward Traffic Channels

Table 5-2: Preamble Code Assignments

Based on a predefined repetition sequence for the Walsh code assigned to an individual user, that user may discern the traffic channel data rate assignment. The repetition pattern for the preamble code is given in Table 5-3. Each of the data rates given in Table 5-3 corresponds to a particular MCS for the traffic channel.

2.1.4 Traffic Channel

The MCSs along with their associated number of slots and total time-multiplexed chips are given in Table 5-4.

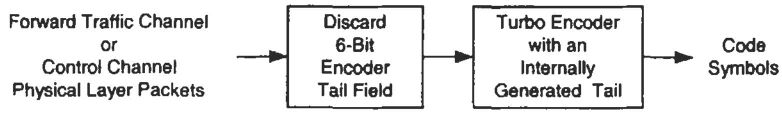
The data rates are uniquely associated with a modulation and code rate, and each frame occupies a fixed number of slots. Each block of traffic channel or control channel bits is passed into a turbo encoder to create channel symbols, as in Figure 5-3. This turbo encoding process is also depicted in Figure 5-1.

After turbo encoding, code symbol scrambling is performed (refer to Figure 5-1). Unlike the IS-95/cdma2000 forward link scrambling, which uses the long code generator, 1X-EV-DO uses a code generator based on the polynomial $h(D) = D^{17} + D^{14} + 1$. The initial state of this 17-tap register is based on the 6-bit MAC Channel code index i (represented as $r_5r_4r_3r_2r_1r_0$) and 4 bits denoting the data rate used for the packet being transmitted (represented as $d_3d_2d_1d_0$). The other 7 bits are set to 1 (i.e., the initial state is $1111111r_5r_4r_3r_2r_1r_0d_3d_2d_1d_0$). This scrambling serves to help reduce the peak-to-average ratio of the symbols being transmitted after the ensuing variable modulation step.

Data Rate (kbps)	Values per Physical Layer Packet		
	Slots	32-Chip Preamble Sequence Repetitions	Preamble Chips
38.4	16	32	1,024
76.8	8	16	512
153.6	4	8	256
307.2	2	4	128
614.4	1	2	64
307.2	4	4	128
614.4	2	2	64
1,228.8	1	2	64
921.6	2	2	64
1,843.2	1	2	64
1,228.8	2	2	64
2,457.6	1	2	64

Table 5-3: Preamble Repetition Pattern

After block interleaving, the variable modulator uses one of the modulation methods (QPSK, 8-PSK, 16-QAM), depending on the data rate chosen for transmission, and yields in-phase (I) and quadrature (Q) output signals. These signal symbols are repeated or punctured depending on the necessary number of symbols to be transmitted (as a function of the data rate) and then the symbols are demultiplexed over 16 Walsh codes each of length 16 chips. Each of these Walsh codes is transmitted with equal power (i.e., 1/16 of the total power).



Data Rate (kbps)	Slots Used	Total Code Rate	Bits per Packet	Bits per Packet	Symbols per Packet
38.4	16	1/5	1,024	1,018	5,120
76.8	8	1/5	1,024	1,018	5,120
153.6	4	1/5	1,024	1,018	5,120
307.2	2	1/5	1,024	1,018	5,120
614.4	1	1/3	1,024	1,018	3,072
307.2	4	1/3	2,048	2,042	6,144
614.4	2	1/3	2,048	2,042	6,144
1,228.8	1	1/3	2,048	2,042	6,144
921.6	2	1/3	3,072	3,066	9,216
1,843.2	1	1/3	3,072	3,066	9,216
1,228.8	2	1/3	4,096	4,090	12,288
2,457.6	1	1/3	4,096	4,090	12,288

Figure 5-3: Traffic Channel Turbo Encoding

2.2 Reverse Link

The 1X-EV-DO reverse link, in addition to providing sufficient feedback for control of the forward link (particularly link adaptation), also provides different reverse traffic data rates. The basic reverse link channelization structure is shown in Figure 5-4, with the reverse link spreading depicted in Figure 5-5.

The 1X-EV-DO reverse link has one common channel, the access channel. The dedicated channel is the reverse traffic channel, which collectively includes the pilot, reverse rate indication (RRI), data rate control (DRC), acknowledgement (ACK), and data channels.

The spreading in Figure 5-5 still uses the short codes of cdma2000/IS-95. The long code of cdma2000/IS-95 is also used, but the user uses two different masks on the long code generator. One of the masks is the in-phase mask, which generates U_I, and the other is the quadrature mask, which generates U_Q.

Data Rate (kbps)	Number of Values per Physical Layer Packet				
	Slots	Bits	Code Rate	Modulation Type	Total Chips (Preamble, Pilot, MAC, Data)
38.4	16	1,024	1/5	QPSK	1024, 3072, 4096, 24576
76.8	8	1,024	1/5	QPSK	512, 1536, 2048, 12288
153.6	4	1,024	1/5	QPSK	256, 768, 1024, 6144
307.2	2	1,024	1/5	QPSK	128, 384, 512, 3072
614.4	1	1,024	1/3	QPSK	64, 192, 256, 1536
307.2	4	2,048	1/3	QPSK	128, 768, 1024, 6272
614.4	2	2,048	1/3	QPSK	64, 384, 512, 3136
1,228.8	1	2,048	1/3	QPSK	64, 192, 256, 1536
921.6	2	3,072	1/3	8-PSK	64, 384, 512, 3136
1,843.2	1	3,072	1/3	8-PSK	64, 192, 256, 1536
1,228.8	2	4,096	1/3	16-QAM	64, 384, 512, 3136
2,457.6	1	4,096	1/3	16-QAM	64, 192, 256, 1536

Table 5-4: MCS and Slot Assignments

The basic frame is 16 slots (each slot is 1.33 ms), which means 26 2/3-ms framing, unlike the 20 ms framing used in cdma2000/IS-95. Unlike the 1X-EV-DO forward link, the reverse link does not use link adaptation through modification of modulation. The basic modulation is BPSK over quadrature spreading.

2.2.1 Access Channel

The 1X-EV-DO access channel makes use of the pilot and data channels depicted in Figure 5-4. An access probe consists of a preamble where only the pilot channel is transmitted for two frames. This is followed by one or more access channel packets, which are transmitted over the data channel. An example of the timing of this transmission sequence is given in Figure 5-6.

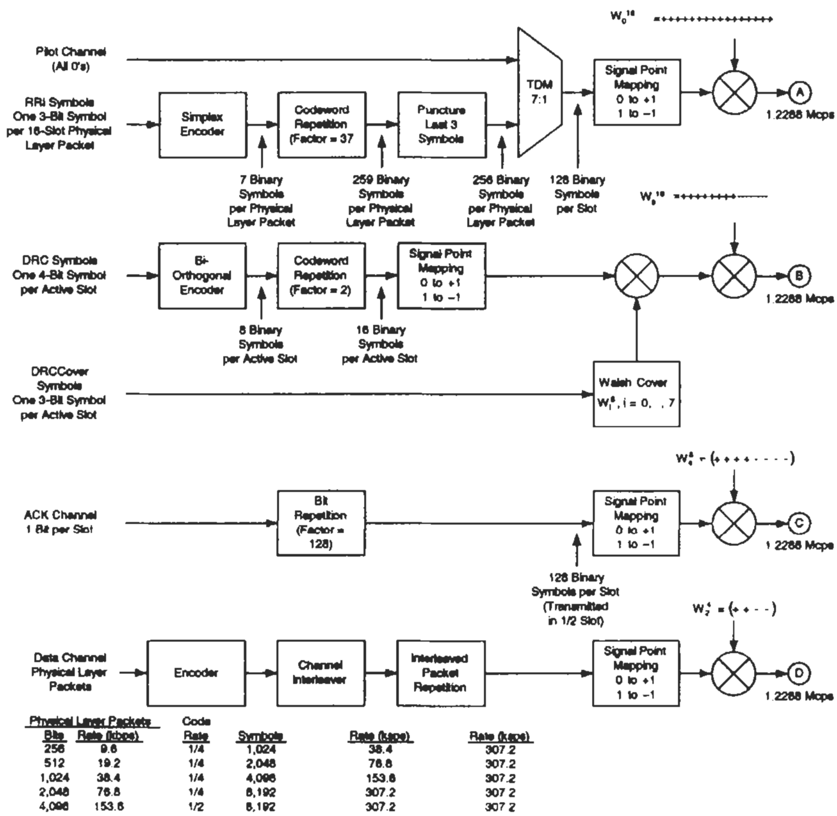


Figure 5-4: Reverse Link Channels

2.2.2 Data Rate Control (DRC) Channel

The DRC channel is important to make the basic enhancements of 1X-EV-DO work. In particular, link adaptation is dependent on properly transmitted DRC symbols, as these tell the network which data rates may be used on the forward link based on the channel conditions the mobile is seeing. The DRC symbol is a 4-bit number corresponding to one of the possible data rates on the forward link. This is mapped to an 8-bit biorthogonal Walsh code (see Section 2.1.3 for a definition of biorthogonality).

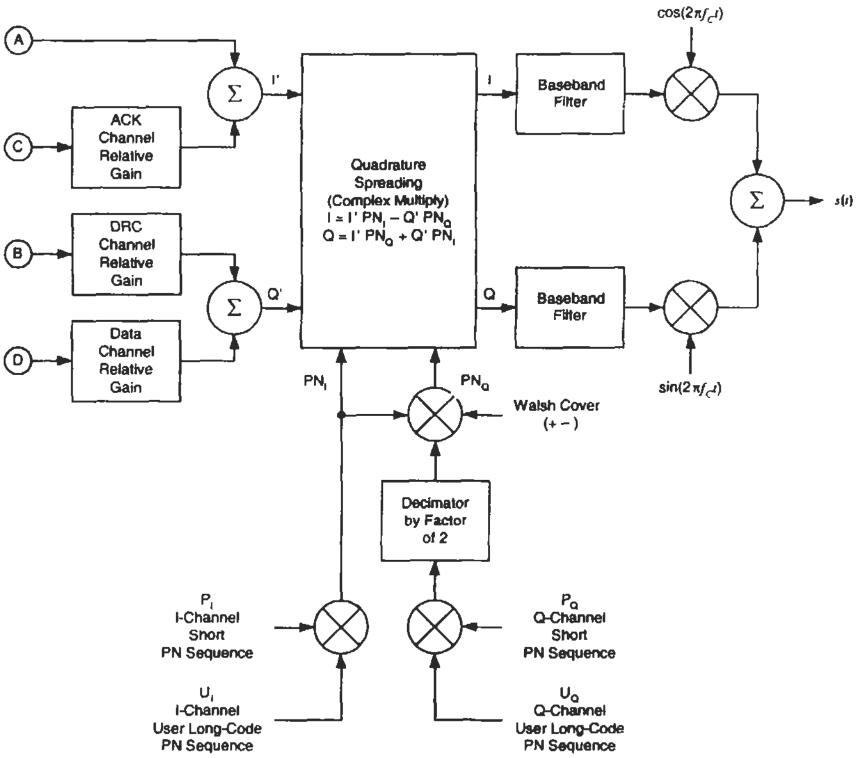


Figure 5-5: Reverse Link Spreading

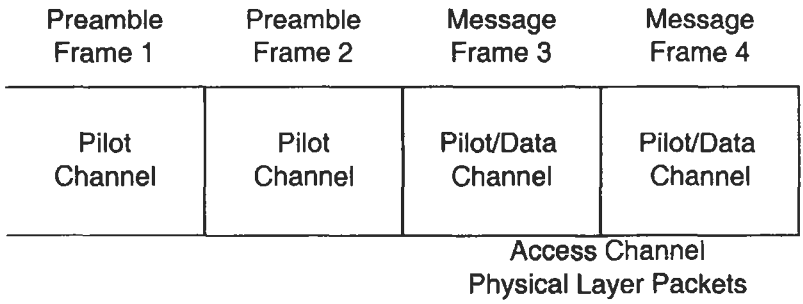


Figure 5-6: Access Probe

A 3-bit *DRC cover* is used to index a sector in the active set. This sector is the sector the mobile is selecting as the best serving sector. This cover selects one of eight 8-bit Walsh codes, which is used to further modulate the biorthogonal code generated from the DRC symbol. Recall that soft handoff is not easy to implement in its conventional sense when multiple users share the forward link resource. Therefore, the mobile must "point" its DRC request to a base station (or sector) which would best serve it by providing the highest data rate.

The DRC is sent at a 600 Hz rate, to match the slotting time of the forward link. As the coherence time of the channel decreases (i.e., mobile velocity increases), it is increasingly difficult for the DRC to track the changing forward link radio environment. Therefore, sometimes mobile channel predictors may be used in the mobile station to improve link adaptation performance.

2.2.3 Reverse Rate Indication (RRI) Channel

The RRI is a 3-bit symbol set once per frame to indicate to the base station receiver at what data rate the reverse traffic is being sent. Each RRI symbol is translated to a predefined codeword, which is repeated and time-multiplexed into the pilot channel. The RRI mappings are given in Table 5-5.

Data Rate (kbps)	RRI Symbol	Predefined Codeword
0	000	0000000
9.6	001	1010101
19.2	010	0110011
38.4	011	1100110
76.8	100	0001111
153.6	101	1011010
Reserved	110	0111100
Reserved	111	1101001

Table 5-5: RRI Mappings

This approach to reverse link data rate control is a significant departure from IS-95/cdma2000, in which data rates on the reverse link were strictly controlled and assigned by the network. In 1X-EV-DO, data rates are adapted by the mobile station based on implementation-dependent criteria.

2.2.4 Data Channel

The data channel involves the transmission of reverse traffic over a specified Walsh code channel on the reverse link. It is used for access probes, but can also be used during traffic operation. The basic data rates, code rates, and frame sizes for the data channel are given in Table 5-6 and in Figure 5-4.

Data Rate (kbps)	9.6	19.2	38.4	76.8	153.6
Reverse Rate Index	1	2	3	4	5
Code Rate	1/4	1/4	1/4	1/4	1/2
Bits per Physical Layer Packet	256	512	1,024	2,048	4,096
Turbo Encoder Code Rate	1/4	1/4	1/4	1/4	1/2

Table 5-6: Data Channel Parameters

2.2.5 Acknowledgement (ACK) Channel

The ACK channel is used for indicating to the network whether the most recently received forward link traffic channel packet was received correctly. If the ACK bit is set to '1', this is treated as a NAK (negative acknowledgement), meaning that the forward link packet was received incorrectly.

The ACK channel is only sent when a forward link packet has been received. Otherwise it is gated off. If the forward link packet was sent during slot n, the ACK bit is sent during slot $n + 3$. The ACK bit is transmitted over 1024 chips, i.e., a half-slot. The network may then send a retransmission or partial retransmission of the packet in response to a NAK reception.

3. 1X-EV-DV

The 1X-EV-DV standardization effort was initiated in the 3GPP2 in the fall of 2000, and several proposals were submitted. The most complete proposal for 1X-EV-DV presented during this time was the 1XTREME proposal ([3]–[6]), cosponsored by Nokia, Motorola, and Texas Instruments. The 1XTREME proposal put forward a system whose voice modes were fully compatible with cdma2000 1X voice modes, yet also introduced shared channel services (like 1X-EV-DO) for enhanced data services. Moreover, a medium access control layer was introduced to support enhanced data services for the forward link.

Moreover, the 1XTREME proposal reuses all common channels from cdma2000 1X, i.e., forward paging channel (FPCH), forward sync channel (FSYNC), reverse access channel (RACH), etc. This is due to the fact that 1XTREME simply introduces a new dedicated traffic mode of operation to enhance forward link data rates.

The fundamental concept behind 1XTREME is the approach of flexible resource allocation to users based on the offered load. In other words, 1XTREME was designed to provide the freedom to allocate radio resources in a prioritized fashion: available resources may be prioritized to voice or circuit-switched data users, and the remaining resources may be allocated to best-effort data users. However, the best-effort data users still may receive enhanced data rates through the use of bandwidth-efficient modulation and fast ARQ mechanisms. Moreover, the voice and circuit-switched users in a

1XTREME system essentially employ a cdma2000 air interface, thereby leveraging the cdma2000 standard.

The means of flexible resource allocation can be seen in the use of *code trees*. Recall that cdma2000 and IS-95 employ Walsh codes to achieve forward link channelization, and that these Walsh codes may be recursively generated (i.e., longer length Walsh codes are generated from shorter length Walsh codes). Thus any Walsh code $W_m^{\ n}$ of length n can be used to generate two orthogonal Walsh codes $W_m^{\ n}W_m^{\ n}$ and $W_m^{\ n}\overline{W_m}^{\ n}$, both of length $2n$ ($\overline{W_m}^{\ n}$ being a bit-wise inversion of $W_m^{\ n}$).

Therefore, the Walsh codes used by the network for forward link transmission may be generated as a code tree whose branches yield Walsh codes of larger and larger length. If a group of these codes of fixed length are allocated for best-effort data users, the effective data rates to these data users can be increased. However, as more voice or circuit-switched users are allocated codes, the group of codes shared among best-effort data users will shrink.

In 1XTREME, this shared group of codes is the set of Walsh codes of length 16. Clearly, not all of them can be used, as certain common channels (e.g., forward pilot [F-PICH] or forward paging [F-PCH]) have predefined Walsh codes which are necessary for mobile stations to acquire and initiate access to the network. However, each of these codes may be taken away from this shared group of Walsh codes to generate four Walsh codes of length 64 (i.e., spreading factor 64); thus four voice users may be serviced when a single code from this shared group of length-16 Walsh codes is removed. The basic 1XTREME code tree concept is shown in Figure 5-7.

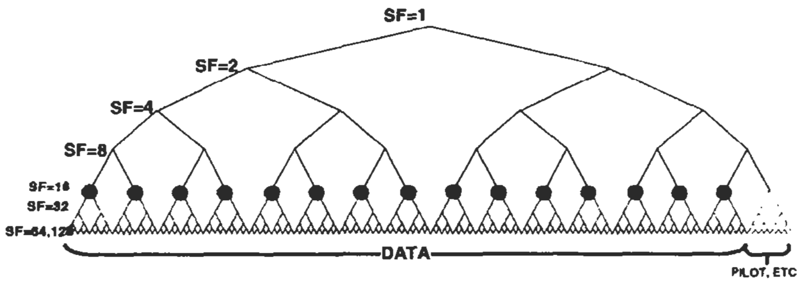


Figure 5-7: 1XTREME Code Tree

Much like 1X-EV-DO, the 1XTREME system makes use of modulation constellations such as 8-PSK, 16-QAM, and 64-QAM to achieve higher data rates over this shared group of codes. In this manner, best-effort data users may be allocated resources for fixed durations of time and still benefit from increased throughput over the forward link. However, much like the traffic channel preamble and the DRC in 1X-EV-DO, 1XTREME requires control channels in both the forward and reverse links to allocate the shared resources among the different best-effort data users.

In addition to the use of a shared channel on the forward link, 1XTREME also employs two features that are not found in cdma2000: fast cell site selection (FCSS) and fast ARQ (also MAC-ARQ). FCSS is a means of providing for mobility for best-effort data users. Like 1X-EV-DO users, 1XTREME users do not employ soft handoff on the forward link, as it is difficult to coordinate identical, simultaneous transmissions of data to the same mobile station over multiple shared channels. FCSS is provided for the mobile to "echo" information about packets it has recently received over the shared channel during a transition from one base station to another to provide for fast synchronization of the packet transmission queues between the new and old base stations.

Fast ARQ is a means of providing fast retransmissions of frames to the mobile station over the shared channel when the mobile fails to deliver an explicit acknowledgement for these frames over the reverse link. Since the retransmission is identical to the original transmitted frame, the mobile can actually perform maximal ratio combining of the channel symbols associated with each frame and thereby improve the decoding performance.

3.1 Forward Link

The 1XTREME forward link consists of three new channels: the forward shared channel (FSHCH), the forward dedicated pointer channel (FDPTRCH), and the forward shared control channel (FSHCCH). The basic spreading mechanism of cdma2000 is still used (see Figure 5-8); however, the inputs to the spreader may be the new channels or existing cdma2000 channels (e.g., forward fundamental channels F-FCHs, FPICH, etc.).

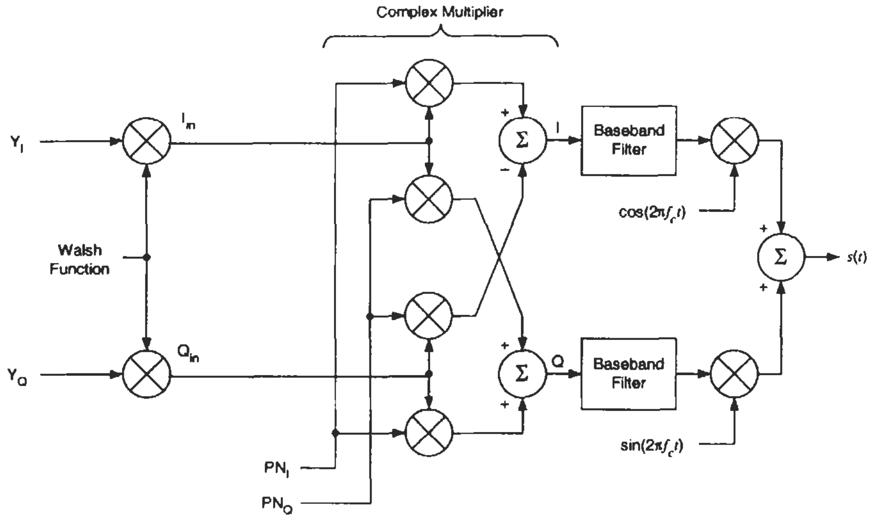
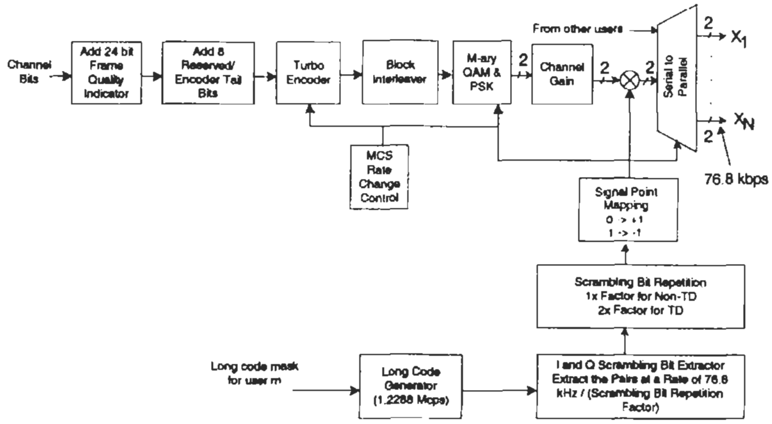


Figure 5-8: 1XTREME (cdma2000) Forward Link Spreading

3.1.1 Forward Shared Channel

The FSHCH carries payload to best-effort data users. It is both time-multiplexed (like 1X-EV-DO) and code multiplexed, in that among the available length-16 Walsh codes that may be allocated to the forward shared channel, at any given instant these codes may be allocated to one or more users. The basic transmission and modulation sequence is shown in Figure 5-9.

After multiplexing the constellation symbols onto N parallel streams, these streams may be spread by different length-16 Walsh codes and demultiplexed for complex spreading (see Figure 5-10).



Total Channel Bits/Frame	Data Rate (kbps)	R	Modulation	Symbols	Data Rate (ksps)
384 x N - 32	76.8 x N	1/2	QPSK	384 x N	76.8 x N
576 x N - 32	115.4 x N	3/4	QPSK	384 x N	76.8 x N
768 x N - 32	153.6 x N	1/2	16-QAM	384 x N	76.8 x N
864 x N - 32	172.8 x N	3/4	8PSK	384 x N	76.8 x N
1152 x N - 32	230.4 x N	3/4	16-QAM	384 x N	76.8 x N
1728 x N - 32	345.6 x N	3/4	64-QAM	384 x N	76.8 x N

Figure 5-9: Forward Shared Channel

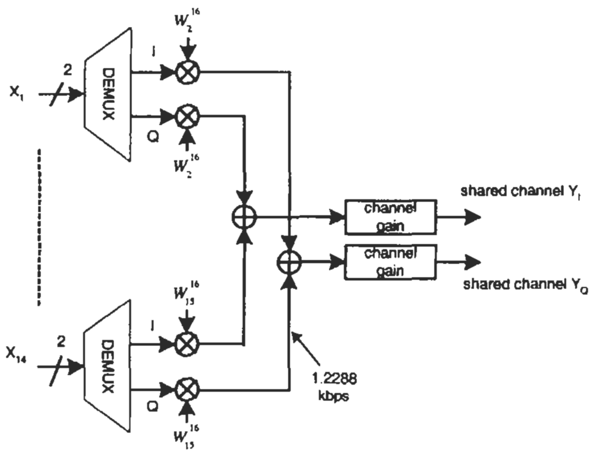


Figure 5-10: Demultiplexing of Parallel Streams for FSHCH

3.1.2 Forward Dedicated Pointer Channel

The forward dedicated pointer channel (FDPTRCH) is assigned to each best-effort data user who may actively receive data on the FSHCH. The transmission sequence is pictured in Figure 5-11. An FDPTRCH is assigned to each mobile station that can actively receive data on the FSHCH. The FDPTRCH carries a 7-bit payload; the first three of these bits are used as a "pointer" field. The pointer field is nominally '000', which implicitly indicates to the mobile that it is not going to receive any concurrent data on the FSHCH. On the other hand, if the pointer field is not '000', then the pointer field indicates to the mobile that it is receiving concurrent data on the FSHCH, and the value of the pointer field provides the mobile a reference to the code channel over which a forward shared common control channel (FSHCCH) is transmitted. The mobile must read the FSHCCH to determine the MCS and number of Walsh codes over which the FSHCH is transmitting data to it. In addition, the mobile receives power control commands time-multiplexed with the FDPTRCH channel bits at an 800 Hz rate; the timing of the insertion of these power control commands is shown in Figure 5-12.

Since the number of mobiles that may be monitoring FDPTRCHs may be large, it would be desirable to minimize the amount of power allocated to these channels and minimize the code space taken up by these channels. This is accomplished by using Walsh codes of length 512, thus increasing the processing gain and reducing the number of codes which must be de-allocated from the code tree when an FDPTRCH is active (see Figure 5-7). However, this results in a limited payload that can be delivered during a 5 ms period. As a result, convolutional coders would not be desirable for forward error correction purposes.

Rather, the FDPTRCH is encoded using a Bose-Chaudhuri-Hocquenghem (BCH) code. BCH encoding involves the modulo-2 inner product of the input bits to the code with the column of the BCH code generator matrix G associated with the output symbol. Therefore, an (n,k) BCH code will encode k input bits into n output bits using a generator matrix G of dimension n rows by k columns. In the case of the FDPTRCH, G would be of dimensions $n = 7$ rows by $k = 16$ columns. Therefore, if the 7 input bits are arranged in a row vector v, then the output row vector o of 16 bits may be determined as

$$o = vG \tag{5-2}$$

BCH codes generally have limited error detection and correction capability; however, since the payload is small (7 bits), this amount of error detection and correction is sufficient for the FDPTRCH.

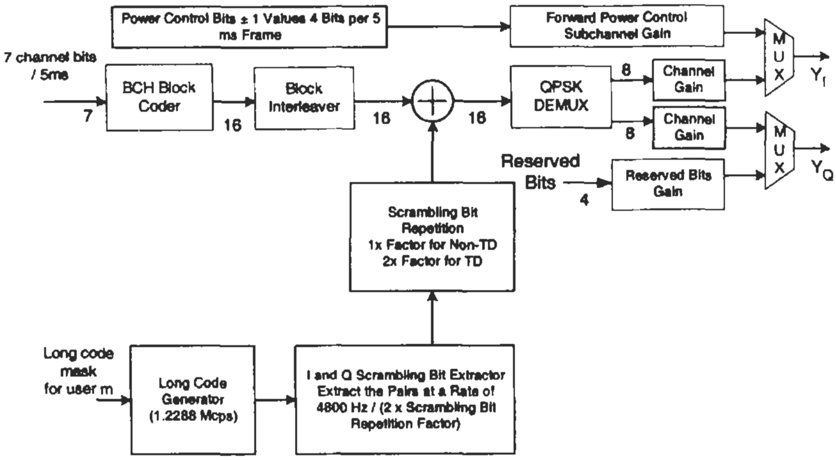


Figure 5-11: 1XTREME Forward Dedicated Pointer Channel

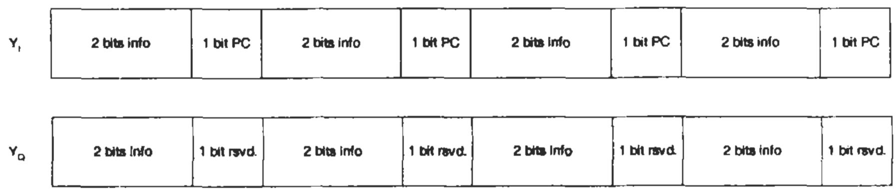


Figure 5-12: FDPTRCH Timing

3.1.3 Forward Shared Control Channel

The forward shared control channel (FSHCCH) includes the information necessary for the mobile to demodulate and decode an incoming frame on the FSHCH. The mobile is triggered to read the FSHCCH when it receives a non-'000' pointer field on the FDPTRCH. This field indicates an index to a length-256 Walsh code over which the intended FSHCCH is transmitted. The possible pointer field values on the FDPTRCH and the associated FSHCCH Walsh codes are given in Table 5-7.

PTR	F-DPTRCH Format	SHCCH Walsh Code
'000'	NULL	N/A
'001'	ASSIGNMENT	80
'010'	ASSIGNMENT	208
'011'	ASSIGNMENT	48
'100'	ASSIGNMENT	176
'101'	ASSIGNMENT	112
'110'	ASSIGNMENT	240
'111'	Reserved	N/A

Table 5-7: Pointer Field of FDPTRCH

The basic transmission sequence of the FSHCCH is shown in Figure 5-13. Note that the payload size (26 bits) is larger than that of the FDPTRCH. Given that this payload is appended with a 10-bit cyclic redundancy check (for error detection), a (48,36) BCH code is used for forward error correction.

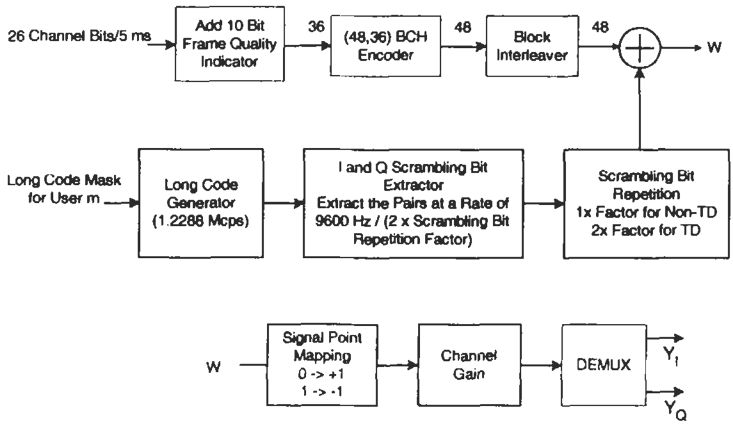


Figure 5-13: 1XTREME Forward Shared Control Channel

The payload carried over the FSHCCH has to contain the minimum information necessary to communicate to a single mobile station the following pieces of information necessary for FSHCH allocation:

a. MCS used
b. The "base code," i.e., the first code index among the Walsh codes allocated to the mobile station
c. The "last code," i.e., the last code index among the Walsh codes allocated to the mobile station.

These pieces of information are sufficient to allocate all or part of the available Walsh codes on the FSHCH to a single mobile station. It should be noted that this information might be signaled in other formats (e.g., base code and number of codes allocated instead of last code).

In addition, the FSHCCH carries information necessary for the proper functioning of the 1XTREME ARQ mechanism; this information will be examined in more detail in Section 3.4.

3.2 Reverse Link

The 1XTREME reverse link leverages heavily from the cdma2000 reverse link. In order to ensure minimal changes to the cdma2000 reverse link, the primary modification is the addition of a new code channel for control information related to the FSHCH allocation.

The basic channelization and spreading structure are given in Figure 5-14. The primary difference between the cdma2000 reverse link and the 1XTREME reverse link is the presence of the reverse quality indicator and echo channel. However, some changes to the reverse supplemental channel (RSCH) are also necessary to increase data rates.

3.2.1 Reverse Quality Indicator and Echo Channel

The reverse quality indicator and echo channel (RQIECH) is used for delivery of three types of control information:

a. Signal quality measurements of the forward pilot channel (FPICH) for allocation of the FSHCH.
b. Fast acknowledgements of FSHCH frames for ARQ.
c. Information for fast cell site selection.

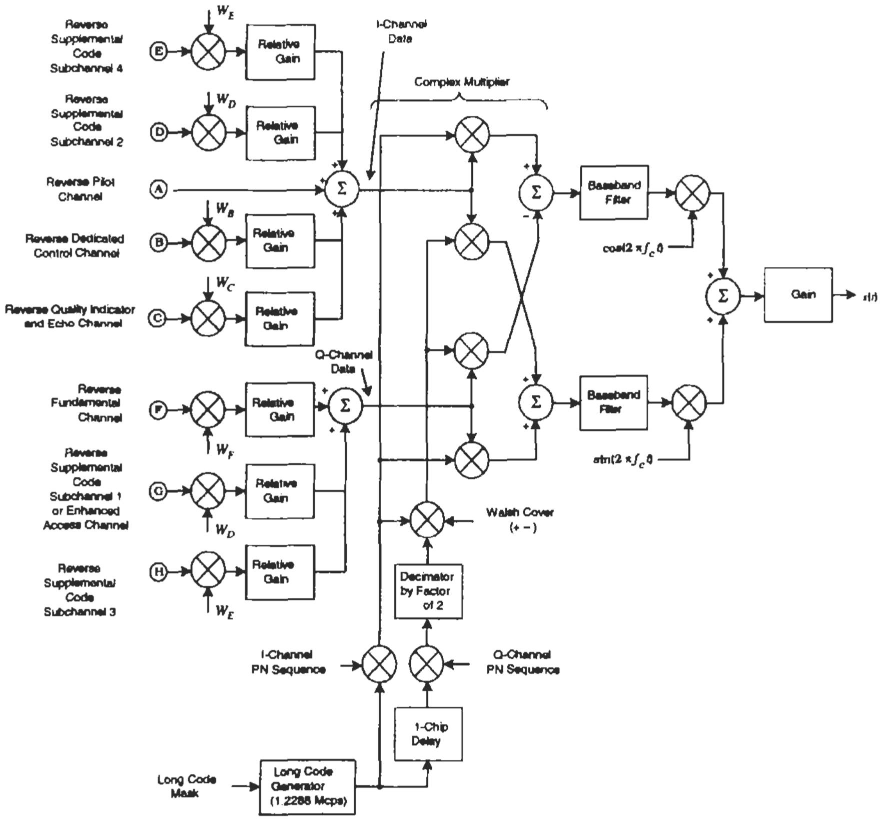


Figure 5-14: 1XTREME Reverse Link Spreading

The RQIECH can be "half-rate" or "full-rate," depending on the type of information that will be transmitted (see Figure 5-15 and Figure 5-16). In addition, an acknowledgement bit (ACK bit) is multiplexed into the RQIECH; this ACK bit is transmitted through the reverse acknowledgement indicator subchannel RAISCH.

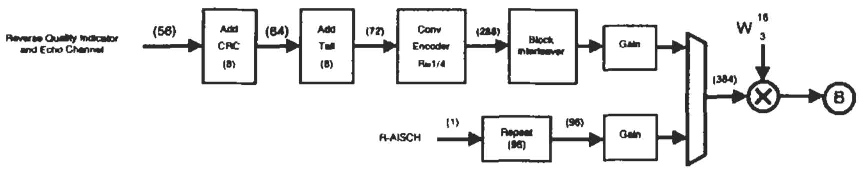


Figure 5-15: Full-rate Reverse Quality Indicator and Echo Channel

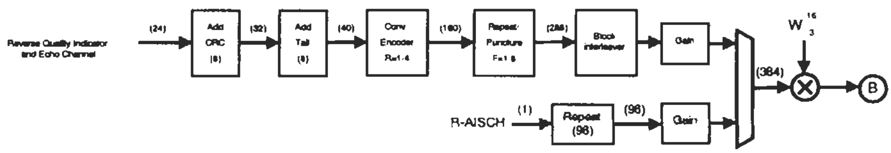
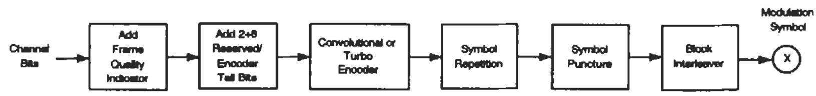


Figure 5-16: Half-rate Reverse Quality Indicator and Echo Channel

3.2.2 Reverse Supplemental Channel

The reverse supplemental channel (RSCH) is nearly identical to the cdma2000 RSCH (see Figure 5-17). However, in order to achieve data rates up to 614 kbps, more code channels are employed, and a pure QPSK transmission method is used. As a result, a demultiplexing step is necessary for RSCH transmission (see Figure 5-18).



Bits/Frame	Bits	Data Rate (kbps)	R	Factor	Deletion	Symbols	Rate (ksps)
28 Bits/5 ms	12	9.6	1/4	2x	None	384	76.8
72 Bits/5 ms	16	19.2	1/4	1x	None	384	76.8
168 Bits/5 ms	16	38.4	1/4	1x	None	768	153.6
360 Bits/5 ms	16	76.8	1/4	1x	None	1,536	307.2
744 Bits/5 ms	16	153.6	1/4	1x	None	3,072	307.2
1,512 Bits/5 ms	16	307.2	1/4	1x	None	6,144	307.2
2,280Bits/5 ms	16	460.8	3/8	1x	None	6,144	307.2
3,048 Bits/5 ms	16	614.4	1/2	1x	None	6,144	307.2
76 Bits/10 ms	12	9.6	1/4	2x	None	768	76.8
168 Bits/10 ms	16	19.2	1/4	1x	None	768	76.8
360 Bits/10 ms	16	38.4	1/4	1x	None	1,536	153.6
744 Bits/10 ms	16	76.8	1/4	1x	None	3,072	307.2
1,512 Bits/10 ms	16	153.6	1/4	1x	None	6,144	307.2
3,048 Bits/10 ms	16	307.2	1/4	1x	None	12,288	307.2
4,584 Bits/10 ms	16	460.8	3/8	1x	None	12,288	307.2
6,120 Bits/10 ms	16	614.4	1/2	1x	None	12,288	307.2
172 Bits/20n ms	12	9.6/n	1/4	2x	None	1,536	76.8/n
360 Bits/20n ms	16	19.2/n	1/4	1x	None	1,536	76.8/n
744 Bits/20n ms	16	38.4/n	1/4	1x	None	3,072	153.6/n
1,512 Bits/20n ms	16	76.8/n	1/4	1x	None	6,144	307.2/n
3,048 Bits/20n ms	16	153.6/n	1/4	1x	None	12,288	307.2/n
6,120 Bits/20n ms	16	307.2/n	1/4	1x	None	24,576	307.2/n
9,192 Bits/20n ms	16	460.8/n	3/8	1x	None	24,576	307.2/n
12,264 Bits/20n ms	16	614.4/n	1/2	1x	None	24,576	307.2/n

Figure 5-17: 1XTREME Reverse Supplemental Channel

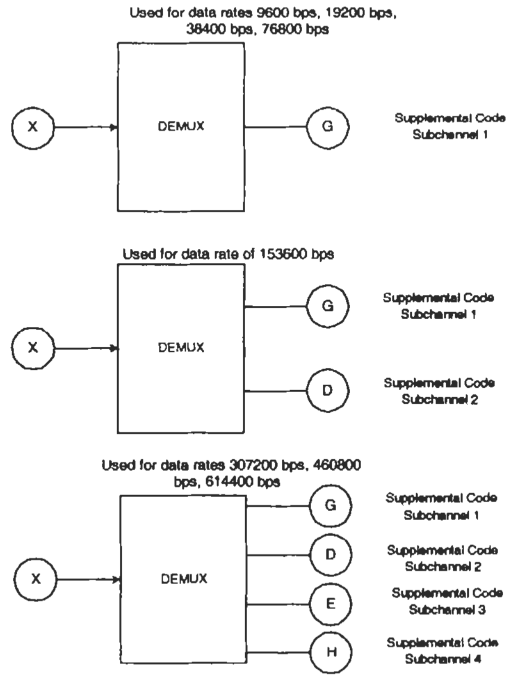


Figure 5-18: RSCH Demultiplexing

3.3 Fast Cell Site Selection

The 1XTREME system supports mobility through the network based on the use of an *eligible set*. The eligible set is the set of base stations (or sectors) to which the mobile can execute a handoff. The mobile can only receive data over the FSHCH from one base station or sector in the eligible set at any given instant in time, so it is important that the wireless network can quickly execute the handover when the mobile station moves from one base station or sector to another.

The network must ensure that each base station in the eligible set can allocate an FDPTRCH at any given instant in time. This is necessary for the mobile to quickly resume service after executing a fast cell site selection. The mobile indicates the selection of the new base station through the *transmit sector indication* (TSI) field in the message transmitted on the RQIECH. The TSI is 9 bits in length, and covers all the possible 512 PN offsets that a base station may use in a cdma2000 system. When the network detects a change in the TSI field received over the RQIECH (clearly to another TSI in the eligible set), an indication of handover is determined. Moreover, if the base station or sector corresponding to the old TSI (denoted as the *serving site*) detects a change in the TSI field in the incoming RQIECH frame, it must cease transmitting over the FSHCH within 10 ms of reception of the frame. In addition, the base station corresponding to the new TSI (denoted as the *selected site*) must initiate transmission over the FDPTRCH to the mobile within 10 ms of receiving the RQIECH frame. It should be noted that the selected site is not required to actually start sending traffic over the FSHCH within 10 ms of a fast cell site selection. In fact, the selected site may delay delivery of traffic to the mobile station due to many reasons, among them:

a. Delays due to serving other users on the selected site.
b. Delays due to synchronization with the serving site for transfer of status information about the mobile related to the call.
c. Lack of data to send, i.e., the fast cell site selection took place during a period of low traffic activity.

The mobile station is still in soft handoff on the reverse link. This means that all the base stations in the eligible set are listening to the reverse control and traffic channels sent by the mobile, e.g., the RQIECH, the R-SCH, etc. However, a base station or sector may not transmit to the mobile over the FSHCH unless it has been selected.

Among the reasons that a selected site may delay in providing traffic over the FSHCH to the mobile during a fast cell site selection is synchronization delays. Basically, several packets may have been delivered over the FSHCH to a particular mobile by the serving site. Some of those packets may not have been delivered or may not have been acknowledged, and therefore are in the serving site's transmit buffer. In order for the selected site to seamlessly resume service to the mobile, it must know what the transmit buffer status is in the serving site. This synchronization step can be performed by direct base-station-to-base-station communication through a wired interface. However, in some networks, this kind of communication may be relatively low bandwidth and thus the service delays to the mobile can be large.

1XTREME offers an alternative method of providing this synchronization by using the full-rate RQIECH. The full-rate RQIECH carries a larger payload and thus more information related to the serving site transmit buffer status can be provided. In particular, each 5-ms FSHCH frame is composed of one or more *protocol data units* (PDUs) that carry traffic or signaling.

1XTREME allows for the appending of a sequence number to each of these PDUs so that the mobile station may keep track of the serving site transmit buffer status. When the mobile station is ready to transition from the serving base station to a selected base station, it changes the TSI field in the outgoing RQIECH message and can be configured to "echo" the sequence number corresponding to the most recently received in-sequence PDU; as a result, the larger payload provided by the full-rate RQIECH is required.

For most types of data applications, a *radio link protocol* (RLP) is used. This provides a layer of reliability below the application (most likely being delivered over an IP stack) by utilizing retransmissions of missing frames of data. It also is important for segmenting and reassembling data that is either being sent to or received from a higher layer. If an RLP instance is used to generate each of the PDUs, then the sequence number used for echo information may be derived from the RLP PDU itself (as it has its own sequencing mechanism).

The most likely network architecture for which the echo information is necessary is depicted in Figure 5-19. In this figure, a central RLP entity (located in the BSC most likely), delivers identical PDUs to all the base stations in the eligible set for delivery on the forward link to a particular mobile. However, only the serving base station knows which PDUs have actually been sent on the FSHCH. Therefore, the selected base station can use the echo information to update which PDUs have not successfully been sent to the mobile station without having to directly communicate to the serving base station.

It should be noted that the echo information is not mandatory in a 1XTREME system to make fast cell site selection work. If the network communication between the serving and selected base stations is sufficiently fast and reliable, it would be desirable to synchronize the two base stations through a direct network interface rather than through the mobile station. Even if network communication does not take place in a relatively timely fashion when compared to using the echo information, network synchronization may still be desirable due to the following reasons:

a. Reliability of network communication almost always exceeds wireless communication (such as the RQIECH).
b. If the 1XTREME data user is relatively stationary (as is the common exhibited behavior among data users), then the fast cell site selection procedure will not take place very often. As a result, the delays suffered due to network synchronization are tolerable when compared to overall system throughput.

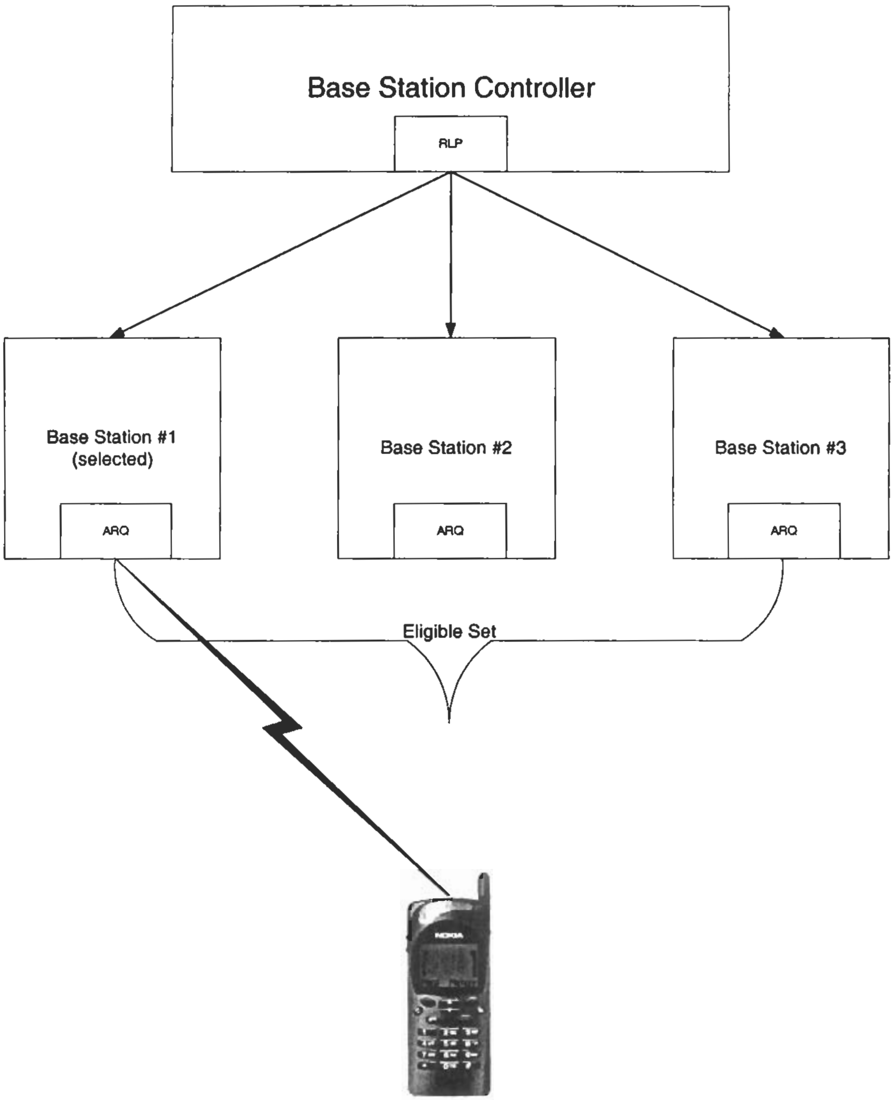


Figure 5-19: Network Architecture for Fast Cell Site Selection

3.4 Fast ARQ

The fast ARQ mechanism (also called MAC-ARQ, or simply "MARQ") in 1XTREME is important in ensuring that some performance loss can be recovered due to the fact that the forward shared channel (FSHCH) is not in soft handoff. Soft handoff provides some limited diversity gain, but it is difficult to implement soft handoff in 1XTREME on the FSHCH due to the fact that each different base station in the eligible set is allocating resources (i.e., number of codes, modulation, scheduling period) to individual mobiles according to each base stations criteria (offered load, forward link signal quality as seen by the mobile from the base station). However, given a fast ARQ mechanism, some diversity gain can be attained by the mobile station for the FSHCH.

ARQ (automatic repeat request) is used in most data protocols to ensure reliability of packet data. In an IP stack, the main layer of reliability is provided through a protocol known as *transmission control protocol* (TCP). The TCP layer, which is immediately above the IP layer in the TCP/IP stack model, retransmits missing octets (8-bit groupings) of data when these octets have not been acknowledged (ACK'ed) by the receiver within a prespecified amount of time. Unfortunately, in wireless data applications TCP tends to request so many retransmissions due to the severity of the wireless transmission channel that often the TCP connection may be terminated due to poor quality.

To overcome this problem, a radio link protocol is used (as briefly discussed in the previous section). RLP generally provides a fixed number of retransmissions for missing PDUs of data (which would include several octets). The maximum number of retransmissions is fixed. In cdma2000, RLP is NAK-based, meaning that retransmissions only occur upon the transmitter receiving a negative acknowledgement (NAK) for one or more missing frames. This method of ARQ is also known as *selective repeat request*.

There are two reasons why RLP cannot provide the functionality needed for MARQ in 1XTREME.

a. As shown in the previous section on fast cell site selection, RLP may terminate at the BSC; this results in network delays in servicing retransmission requests at the RLP layer.

b. The FSHCH frame may contain several PDUs, not all of which come from RLP. In fact, some may be layer 3 messaging intended for the mobile station.

As a result, MARQ provides retransmissions quickly in 1XTREME. 1XTREME employs a stop-and-wait hybrid ARQ method, wherein each packet received by the handset must be acknowledged on a dedicated feedback channel to the base station. This dedicated feedback takes the form of the reverse acknowledgement indicator subchannel (RAISCH), which is time-multiplexed into the RQIECH as a single bit (taking the values +1 for an ACK or -1 for a NAK). However, the mobile does not discard the received soft information associated with the incorrectly received packet. Rather, the mobile buffers the data and coherently combines the buffered data with the received soft information of the retransmission of the bad packet. This type of packet combining provides increased reliability in CDMA systems.

In addition, 1XTREME uses n-phase stop-and-wait MARQ. By "n-phase" it is meant that multiple ARQ instances are employed in consecutive time slots (i.e., 5-ms frame durations). For instance, assume 3 ARQ channels are used:

- In time slot t, the mobile will receive a packet corresponding to phase 1.
- In time slot $t + 1$, the mobile will receive a packet corresponding to phase 2.
- In time slot $t + 3$, the mobile will receive a packet corresponding to phase 3.
- In time slot $t + 4$, the mobile will receive a packet corresponding to phase 1 again, and so on.

The mobile must keep separate packets received for different phases for packet combining and packet acknowledgements; however, once all packets for each phase are received correctly, the mobile may reorder these packets for delivery to higher layers (e.g., RLP).

In order to ensure the mobile does not try to combine packets from one ARQ phase with another, outband signaling is sent on the FSHCCH concurrently with each frame the mobile receives on the FSHCH. Three pieces of information are received on the FSHCCH:

a. The ARQ phase of the frame being sent on the FSHCH (also called "AI" for ARQ instance).

b. The AI sequence number (also called "AI-SN"), usually one bit long.
c. An ARQ abort indication ("AI-ABI"), to tell the mobile to clear packet combining buffers for the AI.

The AI-SN helps the mobile identify retransmissions corresponding to a particular AI on the FSHCH. It also lets the mobile know when an ACK has been misinterpreted as a NAK and vice versa; in the former case, the AI-SN is not incremented, while in the latter case it is. In addition, there is a maximum number of retransmissions allowed for MARQ (up to 10, the exact number determined during call set-up). The AI-SN being incremented will let the mobile know that the maximum number of retransmissions for an AI has been exhausted. In addition, the AI-ABI can be set in such an event to also indicate to the mobile to clear out its packet recombination buffers (the AI-ABI may also be set due to other reasons by the base station such as transmission buffer management). It should be noted that when a fast cell site selection takes place, all AIs are aborted and reset.

If the mobile receives an FSHCH frame on time slot t, it is not mandatory that the mobile receive another FSHCH frame on time slot $t + 1$. The base station may assign the FSHCH to another mobile at that time (due to some quality-of-service requirements or to maximize system throughput). However, the mobile must deliver the ACK/NAK corresponding to the frame received on time slot t within a prespecified time after receiving this frame. If the number of ARQ phases used for the 1XTREME session is denoted as $NARQP$, then the mobile must send the ACK/NAK out to be successfully received and decoded by the base station within $NARQP - 1$ slots after receiving the frame. An example of the timing for $NARQP = 3$ is shown in Figure 5-20.

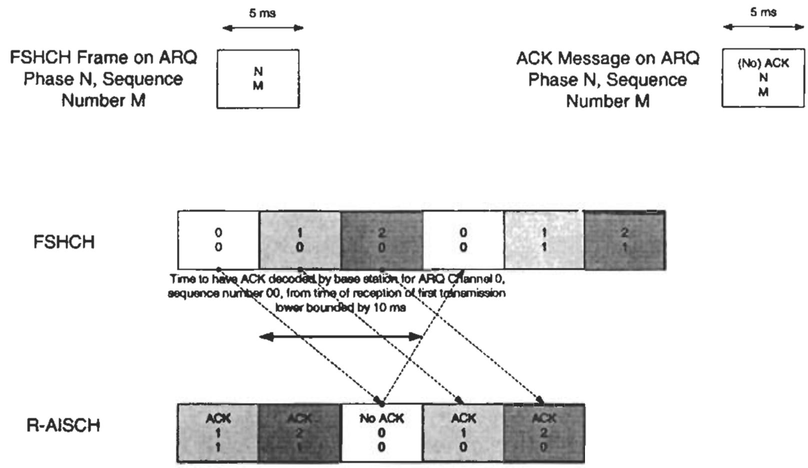


Figure 5-20: *NARQP* = 3 Timing

The mobile station does not deliver ACK/NAKs precisely at slot boundaries. The actual physical layer timing is more precise. This is due to the fact that normally when using coherent receivers in the reverse link (as in cdma2000, 1X-EV-DO and 1XTREME), the base station receiver suffers some delay due to the channel estimation process. In addition, after detecting an ACK or NAK, the base station has more processing to perform to prepare a retransmission for delivery. As a result, the actual timing of the ACK/NAK delivery from the mobile is not precisely within *NARQP* − 1 slots after receiving the frame; rather it commences at T_{AISCH} milliseconds after the reception of the corresponding FSHCH frame, T_{AISCH} determined as

$$T_{AISCH} = 5(NARQP - 2) - 1.25 \tag{5-3}$$

Using this timing, the mobile station can make a determination as to when to deliver the ACK/NAK indication by multiplexing the RAISCH into the outgoing RQIECH frame. This precise timing for *NARQP* = 4 is shown in Figure 5-21.

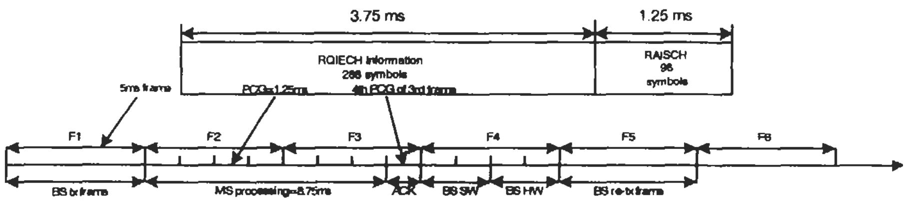


Figure 5-21: RAISCH Timing for *NARQP* = 4

In fact, the mobile station must deliver ACK/NAKs not only when it is receiving FSHCH frames. This is due to the fact that the mobile station may have missed an FSHCH frame also due to not receiving an FDPTRCH or FSHCCH correctly. Therefore, the mobile must ACK or NAK frames on the FDPTRCH. If the FDPTRCH is received correctly and points to an FSHCCH, then the mobile must determine if the FSHCCH is received correctly. Finally, the mobile station may check the FSHCH frame reliability. This entire process is described in the flowchart in Figure 5-22.

4. CONCLUSIONS

The 1X-EV systems are different from the previous IS-95 and cdma2000 systems in several areas, including the type of modulation and coding used, the approach to handovers, and the use of ARQ methods. As a result, as these systems come into deployment, many of these features will most likely be modified from their current state (which also occurred with IS-95 systems in their early deployment). In addition, 1X-EV-DV systems face the additional task of real-time resource allocation between voice and data users, where the data users are packet-switched rather than circuit-switched. Nevertheless, these systems can provide enhancements in data throughput as compared to cdma2000 and IS-95.

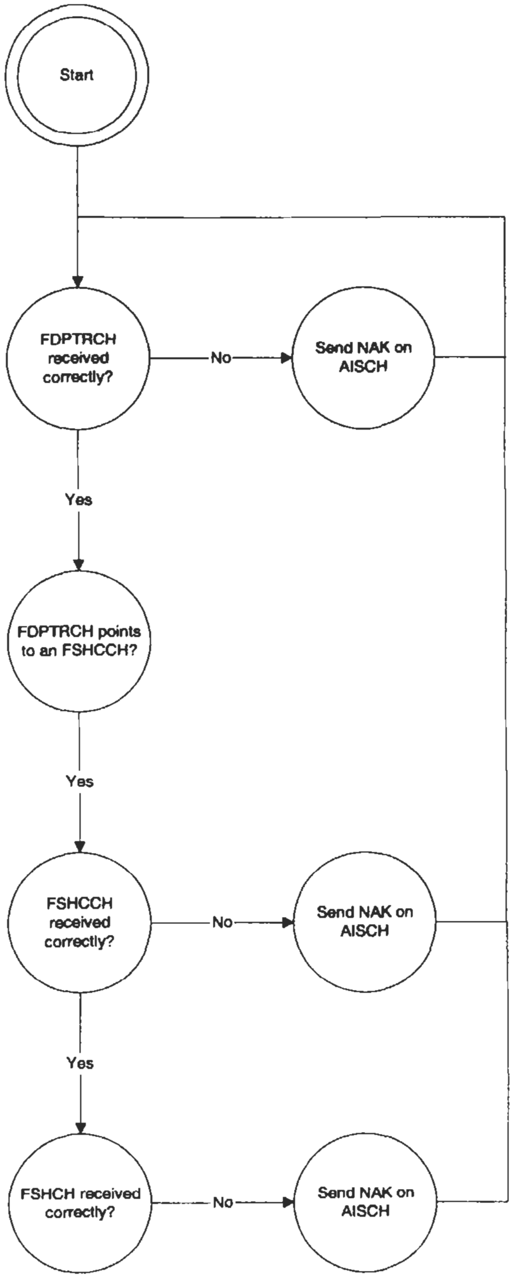


Figure 5-22: 1XTREME MAC ARQ Decision Flow

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Chapter 6

WCDMA Overview

Key Topics: WCDMA, TDD, FDD, HSDPA, UMTS

Overview: A description of the wideband CDMA air interface is provided in this section. The basic spreading and modulation parameters are introduced. Moreover, a standardization history and present spectrum plan are also discussed.

1. INTRODUCTION

1.1 Standardization History

The wireless evolution towards a third-generation global standard started in 1985. The International Telecommunications Union (ITU) has attempted to harmonize the third-generation technologies into a common global radio interface. In April 1997, ITU asked for candidate technologies for the International Mobile Telephony 2000 (IMT-2000) project. Several proposals were received, including WCDMA (ETSI, ARIB), UWC-136 (UWCC), cdma2000 (TIA).

In January 1998, ETSI TC-SMG decided to select WCDMA as its Universal Mobile Telephony System (UMTS) radio technology. The Japanese operator NTT DoCoMo also supported this proposal. During 1998, ETSI and ARIB agreed to make a common UMTS standard under the Third Generation Partnership Project (3GPP). The outcome of the 3GPP work will be a complete set of specifications defining the 3G-network functionality, procedures, and service aspects.

The 3GPP plans to release the specifications on a yearly basis. The first release, Release 99 (3GPP R99), focuses on backwards-compatibility and interoperability with GSM networks. The second release, 3GPP R00, includes two parts (3GPP R4 and 3GPP R5). Release 4 contains minor modifications to Release 99, and the changes in the UMTS core network circuit-switched data flows and control mechanisms. Release 5 contains the items such as High Speed Downlink Packet Access (HSDPA), Radio Resource Management (RRM), uplink synchronous transmission, and IP-based transport.

1.2 Spectrum Allocation for Third-Generation Systems

For 3GPP WCDMA systems, there are two major variants

- FDD: Frequency Division Duplex
- TDD: Time Division Duplex

FDD uses two radio frequencies separately for uplink and downlink transmissions. A pair of 60 MHz frequency bands is allocated for the uplink and downlink spectrum (paired bands). For TDD mode, uplink and downlink transmissions use the same radio frequency with synchronized time intervals. Two frequency bands of 20 and 15 MHz are allocated for TDD operation (unpaired bands). The spectrum allocation for these two modes is

- 1920–1980 MHz: FDD Uplink
- 2110–2190 MHz: FDD Downlink
- 1900–1920 MHz: TDD
- 2020–2025 MHz: TDD

The 3G frequency band allocation for Europe, the USA, and China is shown in Figure 6-1. In most countries, the FDD (paired) spectrum is identical to the one used in Europe, including China, Korea, and Japan. The USA is the exception (in addition to Canada and some other countries), where the 1900 MHz spectrum is used for PCS (PCS1900)

1.3 WCDMA Parameters and Features

Compared to IS-95/cdma2000's 1.25 MHz carrier bandwidth (1.2288 Mcps spreading rate), WCDMA uses a 5 MHz bandwidth per carrier (3.84 Mcps spreading; see Figure 6-2). Note that the TDD mode also allows 1.28 Mcps spreading as an option. Because of the higher carrier bandwidth/

spreading compared to IS-95, WCDMA provides an increase in the processing gain G_p for the same data rate and higher receiver multipath resolution[1] (see also Chapter 3).

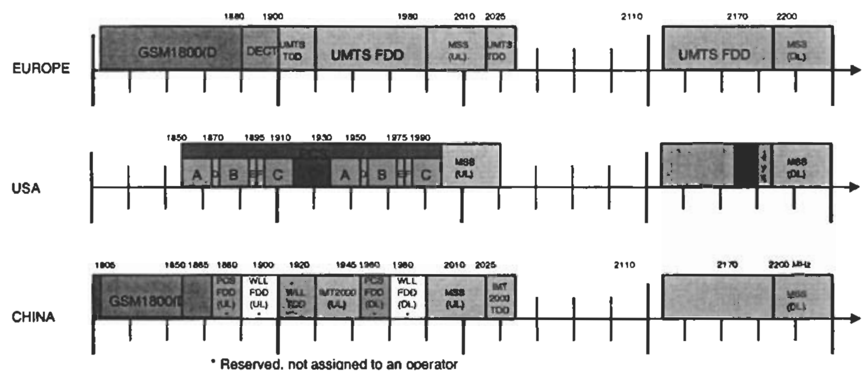


Figure 6-1: Spectrum Usage in Europe, the USA, and China

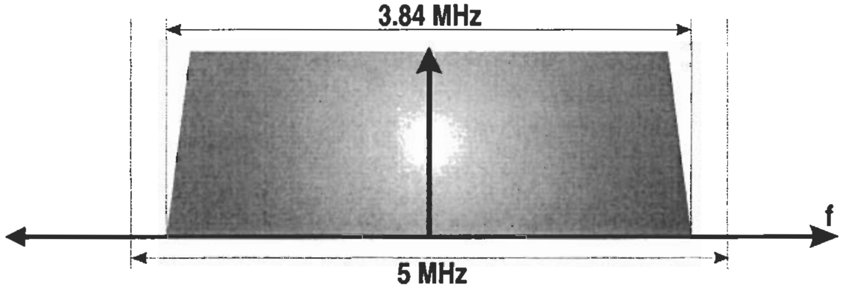


Figure 6-2: WCDMA Carrier Bandwidth

WCDMA uses a 10 ms radio frame, and is subdivided into 15 time slots (0.667 ms per slot). For 3.84 Mcps rate, there are 38400 chips per frame and 2560 chips per slot.

Table 6-1 shows the major WCDMA parameters. In the following sections, every parameter will be explained in detail.

[1] The number of resolvable multipaths is approximately W/B_c, where B is the carrier bandwidth, and B_c is the coherence bandwidth.

Frequency bands	DL: 2110–2170 MHz UL: 1920–1980 MHz	FDD mode (For Region 2* DL: 1930–1990 MHz UL: 1850–1910 MHz) *Region 2: PCS bands
UARFCN[2]	$N_d = 5 * F_{downlink}$ $N_u = 5 * F_{uplink}$	$0.0\ MHz \leq F_{uplink} \leq 3276.6\ MHz$ where $F_{uplink}/F_{downlink}$ is the uplink/downlink frequency in MHz
Chip rate	3.84 Mcps	
Frame duration	10 ms	
Spreading factors	4 – 512 (FDD DL) 4 – 256 (FDD UL)	Spreading factors are 1–16 for TDD mode.
Modulation symbol rate	960 ksps – 7.5 ksps (FDD DL) 960 ksps – 15 ksps (FDD UL)	
Channelization codes	OVSF codes	
Scrambling codes		
Modulation	DL: QPSK UL: HPSK	
Fast power control	1500 Hz	

Table 6-1: WCDMA Parameters Table

Note that for the FDD 1920–1980 MHz / 2110–2170 MHz bands, the UARFCNs are 9612–9888 (uplink, or UL) and 10562–10838 (downlink, or DL) [1].

[2] UARFCN: UTRA Absolute Radio Frequency Channel Number.

2. WCDMA CHANNELS

2.1 Radio Interface Architecture

In order to handle the network complexity, WCDMA uses a layered architecture like other wireless systems. The layered system enables easier network modifications, which helps for system evolution. For the radio interface between base station (denoted as Node B) and the mobile station (denoted as user equipment, or UE), a 3-layer protocol is implemented. These layers are

- The network layer (L3);
- The data link layer (L2);
- The physical layer (L1).

Layer 3 provides the functions for radio resource management (RRM) and radio resource control (RRC), mobility management (MM), connection management (CM), and logical link control (LLC). It is divided into the control plane (C-plane) and the user plane (U-plane) (see Figure 6-3). The control plane carries the application interface and the signaling bearers. The user plane offers frame protocol (FP) for user data transfer through the interface and the underlying transport protocols.

Layer 2 provides services and functionality such as medium access control (MAC), radio link control (RLC), packet data convergence protocol (PDCP), and broadcast/multicast control (BMC) (see Figure 6-3). Note that PDCP and BMC only exist in the U-plane.

Layer 1 transports information to and from the MAC and higher layers.

2.1.1 Logical, Transport, and Physical Channels

WCDMA system also uses a 3-layer channel structure to carry control information and user data between layer 2 and layer 1. These three types of channels are logical channels, transport channels, and physical channels. In the following subsections, the functionality and relationship between these different channels will be described.

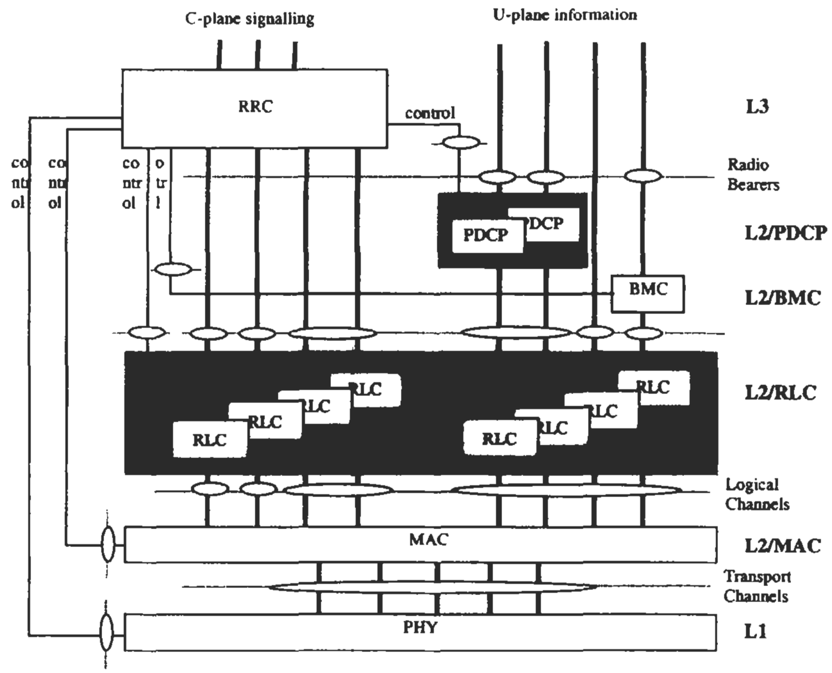


Figure 6-3: WCDMA Radio Interface Protocol Model

2.1.1.1 Logical Channels

The logical channels provide data transfer service from the MAC layer, and they are categorized into two groups: control channels and traffic channels (see Figure 6-4).

Control Channels (CChs):

- BCCH — Broadcast control channel: A downlink channel for broadcasting system control information, such as the spreading code values of a cell and neighboring cells, the allowed transmitted power, and other system parameters.
- PCCH — Paging control channel: A downlink channel used for transferring paging information. It is utilized when the network wants to communicate with the UE but does not know its exact location.
- CCCH — Common control channel: Bi-directional channel for transmitting control information between the network and UEs.

- DCCH — Dedicated control channel: A point-to-point bidirectional channel that transmits dedicated control information between a single UE and the network.

Traffic Channels (TrChs):

- DTCH — Dedicated traffic channel: A point-to-point channel that is dedicated to one UE service for user information transfer.
- CTCH — Common traffic channel: A point-to-multipoint downlink channel to transmit dedicated user information for all or a group of specified UEs.

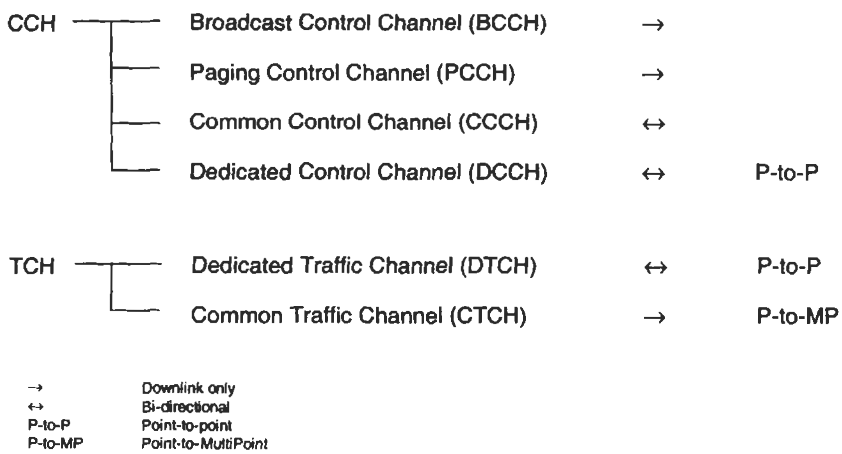


Figure 6-4: Different Types of Logical Channels

2.1.1.2 Transport Channels

Transport channels (TrChs) are responsible for mapping L2 information to L1. All transport channels are unidirectional. The connection between Node B and one UE can carry one or several transport channels simultaneously (both uplink and downlink). Transport channels are classified into two groups — dedicated transport channel and common transport channels (see Figure 6-5).

Common TrCH	Broadcast Channel (BCH)	DL
	Forward Access Channel (FACH)	DL
	Paging Channel (PCH)	DL
	Random Access Channel (RACH)	UL
	Common Packet Channel (CPCH)	UL
	Downlink Shared Channel (DSCH)	DL
Dedicated TrCH	Dedicated Channel (DCH)	DL/UL

Figure 6-5: Transport Channel types

Dedicated Transport Channel

- DCH — Dedicated channel: This is the only type of dedicated transport channel. The DCH is transmitted over the entire cell and carries all the information intended for the given user from the higher layers, which includes data and control information (UE measurement, handover commands, etc.). A DCH supports variable data rates and service multiplexing. Several DCHs can be processed and multiplexed together by the same coding and multiplexing unit. The single output data stream is called the Coded Composite Transport Channel (CCTrCh). For the downlink, multiple CCTrChs can be used simultaneously for one UE. For the uplink, only one CCTrCh can be used (FDD mode). Figure 6-6 demonstrates how multiple uplink DCHs are multiplexed into a CCTrCH and then split into different physical channels. More detailed procedures are provided in Section 3.1.1.

Common Transport Channels

For the common transport channels, resources are shared between all or a group of users in a cell. Soft or softer handover cannot be used for these channels. The following are the common transport channel types:

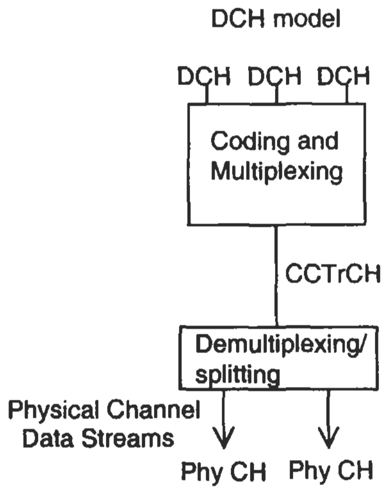


Figure 6-6: Uplink DCHs Multiplexing/Demultiplexing Example

- BCH — Broadcast channel: A downlink channel that carries system and cell-specific information, and is transmitted over the entire cell.
- FACH — Forward access channel: A downlink channel that carries control information to UEs. The FACH can be transmitted over the entire cell or part of a cell.
- PCH — Paging channel: A downlink transport channel that is always transmitted over the entire cell. The PCH carries the information relevant to the paging procedure.
- RACH — Random access channel: An uplink channel that carries control information from the terminal, such as RRC connection set-up requests. The RACH is not collision-free, and is received from the entire cell.
- CPCH — Common packet channel: An uplink packet-based transport channel. CPCH is associated with a dedicated channel on the downlink that provides power control and CPCH control commands for the uplink CPCH.
- DSCH — Downlink shared channel: A downlink channel shared by several UEs. The DSCH carries dedicated user data and/or control information, and is associated with one or several downlink DCHs.

Compressed Mode

Compressed mode is defined as the mechanism whereby certain idle periods are created in radio frames so that the UE can perform measurements during these periods (more details can be found in [3]).

Compressed mode is obtained through layer 2 by using transport channels provided by layer 1 as follows:

- Compressed mode is controlled by the radio resource control (RRC) layer, which configures layer 2 and the physical layer;
- The number of occurrences of compressed frames is controlled by RRC, and can be modified by RRC signaling;
- It is the responsibility of layer 2, if necessary and if possible, to either buffer some layer 2 protocol data units, i.e., PDUs (typically at the RLC layer for NRT services), or to rate-adapt the data flow (similarly to GSM) so that there is no loss of data due to the use of compressed mode. This will be service-dependent and controlled by the radio resource control layer.

For measurements in compressed mode, a transmission gap pattern sequence is defined. A transmission gap pattern sequence consists of alternating two transmission gap subpatterns, and each of these patterns in turn consists of one or two transmission gaps. The transmission gap pattern structure, position, and repetition are defined with physical channel parameters described in [3].

The UE can support a certain number of simultaneous compressed mode pattern sequences, which is determined by the UEs capability to support each of the measurement types categorized by the standard. For example, a UE supporting FDD and GSM shall support four simultaneous compressed mode pattern sequences and a UE supporting FDD and TDD shall support two simultaneous compressed mode pattern sequences.

2.1.1.3 Physical Channels

Physical channels are specified by the carrier frequency, codes (channelization code and scrambling code), and phase (0 or $\pi/2$ for uplink). Similar to transport channels, both uplink and downlink physical channels can be classified as dedicated and common channels.

2.1.1.3.1 Uplink Physical Channels

2.1.1.3.1.1 Dedicated Uplink Physical Channels

There are two types of dedicated uplink physical channels:

- DPDCH (Dedicated physical data channel): The DPDCH channel carries the user data and higher layer signaling from the DCH transport channel, and its bit rate can be changed frame-to-frame (10 ms). The spreading factor for the DPDCH ranges from 4 to 256.
- DPCCH (dedicated physical control channel): The DPCCH channel contains control information such as the pilot bits, feedback information (FBI), transmit power control (TPC), and an optional transport format combination indicator (TFCI). In order to maintain accurate channel estimation, for higher data rates the DPCCH transmit power level needs to be relatively higher than that of the lower bit rates. The spreading factor for the DPCCH is always 256 (15 ksps channel symbol rate). The constant bit rate is to ensure reliable detection.

The DPDCH and DPCCH are I/Q multiplexed within each radio frame with complex scrambling. Figure 6-7 shows the frame structure for the DPDCH and DPCCH.

2.1.1.3.2 Common Uplink Physical Channels

2.1.1.3.2.1 Physical Random Access Channel (PRACH)

The PRACH carries the RACH information from the transport channel. The UE can start transmission at certain predefined time offsets (access slots). The PRACH message consists of two parts, a data part to which the RACH transport channel is mapped, and a control part that carries layer 1 control information (see Figure 6-8). The spreading factor for the data segment ranges from 256 (15 ksps) to 32 (120 ksps). The spreading factor for the control segment is always 256. Note that the data and control segments are transmitted in parallel. More information regarding the RACH structure and transmission can be found in [2].

2.1.1.3.2.2 Physical Common Packet Channel (PCPCH)

The PCPCH carries the transport channel CPCH information. For the CPCH, transmission is based on the Digital Sense Multiple Access-Collision Detection (DSMA-CD) approach. The UE starts transmitting at a defined slot timing that is identical to the RACH. The PCPCH access transmission includes one or several Access Preambles (A-P) over 4096 chips, one Collision Detection Preamble (CD-P), one DPCCH Power Control Preamble (PC-P), and a message part. The message part includes data and control segments, and its frame structure is very similar to DPDCH and DPCCH, mentioned in 2.1.1.3.1.1 and Figure 6-7.

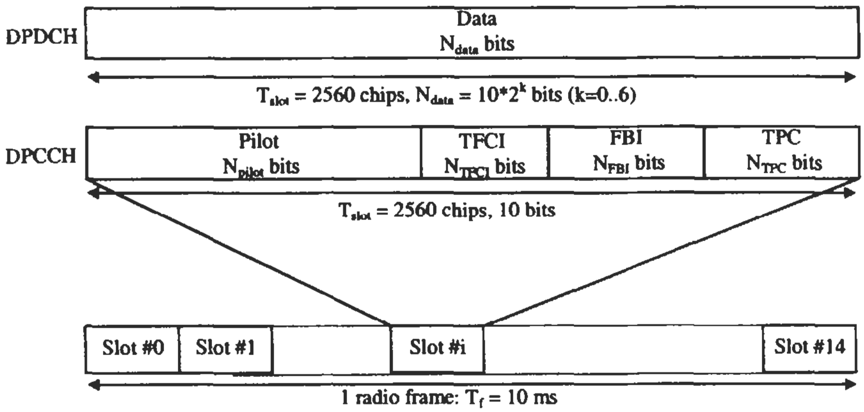


Figure 6-7: Uplink Dedicated Physical Channel Structure

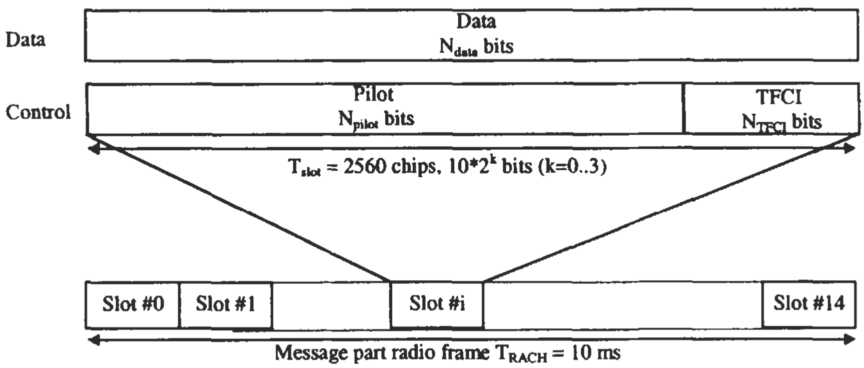


Figure 6-8: PRACH Message Part Structure

2.1.1.3.3 Downlink Physical Channels

2.1.1.3.3.1 Dedicated Downlink Physical Channels

The L2 DCH information is carried by the Downlink Dedicated Physical Channels (downlink DPCHs). Similar to the uplink dedicated channel, the downlink DPCH consists of two types of channels — DPDCH and DPCCH. Unlike the uplink, where the DPDCH and DPCCH are I/Q multiplexed with each frame, downlink DPCHs are time-multiplexed with layer 1 related control information (such as pilot bits, UL TPC, and optional TFCI; see Figure 6-9). The DPCH spreading factor can be 512 (7.5 ksps), 256 (15 ksps), down to 4 (960 ksps). Uplink I/Q multiplexing is used to ensure continuous transmission in order to reduce audible interference. Downlink time multiplexing is used to save the orthogonal codes. Since the downlink common channels are transmitted all the time, discontinuous transmission (DTX) is not used for the downlink.

The downlink DPCH bit rate can change frame by frame, and lower data rate transmission will be handled by DTX. When the total transmitted bit rate in one downlink CCTrCH exceeds the maximum bit rate for a DL physical channel, then multicode transmission can be used. For multicode operation, several parallel downlink DPCHs are transmitted for one CCTrCH using the same spreading factor. In this case, the layer 1 control information will be transmitted only over the first downlink DPCH.

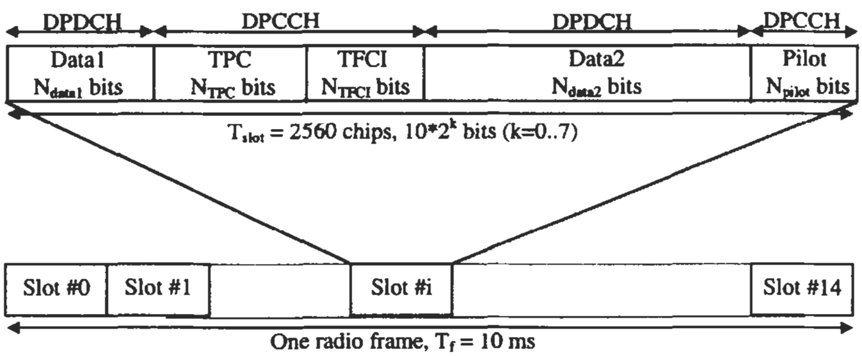


Figure 6-9: Downlink Dedicated Physical Channel Frame Structure

2.1.1.3.3.2 Common Downlink Physical Channels

- Common Pilot Channel (CPICH)

The CPICH is a continuously transmitted pilot channel for downlink channel estimation and for UE intrafrequency measurement of neighboring cells for the soft handovers. The CPICH typically takes 5 to 15 percent of the total BTS transmitted power. This is relatively low overhead compared with IS-95, where the pilot power is typically 20 to 25 percent of the maximum base station transmitted power.

The CPICH uses a predefined bit/symbol sequence, transmitted at 15 ksps (30 kbps) with a spreading factor of 256 (see Figure 6-10). This physical channel can be further divided into two types:

- Primary Common Pilot Channel (P-CPICH): The P-CPICH always uses the same channelization code ($C_{ch,256,0}$), and is scrambled by the primary scrambling code. There is only one P-CPICH per cell and it is broadcast over the entire cell.
- Secondary Common Pilot Channel (S-CPICH): The S-CPICH can use an arbitrary channelization code with a spreading factor equal to 256, and is scrambled by either the primary or a secondary scrambling code. When the S-CPICH is scrambled by secondary scrambling code, it can be used for beam steering in adaptive antenna applications to increase system capacity. Hence, it won't use an additional primary scrambling code and thus cause downlink code planning problems.

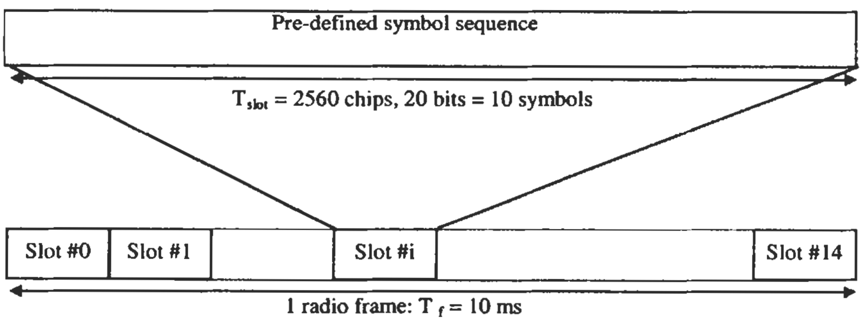


Figure 6-10: Frame Structure for Common Pilot Channel

- Primary Common Control Physical Channel (P-CCPCH)

The P-CCPCH carries the BCH transport channel system and cell information, and is broadcast over the whole cell. It uses a fixed channelization code $C_{ch,256,1}$ (SF = 256), without power control command (TPC), TFCI, or pilot bits. For each 2560-chip slot, the first 256 chips are not transmitted in P-CCPCH. This interval is reserved for the Primary Synchronization Channel (SCH) or Secondary SCH transmission (see Figure 6-11).

- Secondary Common Control Physical Channel (S-CCPCH)

The S-CCPCH carries the FACH and PCH from the transport channel, and these channels can be mapped to the same S-CCPCH or to separate S-CCPCHs. The spreading factor can be from 256 down to 4 (15 to 960 ksps). The S-CCPCH is transmitted only if there is data available. It also supports variable rates when the TFCI bits are included. The frame structure of S-CCPCH is shown in Figure 6-11.

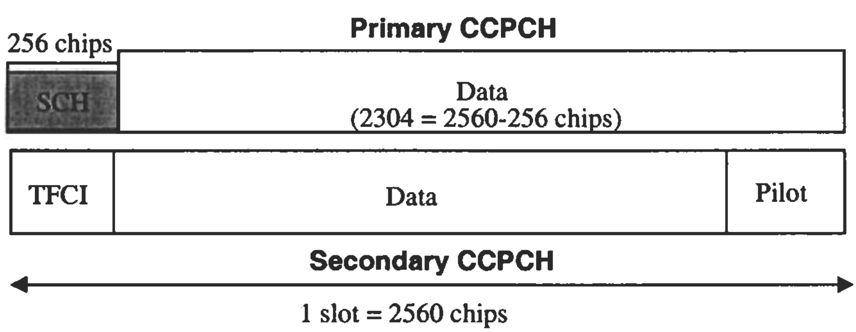


Figure 6-11: P-CCPCH and S-CCPCH Frame Structures

- Synchronization Channel (SCH)

The synchronization channel is used for cell search and slot/frame synchronization. Since the UEs need the SCH information to locate the cell and synchronize before the decoding process, the SCH does not go through the spreading and scrambling process.

The synchronization channel consists of two subchannels: the Primary SCH and the Secondary SCH.

- Primary SCH: It is a 256-chip length code (C_p) and is transmitted at the beginning of every slot (see Figure 6-12). For the UE receiver's to be able to lock on to the strongest SCH among different base stations to obtain slot synchronization, the same Primary SCH code is used for every cell in the system.
- Secondary SCH: The Secondary SCH repeatedly transmits a length-15 sequence of codes $[C_s^{i,0}, C_s^{i,1}, \ldots, C_s^{i,14}]$ with 256 chips per code, where i = 1, 2, ..., 64 corresponds to one of 64 scrambling code groups. Similar to the Primary SCH, each 256-chip code is transmitted at the beginning of every slot (see Figure 6-12). The sequence of the Secondary SCH allows downlink frame synchronization, and indicates to which of the code groups the cell downlink scrambling code belongs.

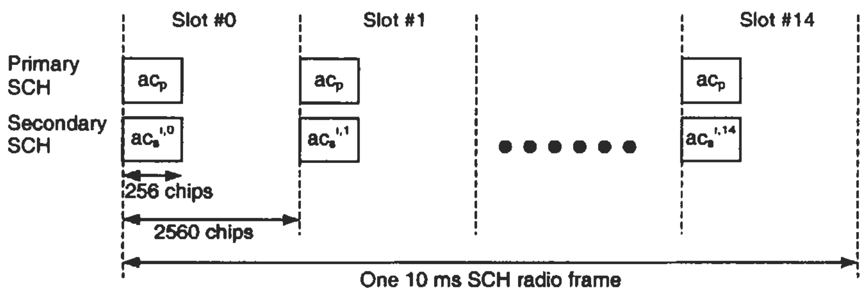


Figure 6-12: Synchronization Channel Structure

Note that in Figure 6-12, the primary synchronization code (C_p) and the secondary synchronization code ($C_s^{i,k}$) are modulated by the symbol "a", which indicates the presence/absence (+1/-1) of transmit-diversity (space-time transmit diversity, or STTD) encoding on the P-CCPCH.

- Acquisition Indictor Channel (AICH)

The AICH carries the layer 1 acknowledgement information corresponding to the PRACH preambles. It is broadcast over the entire cell. When the UE receives the AI acknowledgement, it will start transmitting the PRACH (see Figure 6-13).

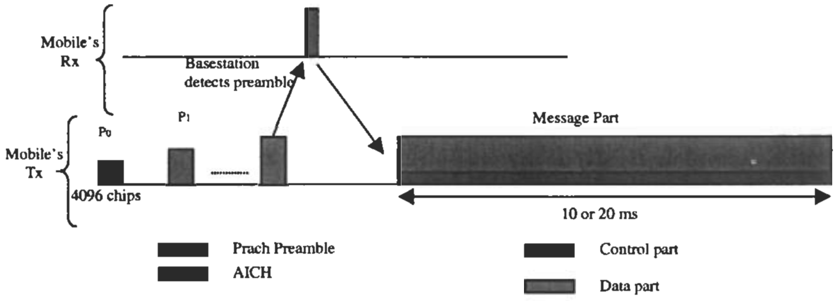


Figure 6-13: PRACH & AICH Transmission Procedure

The AICH uses a fixed rate (256 spreading factor) and consists of a repeated sequence of 15 consecutive Access Slots (AS). There are 5120 chips per AS, therefore, the duration of an AICH is 20 msec. For each access slot, there exist two segments: a 32-length real valued (+1, -1, or 0) Acquisiting Indicator (AI) sequence of symbols with 32 chips in each symbol, and a duration of 1024 bits with transmission off (see Figure 6-14).

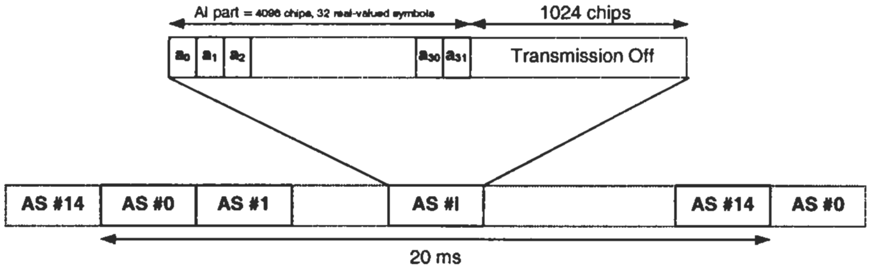


Figure 6-14: Structure of the AICH

- Paging Indictor Channel (PICH)

The Paging Indicator Channel (PICH) is a downlink common channel, which carries the paging indicator for the UEs. The paging channel uses a fixed-rate spreading factor (SF = 256). Therefore one 10 ms PICH radio frame consists of 300 bits (b_0, b_1, ..., b_{299}) (see Figure 6-15). The first 288 bits contains paging indication information, and the last 12 bits are not transmitted. In each PICH frame, N_p paging indicators are transmitted, where N_p = 18, 36, 72, or 144. Each paging indicator corresponds to a certain channel, and a UE only has to wake up from the sleep mode to listen to its specific paging indicator. This will help UE save power in idle mode.

Note that the PICH is always associated with an S-CCPCH, and it is broadcast in the entire cell.

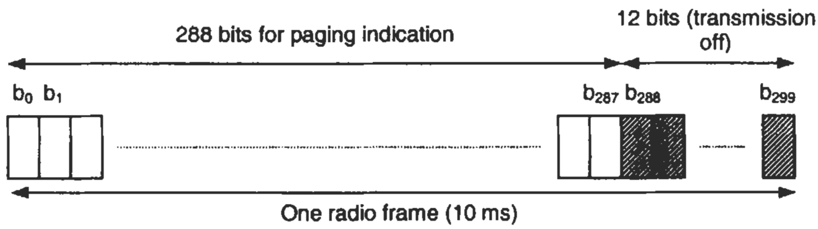


Figure 6-15: Paging Indicator Channel (PICH) Frame Format

- Physical Downlink Shared Channel (PDSCH)

As mentioned in the common transport channel section, the downlink shared channel (DSCH) is shared by a group of downlink users to effectively utilize channelization codes for packet data services. The radio frame–based PDSCH carries the DSCH in a physical channel, and one PDSCH is allocated to a single UE. Within the same radio frame, several PDSCHs can be allocated for a single UE for multicode transmission. Or, different PDSCHs can be used for different UEs for channel sharing using code multiplexing.

The PDSCH is always associated with the DPCH. The TFCI in DPCCH provides the PDSCH information to the UE (such as the transport format parameters and the PDSCH channelization code).

Figure 6-16 illustrates the channel mapping for the Logical, Transport, and Physical channels.

3. WCDMA PHYSICAL LAYER

This section describes the WCDMA physical layer structure and characteristics for FDD mode. The WCDMA physical layer provides the following functions:

- RF processing
- FEC encoding/decoding
- Multiplexing/demultiplexing of the transport channels
- Rate matching

- Mapping transport channels on physical channels
- Power weighting and combining of physical channels
- Spreading/despreading & modulation/demodulation of physical channels
- Frequency and time synchronization (chip, slot, frame)
- Radio measurements including FER, SIR, etc.
- Soft handover execution
- Inner loop power control

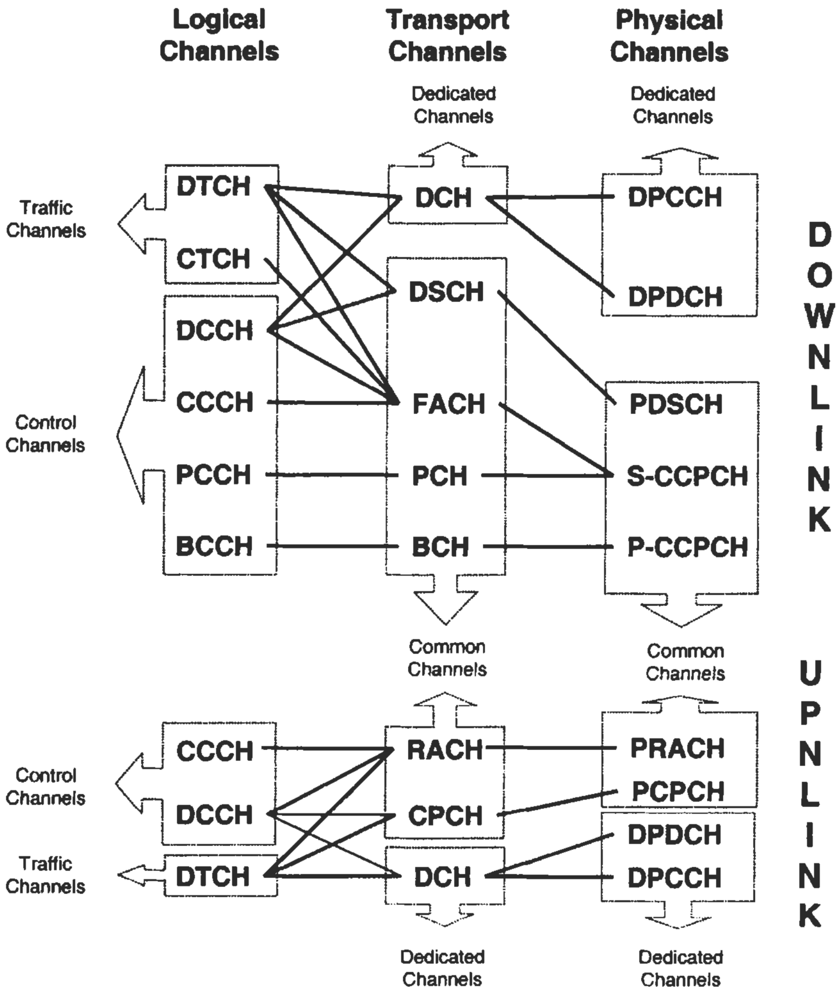


Figure 6-16: Logical-Transport-Physical Channel Mapping

3.1 Data Transmission Processing

WCDMA utilizes a multirate transmission scheme to multiplex different data rates in order to achieve different quality of service (QoS) levels for different users. Figure 6-17 illustrates a simplified transmission block diagram. The transport channel information that includes both user data and control message is processed (coded/interleaved/multiplexed) and mapped to physical channels. In the spreading stage, the physical channels are spread by the channelization code to 3.84 Mcps and then scrambled by the scrambling code for cell (downlink) or UE (uplink) identification. After spreading, the complex-value sequence is QPSK modulated and up-converted to the final radio frequency.

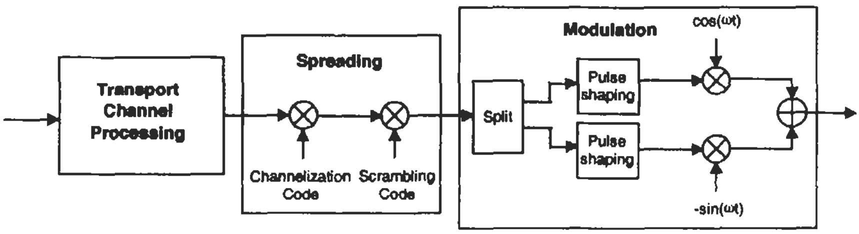


Figure 6-17: Data Transmission Flow Diagram

3.1.1 Transport Channel Processing

Transport channel processing is responsible for mapping layer 2 data onto the physical layer. During this process, the following functions are performed (see Figure 6-18):

- **CRC attachment**: Error detection for transport blocks is made possible by adding Cyclic Redundancy Check (CRC) bits.
- **TrBK concatenation/Code block segmentation**: The transport blocks (TrBKs) are either concatenated or segmented to different coding blocks to fit the available code block size.
- **Channel coding**: Either convolutional codes or turbo codes are used for error correction.
- **Rate matching**: This procedure is used to match the number of transmitted bits to the number of bits available on a single physical channel frame by either puncturing or repetition.
- **Interleaving**: Used to prevent bursty errors.

- **TrCH multiplexing**: When the total transmitted bit rate exceeds the maximum bit rate for a physical channel, then multicode transmission is employed.

3.1.1.1 CRC Attachment

CRC attachment provides error detection capability for the transport blocks. The length of the CRC can be 24, 16, 12, 8, or 0 bits based on the higher layer command. A higher number of CRC bits can provide better error detection capability.

The parity check bits are generated by one of the following cyclic generator polynomials using the entire transport block:

$$
\begin{aligned}
g_{CRC24}(D) &= D^{24} + D^{23} + D^{6} + D^{5} + D + 1 \\
g_{CRC16}(D) &= D^{16} + D^{12} + D^{5} + 1 \\
g_{CRC12}(D) &= D^{12} + D^{11} + D^{3} + D^{2} + D + 1 \\
g_{CRC8}(D) &= D^{8} + D^{7} + D^{4} + D^{3} + D + 1
\end{aligned}
\tag{6-1}
$$

3.1.1.2 TrBK Concatenation/Code Block Segmentation

The transport blocks (TrBKs) in a TTI (transmission time interval) are either concatenated together or segmented to several coding blocks, according to whether the TrBK fits the available code block size (depending on the coding method). Concatenation provides better performance due to the low overhead. Also, because of the larger block size, better channel coding methods can be used to improve performance.

If the transport block does not fit into the maximum available code block, it will be divided into several equal length code blocks. (For convolutional coding, the maximum code block size $Z = 504$. For turbo coding, $Z = 5114$.)

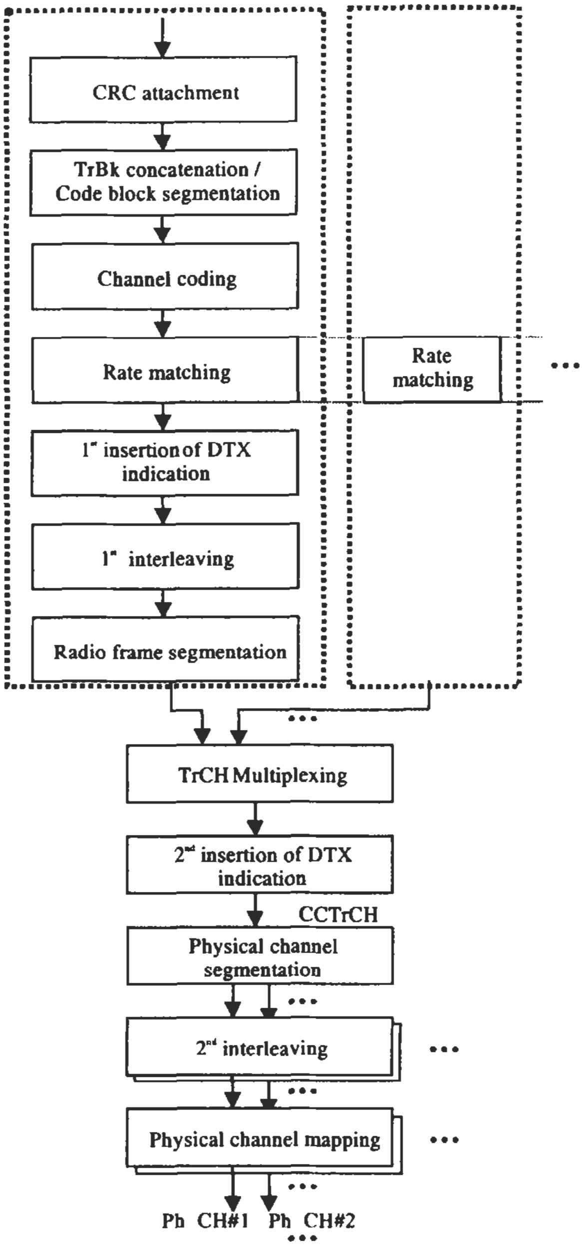


Figure 6-18: Transport Channel Multiplex Structure (DL)

3.1.1.3 Channel Coding

Followed by block concatenation or segmentation, channel coding is performed on the coding blocks. Three options can be implemented on the TrCHs for channel coding:

- Convolutional coding;
- Turbo coding;
- No coding.

Table 6-2 demonstrates the coding schemes and the coding rates for various types of transport channels.

<table>
<tr><th>Type of TrCH</th><th>Coding scheme</th><th>Coding rate</th></tr>
<tr><td>BCH</td><td rowspan="4">Convolutional coding</td><td rowspan="3">1/2</td></tr>
<tr><td>PCH</td></tr>
<tr><td>RACH</td></tr>
<tr><td rowspan="3">CPCH, DCH, DSCH, FACH</td><td>1/3, 1/2</td></tr>
<tr><td>Turbo coding</td><td>1/3</td></tr>
<tr><td colspan="2">No coding</td></tr>
</table>

Table 6-2: Channel Coding Schemes and Rates

3.1.1.3.1 Convolutional Coding

As in cdma2000, convolutional coding is used for lower data rate transmission compared to turbo codes. Constraint length 9 (K = 9) convolutional codes are used in both rates 1/2 and 1/3 for WCDMA FDD application (see Figure 6-19). The initial value of the shift register is the "all 0" state when starting to encode the input bits. Also, eight "0" tail bits are added to the end of the code block before encoding to ensure the encoder returns to the "all 0" state.

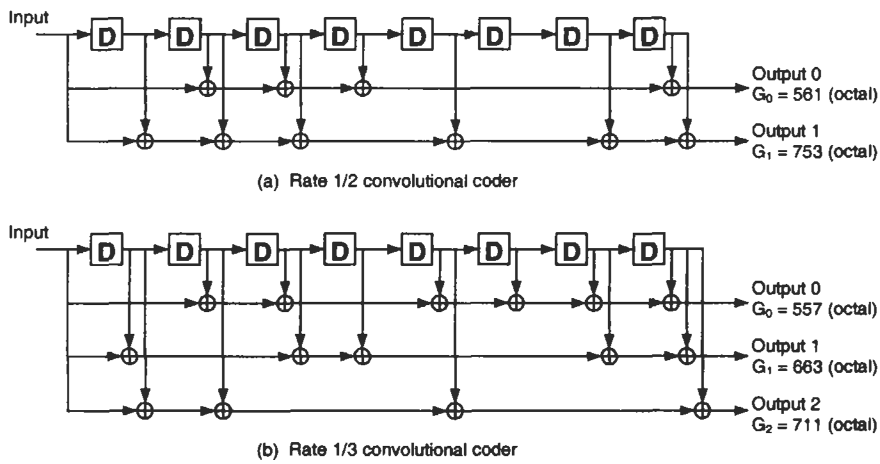


Figure 6-19: Rate 1/2 and 1/3 Convolutional Encoders Used in WCDMA

3.1.1.3.2 Turbo Coding

The turbo code in WCDMA is formed by parallel concatenation of two or more convolutional codes. It was claimed to be able to achieve near the Shannon-limit (see [4]) for error correction performance. The input data sequence is first block-interleaved before encoding to increase coding diversity. The turbo encoder in FDD mode uses a rate of 1/3, 8-state code with transfer function:

$$G(D) = \left[1, \frac{g_1(D)}{g_0(D)}\right], \text{ where} \tag{6-2}$$

$$g_0(D) = 1 + D^2 + D^3$$

$$g_1(D) = 1 + D + D^3$$

Figure 6-20 shows the structure of the turbo encoder. The output from the encoder is x_1, z_1, z'_1, x_2, z_2, z'_2, ... , x_K, z_K, z'_K. Similar to the convolutional encoder, the turbo encoder shift registers begin with all zeros at the start of encoding. Note that the dotted line in Figure 6-20 is only for trellis termination. For the convolutional code, 0's are added to the end of the input data sequence to "flush" the content of the shift register. In the turbo encoder, the feedback from the shift register is taken as the tail bits, after all the information bits are encoded (x_{K+1}, z_{K+1}, x_{K+2}, z_{K+2}, x_{K+3}, z_{K+3}, x'_{K+1}, z'_{K+1}, x'_{K+2}, z'_{K+2}, x'_{K+3}, z'_{K+3}). More information regarding the turbo code internal interleaver can be found in [3].

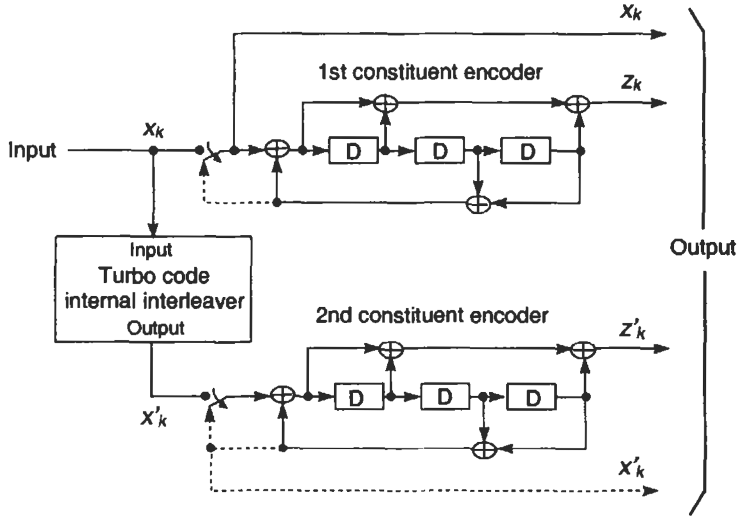


Figure 6-20: Rate 1/3 Turbo Encoder

3.1.1.3.3 Rate Matching

In order to match the number of transmitted bits to the number of bits available on a single physical channel frame, puncturing or repetition is used for rate matching. The rate matching between different transport channels (TrCHs) is provided by a higher layer parameter — the rate-matching attribute. When multiplexing several TrCHs in the same frame, the rate-matching value is calculated using this rate-matching attribute. At the receiver side, the rate-matching parameters can be obtained using the rate-matching attribute and the transport format combination indicator (TFCI).

3.1.1.3.4 Insertion of Discontinuous Transmission (DTX) Indication

Discontinuous transmission is used in the downlink when the data rate is below the maximum transmission rate. In this case, the DTX indication bits are inserted to fill up the radio frame. These DTX indication bits only show when the transmission needs to be turned off, but are not transmitted through the air. The DTX insertion locations are different, depending on whether they are fixed position TrCH[3] or flexible position TrCH[4] (see Figure 6-21):

- First insertion: for the fixed position TrCH only. The DTX indication bits are inserted in the unused channel bit positions.
- Second insertion: for the flexible position TrCH only. The DTX indication bits are inserted at the end of the radio frame.

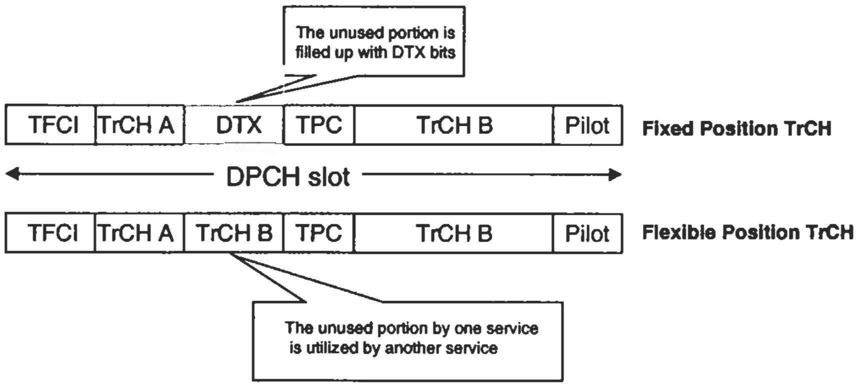


Figure 6-21: Fixed and Flexible Position TrCHs and DTX Example

3.1.1.3.5 Interleaving

There are two interleaving stages during the transmission process. The first interleaving is followed by rate matching (or first DTX indication insertion), which performs radio frame interleaving. The second interleaving is executed just before the physical channel mapping, which carries out the intraframe interleaving.

- **First interleaving:** The first interleaving performs intercolumn permutations, and is used only when the transmission time interval (TTI) is more than 10 ms. The length of the first interleaving has to be 20, 40, or 80 ms.

[3] For fixed position TrCH, service always uses the same bit in the DPCH; if the transmission rate is below the maximum rate, then DTX indication bits are inserted.

[4] For flexible position TrCH, the channel bits unused by one service may be utilized by another service.

- **Second interleaving:** The second interleaving performs intraframe interleaving (for 10 ms radio frames). It utilizes a 30-column block interleaver for intercolumn permutation. For every physical channel, this second interleaving is done separately.

3.1.1.3.6 Radio Frame Segmentation

When the first interleaving is used, the frame segmentation will distribute the interleaver outputs over 2, 4, or 8 consecutive frames according to the interleaving length.

3.1.1.3.7 TrCH Multiplexing

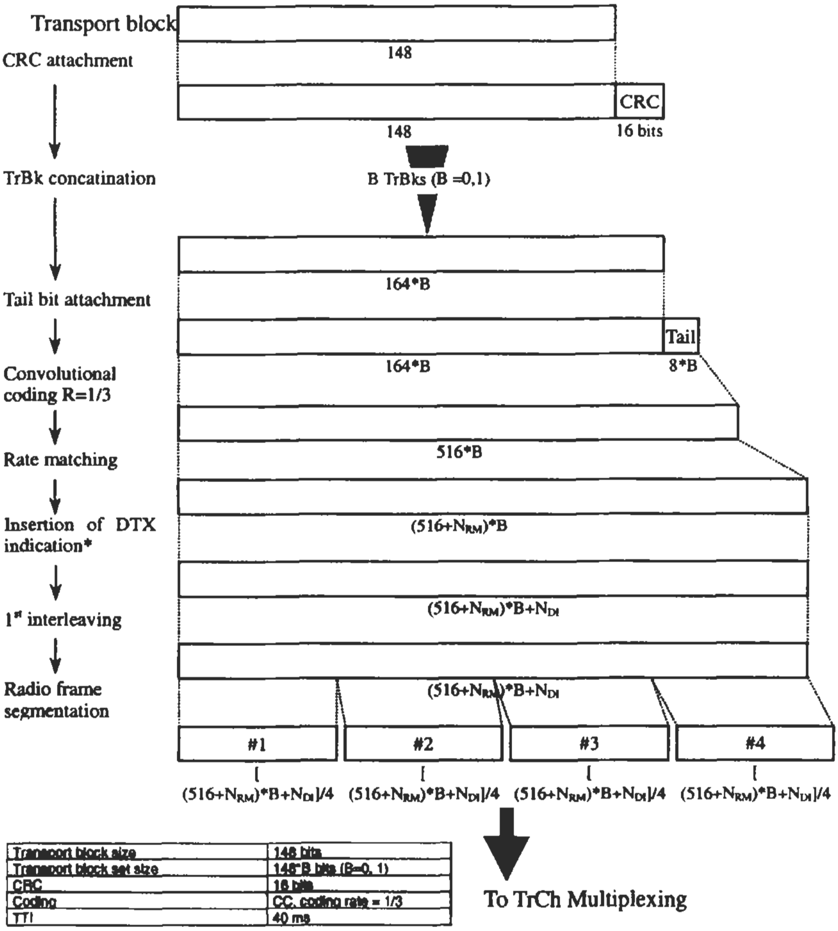
In this stage, several transport channels (each using a 10 ms frame per channel) are serially multiplexed into one coded composite transport channel (CCTrCH). When the total bit rate of a CCTrCH exceeds the capability of one physical channel, several dedicated physical channels (DPCHs) can be transmitted in parallel for one CCTrCH — i.e., a multicode transmission concept. Note that only one control channel (DPCCH) is needed for each CCTrCH connection, and the spreading factor should be identical for all DPCHs under the same CCTrCH.

3.1.1.3.8 Physical Channel Segmentation and Mapping

- Physical channel segmentation: In multicode transmission, physical channel segmentation divides the bits among the different physical channels (PhChs).
- Physical channel mapping: The data from the second interleaver is now mapped to the physical channels (e.g., PCCPCH, DPDCH, PRACH, etc.).

3.1.1.3.9 Channel Coding and Multiplexing Example

In this section, an example is provided to illustrate how the multiplexing and coding work for a downlink channel. In this example, a transport channel with 3.4 kbps data rate (40 ms TTI, 148 bits transport block size) and a 384 kbps packet data (20 ms TTI, 352 bits block size) are multiplexed together and transmitted through several physical channels (see Figure 6-22, Figure 6-23, and Figure 6-24). For the 3.4 kbps transport channel (Figure 6-22), there are 136 bits (3.4 kbps * 40 ms) in one TTI block, which is less than the 148 bits transport block size. Therefore, no segmentation is needed. In the convolutional coding stage, 8 tail bits are added for every 164-bit code block. After the first interleaving, the 40 ms transmission time interval (TTI) is segmented and mapped onto 4 consecutive radio frames.



Transport block size	148 bits
Transport block set size	148*B bits (B=0, 1)
CRC	16 bits
Coding	CC, coding rate = 1/3
TTI	40 ms

* Insertion of DTX indication is used only if the position of the TrCHs in the radio frame is fixed.

Figure 6-22: Channel Coding and Multiplexing Example for 3.4 kbps Data

For the 384 kbps packet data transport channel (Figure 6-23), 24 transport blocks (TrBKs) in a TTI are serially concatenated (X_i = 8448 bits = 352 * 24). The bit stream after concatenation is greater than the maximum code block size Z (which is 5114 for turbo coding), and it is segmented into two blocks ($2 = \lceil 8448/5114 \rceil$). After turbo coding, 12 tail bits are added per segment. Note that the tail bit addition is different from that in the 3.4 kbps TrCh (where a convolutional code is used). In this example, tail bits are added for every 164-bit code block for convolutional coding. For turbo coding, tail bits are added after the entire segment (12672 bits). The 20 ms TTI is then segmented and mapped to 2 radio frames, followed by rate matching and interleaving.

The number of TrChs	1
Transport block size	336 bits
Transport block Set size	336*B bits (B= 0, 1, 2, 4, 8, 12, 16, 20, 24 for TTI=20 ms)
CRC	16 bits
Coding	Turbo coding, coding rate = 1/3
TTI	20 ms

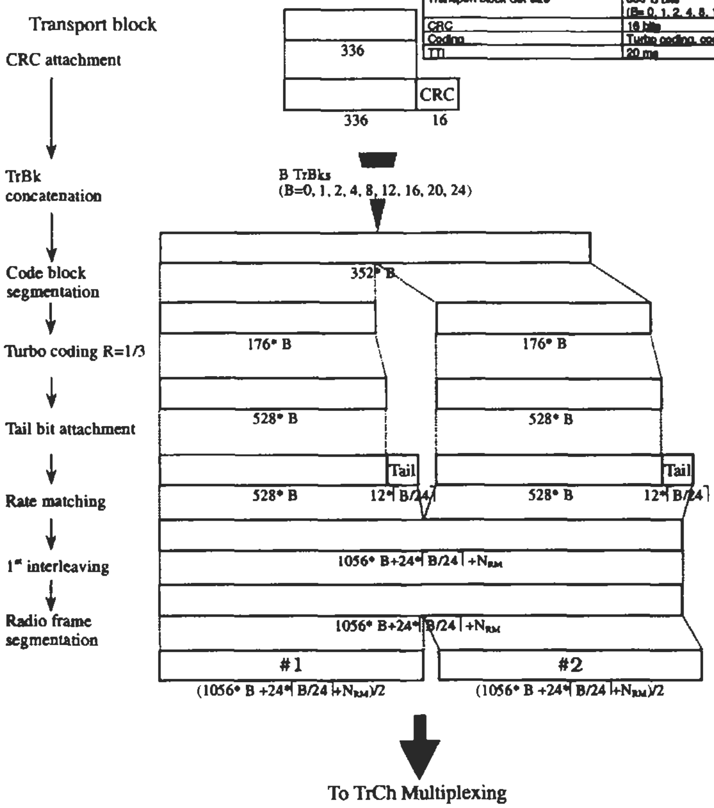


Figure 6-23: Channel Coding and Multiplexing for 384 kbps Packet Data

Before the TrCH multiplexing stage, different transport channels are processed separately. After the radio frame segmentation, every 10 ms radio frame from each TrCH (3.4 kbps TrCH and 384 kbps TrCH) is multiplexed into coded composite transport channels (CCTrCHs). Then, the multiplexed data is divided into *P* segments for *P* physical channels. Followed by the second interleaving, the data is mapped to *P* physical channels (DPCHs) by adding a transport format combination indicator (TFCI), transmission power control (TPC) commands, and pilot information into every slot.

Other multiplexing and channel coding examples can be found in [5].

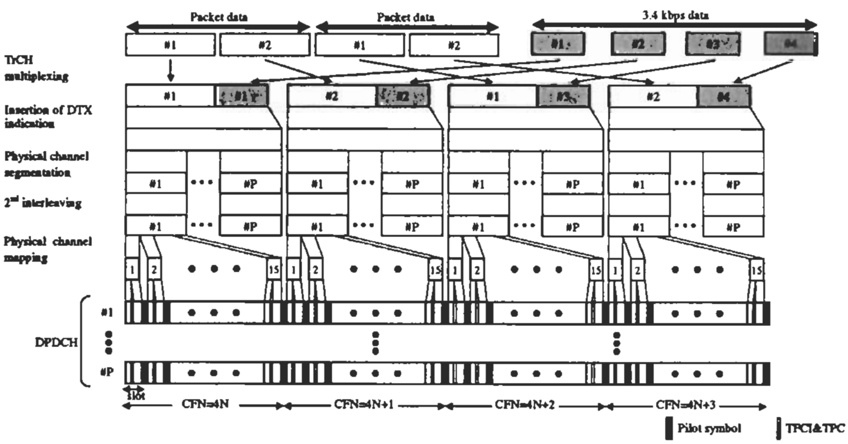


Figure 6-24: Multiplexing Example of 384 kbps and 3.4 kbps Data

3.1.2 Spreading

In the spreading process, the physical channels are first spread to the channel bandwidth (3.84 Mcps) by the channelization code and then scrambled for cell or UE separation (see Figure 6-25). Note that the scrambling codes are used for identifying different mobiles for the UL and different cells for the DL. The scrambling process will not increase the channel bandwidth.

3.1.2.1 Channelization Codes

The channelization codes in WCMDA serve two purposes: the first is to spread the bit rate to the chip rate (3.84 Mcps), and the second is for user identification (for downlink) or the data/control channel separation (for uplink). The channelization codes need to have good orthogonal properties in order to reduce the interference, and different spreading factors can be used for variable rate transmission.

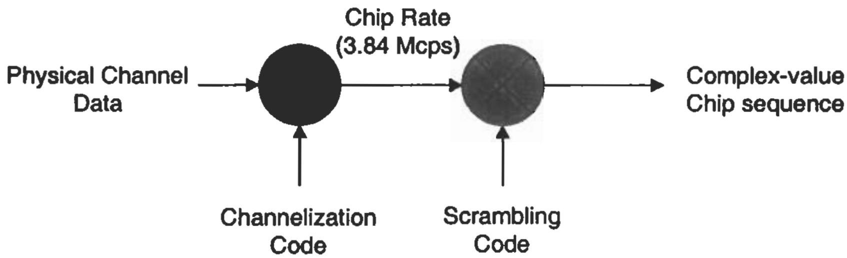


Figure 6-25: Spreading and Scrambling

In WCDMA, Orthogonal Variable Spreading Factors (OVSF) codes are used for channelization codes. The OVSF codes are generated from a single base Walsh-Hadamard matrix (see also Chapter 2, Section 4):

$$\mathbf{H}_{2N} = \begin{bmatrix} \mathbf{H}_N & \mathbf{H}_N \\ \mathbf{H}_N & -\mathbf{H}_N \end{bmatrix} \tag{6-3}$$

The channelization code length is in the form of 2^n, where n is greater than or equal to 2. Note that the length of the code is also equal to the spreading factor (SF = 2^n). The channelization codes in WCDMA can use different length OVSF codes to achieve variable rate transmission. When one OVSF code is selected, the subtree codes will be blocked and won't be usable in the same physical channel. An example is shown in Figure 6-26, where the code $C_{ch,4,1}$ is selected, the spreading codes in the subtree (e.g., $C_{ch,8,2}$, $C_{ch,8,3}$, $C_{ch,16,4}$, ...) are all blocked.

The OVSF codes have good orthogonality properties, but they require perfect synchronization at the symbol level. It is also worth mentioning that the short length channelization code is still orthogonal to the longer length code over the duration of the longer code, if the longer code does not come from a subtree generated through the shorter code. For instance, in the example of Figure 6-26, the code (1,-1) is orthogonal to the code (1,1,1,1) but not to (1,-1,1,-1). The channelization codes can be changed due to the change of service or hand-off process. The downlink channelization codes are managed by the radio network controller (RNC).

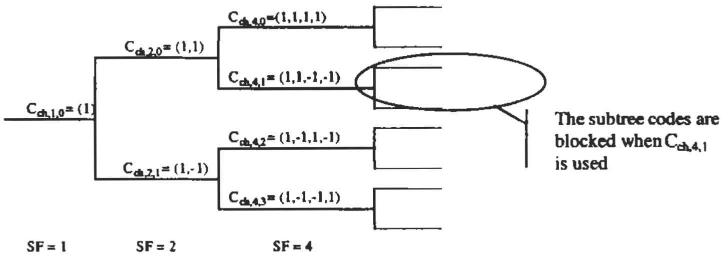


Figure 6-26: Example of OVSF Code

3.1.2.2 Scrambling Codes

The scrambling codes are used to separate different UEs for the UL and different cells (sectors) for the DL. They do not have the spreading effect, and therefore won't affect the transmission bandwidth.

3.1.2.2.1 Uplink Scrambling Codes

For the uplink, each mobile has its own scrambling code, and it can be either a long code or a short code. The short scrambling codes are 256-chip long Extended S(2) code family, and are used when Node B utilizes multi-user detectors or interference cancellation receivers. The long scrambling codes are Gold codes, which are 10 ms in length (38400 chips) to cover a WCDMA frame.

- Long scrambling sequence

  The long scrambling codes are 10 ms frame length. For 3.84 Mcps, they contain 38400 chips. The complex-valued long scrambling sequence $C_{long,n}$ is defined as:

$$C_{long,n}(i) = c_{long,1,n}(i)\left(1 + j(-1)^i c_{long,2,n}\left(2\lfloor i/2 \rfloor\right)\right) \tag{6-4}$$

  where $c_{long,1,n}$ and $c_{long,2,n}$ are generated by the polynomials $X^{25}+X^3+1$ and $X^{25}+X^3+X^2+X+1$ (see Figure 6-27 and [6]).

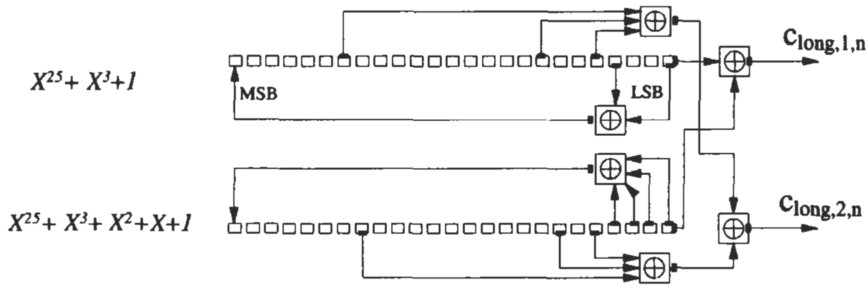


Figure 6-27: Uplink Long Scrambling Sequence Generator

- Short scrambling sequence

  The short scrambling sequences $c_{short,1,n}(i)$ and $c_{short,2,n}(i)$ are defined from a sequence from the family of periodically extended S(2) codes. The complex-valued short scrambling sequence $C_{short,n}$ is defined as:

$$C_{short,n}(i) = c_{short,1,n}(i \bmod 256)\times \left(1 + j(-1)^i c_{short,2,n}\left(2\lfloor (i \bmod 256)/2 \rfloor\right)\right) \tag{6-5}$$

with three generator polynomials $g_0(x) = x^8+x^5+3x^3+x^2+2x+1$, $g_1(x) = x^8+x^7+x^5+x^4+1$, and $g_2(x) = x^8+x^7+x^5+x^4+1$ (see Figure 6-28). The short scrambling sequence is 256 chips in length.

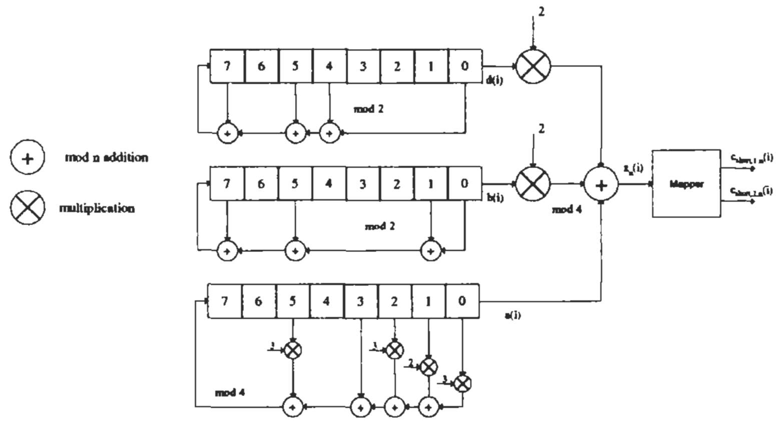


Figure 6-28: Uplink Short Scrambling Sequence Generator

3.1.2.2.2 Downlink Scrambling Codes

For downlink, each Node B or sector has its own scrambling code. There are 2^{18}-1 possible codes, but only 512 codes are used.

More detailed information regarding scrambling codes will be provided in Sections 3.1.2.3 and 3.1.2.4.

3.1.2.3 Uplink Spreading and Modulation

The uplink spreading/scrambling process is illustrated in this section. For the DPCCH/DPDCH spreading, the binary dedicated physical channels are first mapped to +1 or -1 ($0 \rightarrow +1$, $1 \rightarrow -1$) and then spread by the channelization code $C_{d,n}$ or C_c (for DPCCH) (see Figure 6-29). For one DPCCH, up to 6 data channels (DPDCHs) can be transmitted simultaneously. The channelization code of the DPCCH is always $C_{ch,256,0}$. Therefore, we can see that the channelization code for the uplink is to separate the control channel (DPCCH) and data channels (DPDCHs). When only one DPDCH is transmitted, the spreading factor can range from 4 to 256, and it may vary on a frame-by-frame basis. If more than one DPDCH is transmitted, the spreading factor (SF) will be fixed at 4. After the spreading, the signals are weighted by the gain factors β_d (for DPDCH) and β_c (for DPCCH) and then summed for I and Q branches separately. These I & Q signals are then treated as a complex-value chip stream (I+jQ) and scrambled by the scrambling code $S_{dpch,n}$.

The complex-valued uplink scrambling codes are used to identify different UEs. Either long or short scrambling codes can be used for DPCCH/DPDCH scrambling, and the higher layers assign them.

For PRACH or PCPCH, the data and control part are spread by channelization codes C_d or C_c accordingly, and then multiplied by the gain factors. The complex-valued chip sequence is then scrambled by a predetermined scrambling code [6]. Unlike DPCCH/DPDCH, the PRACH and PCPCH do not have simultaneous data channel transmission.

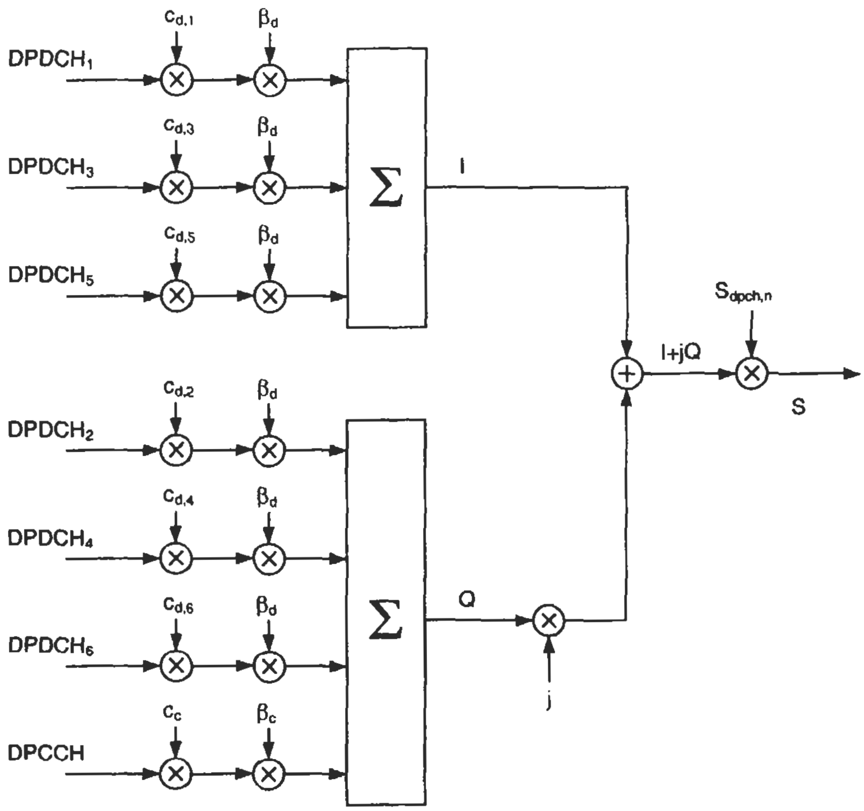


Figure 6-29: Uplink DPCCH and DPDCHs Spreading

The signal is simply pulse-shaped and modulated with the carrier signal. The transmit pulse shaping filter is a root-raised cosine (RRC) filter. The RRC is a type of filter that provides no intersymbol interference (ISI), as it meets the design criterion for zero-ISI (zero crossings at integer multiples of the chip duration; see also Chapter 2, Section 7). The WCDMA RRC is defined with a roll-off factor α =0.22 in the frequency domain. The impulse response of the chip impulse filter $RC_0(t)$ is:

(6-6)

$$RC_0(t) = \frac{\sin\left(\pi \frac{t}{T_C}(1-\alpha)\right) + 4\alpha \frac{t}{T_C} \cos\left(\pi \frac{t}{T_C}(1+\alpha)\right)}{\pi \frac{t}{T_C}\left(1-\left(4\alpha \frac{t}{T_C}\right)^2\right)}$$

3.1.2.4 Downlink Spreading and Modulation

All downlink physical channels (the physical channel outputs from Figure 6-18) except the SCH are first serial-to-parallel mapped into I and Q branches, and spread by the channelization code $C_{ch,SF,m}$. For the initial synchronization process, the UEs need both primary and secondary synchronization codes (P-SCH and S-SCH) to obtain slot and frame synchronization. This has to be achieved before the received data can be sent to a rake receiver for decoding. Hence, the P-SCH and S-SCH do not go through the spreading and scrambling process.

The complex-value spread data, however, is scrambled by the scrambling code $S_{dl,n}$ for cell/sector separation, i.e., each cell/sector will have its own unique scrambling code (see Figure 6-30). Unlike UL scrambling, only long scrambling codes are used in downlink. The code period is 10 ms, and it contains 38400 chips. There are $2^{18} - 1$ scrambling codes that can be generated, but a subset of codes is chosen as the downlink primary and secondary scrambling codes. There are 512 primary scrambling codes ($S_{16*i,dl}$), i = 0,1,...,511. Each primary scrambling code has 15 corresponding secondary scrambling codes $\{S_{16*i+1,dl}, \ldots, S_{16*i+15,dl}\}$, and there is a one-to-one mapping between each primary scrambling code and the 15 secondary scrambling codes.

The 512 primary scrambling codes are further divided into 64 scrambling code groups, each group containing 8 primary codes.

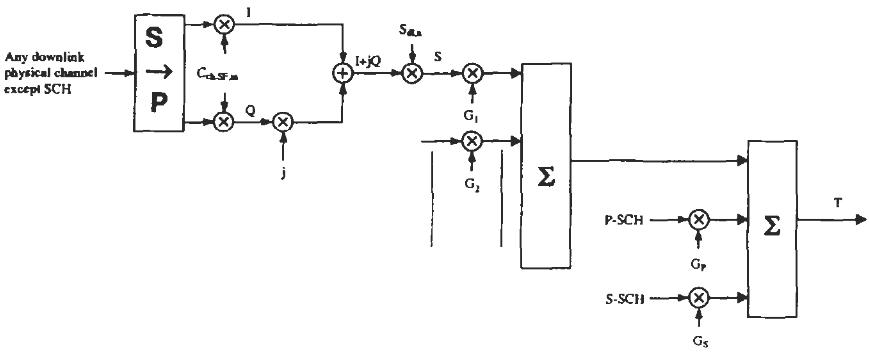


Figure 6-30: Downlink Channels Spreading

The spread signal is simply pulse-shaped (using the filter of (6-6)) and modulated using QPSK modulation for final transmission.

4. HIGH-SPEED DOWNLINK PACKET ACCESS (HSDPA)

In the year 2000, another effort was initiated within the 3GPP to define an evolved system for high-speed data users. This effort was similar to the 1X-EV standardization activity going on within the 3GPP2 (see Chapter 5). In particular, the 3GPP was seeking to enhance the features of the physical downlink shared channel (PDSCH or simply DSCH) so that high-speed data services could be made available to the user without providing circuit-switched connections.

The so-called high-speed DSCH (HS-DSCH), as described in [7], takes advantage of adaptive modulation and coding to enhance data rates to data users in a time-multiplexed manner. The basic modulation and coding is depicted in Figure 6-31. Much like 1X-EV, the number of codes allocated to the HS-DSCH varies and is a function of the offered load. In other words, if a large number of voice users are present, then the number of codes available for the HS-DSCH will most likely become less.

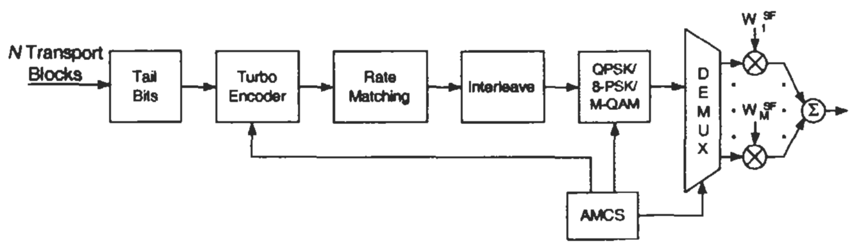


Figure 6-31: HSDPA Downlink Shared Channel Coding and Modulation

An N-channel stop-and-wait fast ARQ method is also described in [7]. This method is almost identical to the fast ARQ method described in the 1XTREME proposal for 1X-EV-DV (see Chapter 5, Section 3.4). As of Summer 2001, the standard was still being defined in the 3GPP with a target date for release of December 2001.

5. CONCLUSIONS

An overview of the WCDMA air interface was provided in this chapter. This along with the previous chapters should provide the reader with an understanding of cellular CDMA systems that exist today. This will also lead to a better understanding when comparing cdma2000 and WCDMA performance (as will be seen in Chapter 7).

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Chapter 7

IS-95, cdma2000, 1X-EV, and WCDMA Performance

Key Topics: Capacity, throughput

Overview: The performance of voice and data service is analyzed with respect to IS-95, cdma2000, and 1X-EV. In particular, the enhancements in voice capacity of cdma2000 over IS-95, due primarily to fast forward power control and coherent reverse link demodulation, are explored. In addition, the achievable data rates in 1X-EV are examined; in particular, the tradeoffs between voice capacity and data throughput for 1X-EV-DV systems. Finally, performance of WCDMA is analyzed to provide a comparison that gives some indications of tradeoffs between IS-95–based systems and WCDMA systems.

1. INTRODUCTION

When IS-95 was originally deployed in the mid-1990s, much of the interest in performance of IS-95 was in the actual voice capacity enhancements seen over existing AMPS, IS-136 and GSM cellular systems. Many of these performance enhancements were predicted by the seminal paper by Gilhousen, et al ([1],[2]) from 1991 which provided a simple capacity analysis of power-controlled CDMA systems versus other cellular systems which existed at that time; the predicted improvement of CDMA systems over AMPS systems was an 18-fold increase in capacity, while over 30-kHz TDMA systems it was 6-fold.

However, such cross-technology comparisons often make assumptions that do not hold up with actual deployments (e.g., perfect power control). However, simulations and theoretical analysis are possible when comparing one CDMA technology with another when both occupy the same bandwidth. This is certainly possible in the case of IS-95, cdma2000, and 1X-EV.

The comparisons become a little trickier when the technologies occupy different bandwidths, such as IS-95 and WCDMA. In this case, an "apples-to-apples" comparison must be made by analyzing the technologies as deployed over identical bandwidths. For instance, three IS-95 carriers may be deployed in the same bandwidth (5 MHz) it takes to deploy one WCDMA carrier. Under these conditions, a comparison between differing-bandwidth technologies may be made.

1.1 Channel Models

There are several channel models that will be referred to in this chapter. Among them are the ITU channel models [3], which were designated as part of the IMT-2000 project for qualifying different candidate technologies as 3G-compliant [3]. These channel models are summarized in Table 7-1, Table 7-2, and Table 7-3. These all describe multipath channels in terms of "taps", where each tap describes the relative delay of arrival of each multipath at the receiver. Moreover, each path power is described relative to the strongest path (note that the strongest path usually arrives earliest, due to having sustained few reflections). These channel models are based on field measurements for indoor, urban, and rural channels.

Another set of channel models that will be examined in this chapter are the models used in the minimum performance specification for IS-95 systems, IS-98 [4]. These models are given in Table 7-4, Table 7-5, and Table 7-6.

Tap	Channel A		Channel B		Doppler spectrum
	Relative delay (ns)	Average power (dB)	Relative delay (ns)	Average power (dB)	
1	0	0	0	0	Flat
2	50	–3.0	100	–3.6	Flat
3	110	–10.0	200	–7.2	Flat
4	170	–18.0	300	–10.8	Flat
5	290	–26.0	500	–18.0	Flat
6	310	–32.0	700	–25.2	Flat

Table 7-1: Indoor Office Test Environment Tapped-delay-line Parameters, Mobile Velocity = 3 km/h

Tap	Channel A		Channel B		Doppler spectrum
	Relative delay (ns)	Average power (dB)	Relative delay (ns)	Average power (dB)	
1	0	0	0	0	Classic
2	110	–9.7	200	–0.9	Classic
3	190	–19.2	800	–4.9	Classic
4	410	–22.8	1 200	–8.0	Classic
5	—	—	2 300	–7.8	Classic
6	—	—	3 700	–23.9	Classic

Table 7-2: Outdoor to Indoor and Pedestrian Test Environment Tapped-delay-line parameters, Mobile Velocity = 3 km/h

Tap	Channel A		Channel B		Doppler spectrum
	Relative delay (ns)	Average power (dB)	Relative delay (ns)	Average power (dB)	
1	0	0.0	0	–2.5	Classic
2	310	–1.0	300	0	Classic
3	710	–9.0	8900	–12.8	Classic
4	1090	–10.0	12900	–10.0	Classic
5	1730	–15.0	17100	–25.2	Classic
6	2510	–20.0	20000	–16.0	Classic

Table 7-3: Vehicular Test Environment, High Antenna, Tapped-delay-line Parameters, Mobile Velocity = 120 km/h

Tap	1-Path		Doppler spectrum
	Relative delay (ns)	Average power (dB)	
1	0	0.0	Classic
2	N/A	N/A	N/A
3	N/A	N/A	N/A

Table 7-4: IS-98 1-Path Tapped-delay-line Parameters, Mobile Velocity = 30 km/h

Tap	2-Path		Doppler spectrum
	Relative delay (ns)	Average power (dB)	
1	0	0.0	Classic
2	2000	0.0	Classic
3	N/A	N/A	N/A

Table 7-5: IS-98 2-Path Tapped-delay-line Parameters, Mobile Velocity = 8 km/h

Tap	3-Path		Doppler spectrum
	Relative delay (ns)	Average power (dB)	
1	0	0.0	Classic
2	2000	0.0	Classic
3	14500	–3.0	Classic

Table 7-6: IS-98 3-Path Tapped-delay-line Parameters, Mobile Velocity = 100 km/h

2. IS-95 AND CDMA2000 VOICE CAPACITY

There are two main enhancements when examining cdma2000 versus IS-95 in term of capacity: fast forward link power control, and coherent reverse link reception. Both of these enhancements generally lead to overall voice capacity enhancements of cdma2000 versus IS-95, but their impacts are different to the different links. It should also be noted that voice capacity enhancements on the forward link should be accompanied by similar capacity enhancements on the reverse link for the performance enhancements to be meaningful, as voice is a symmetric service.

2.1 Forward Link Enhancements

In IS-95, the forward link power control mechanism was based on a slow process using message-based control. The mobile station would be triggered to send a message to the network when its detected frame erasure rate exceeded a threshold. Based on the event of receipt of this message, the network could choose to increase allocated base station transmit power to the mobile station. In cdma2000, however, the mobile could send 800 Hz fast power control commands to the base station and not suffer the delays of Layer 3 messaging.

A set of simulations run for both the IS-95-B and cdma2000 1X forward links actually demonstrates this difference. The simulation parameters are given in Table 7-7.

Parameter	Value
Carrier Frequency	2 GHz
Chip Rate	1.2288 Mcps
Coding Rate	RC3 RC4
Pilot Power	-7.0 dB of total transmitted power
Channel Model	Pedestrian A, 3 km/h Pedestrian B, 3 km/h Vehicular B, 120 km/h 1 path, 30 km/h 2 path, 8 km/h 3 path, 100 km/h
Power Control Rate	IS-95B - Message based cdma2000 — 800 Hz
Power Control Delay	1 PCG
Power Control Step Size	0.5 dB
Power Control Bit Error Rate	4%

Table 7-7: Forward Link Simulation Parameters

The simulation results are provided assuming the mobile is in various degrees of soft handoff, characterized by the "soft handoff imbalance." This is the power differential between the strongest base station and the additional base station the mobile is communicating with during soft handoff. Performance is measured as the mean chip energy for the traffic channel Ec as a fraction of the total transmitted power Ior. Moreover, performance is characterized as changing with *geometry*, which is the ratio of transmitted base station power to the noise floor. It should also be noted that the target frame error rate for both systems was 1%; this is necessary for acceptable voice quality. The simulation results for several channel models with IS-95-B and cdma2000 1X are given below for voice mode (9600 bps).

The effects of fast forward power control are evident when comparing Pedestrian A and Pedestrian B performance at 3 km/hr between IS-95-B and cdma2000. As much as nearly 3 dB of performance enhancement was possible. This benefit practically disappeared for Vehicular B performance models at 120 km/hr and in fact cdma2000 performed slightly worse than IS-95-B; this is due to the worsening of link performance due to inaccurate power control commands when the Doppler bandwidth is large relative to the 800 Hz power control rate. It should be noted that this problem is not insurmountable, as the network can always choose to measure the reliability of power control commands received from the mobile station (based on a velocity estimate) and simply revert to message-based power control.

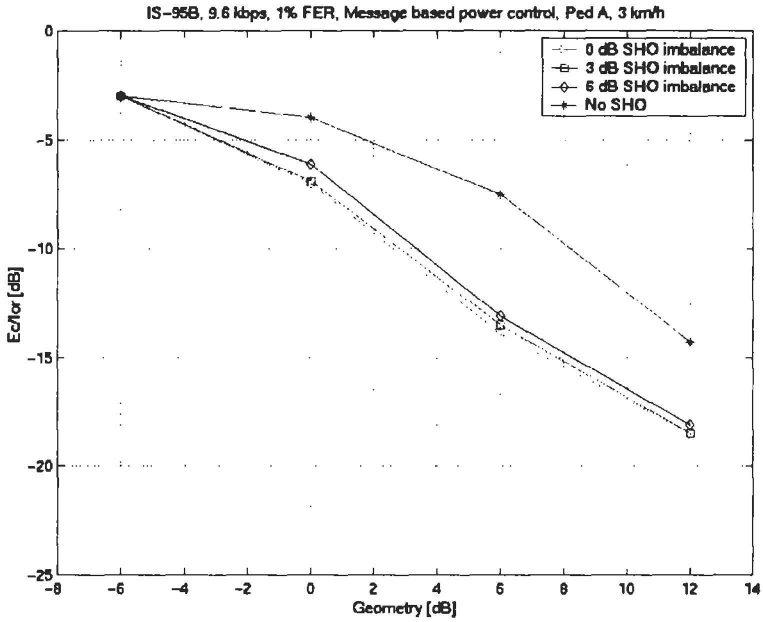

Figure 7-1: Voice Mode, IS-95-B Pedestrian A, 3 km/h

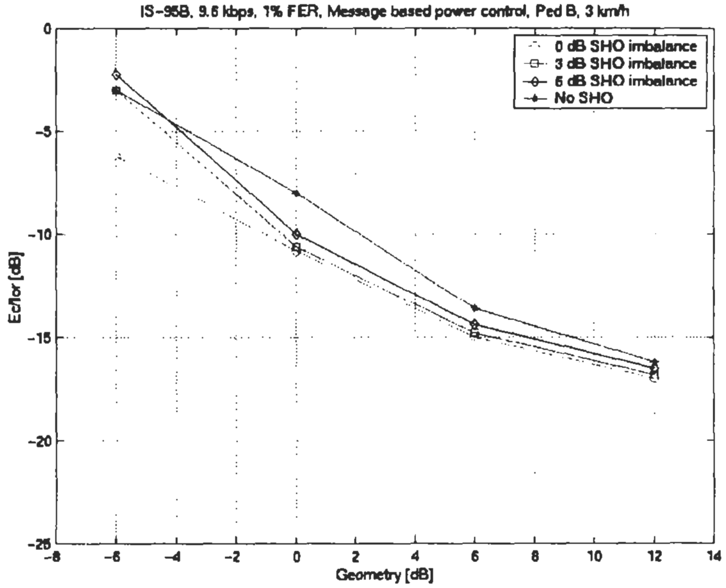

Figure 7- 2: Voice Mode, IS-95-B Pedestrian B, 3 km/h

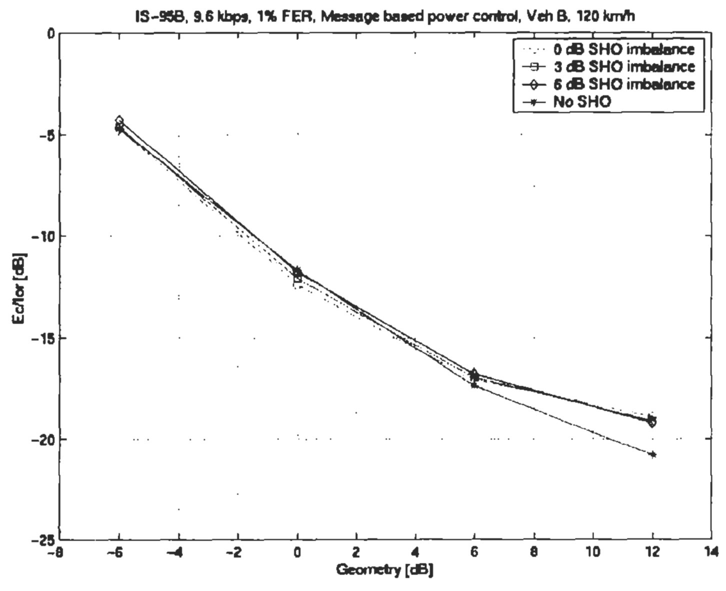

Figure 7- 3: Voice Mode, IS-95-B Vehicular B, 120 km/h

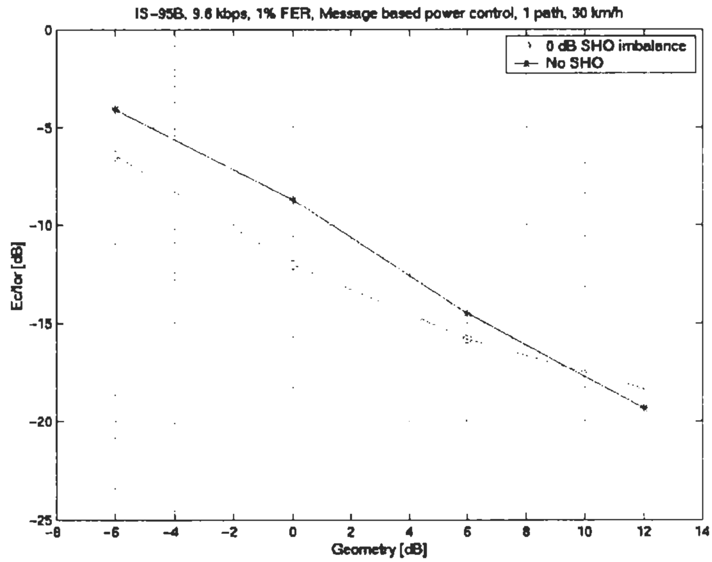

Figure 7- 4: Voice Mode, IS-95-B 1-path, 30 km/h

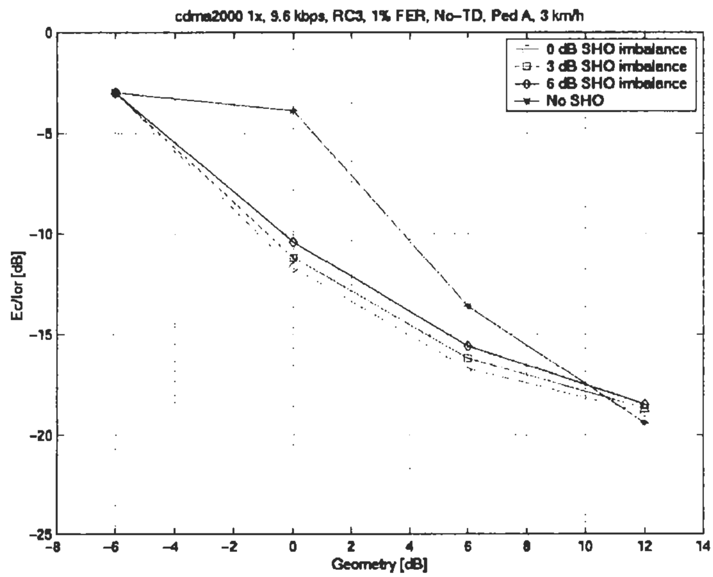

Figure 7- 5: Voice Mode, cdma2000 1X, No Transmit Diversity, Pedestrian A, 3 km/h

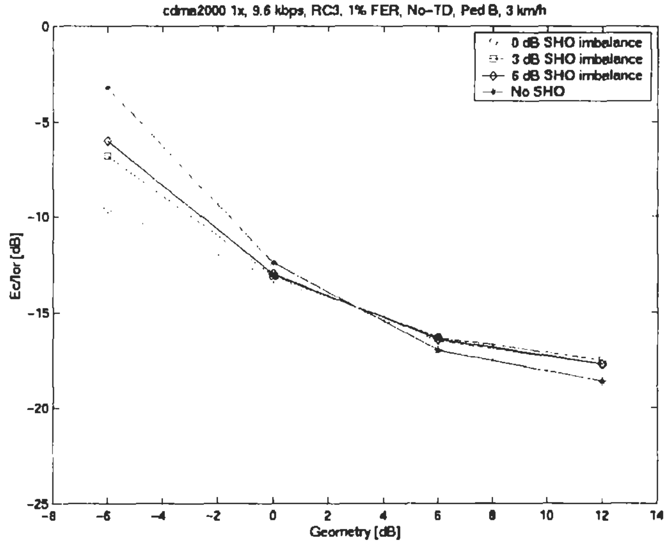

Figure 7- 6: Voice Mode, cdma2000 1X, No Transmit Diversity, Pedestrian B, 3 km/h

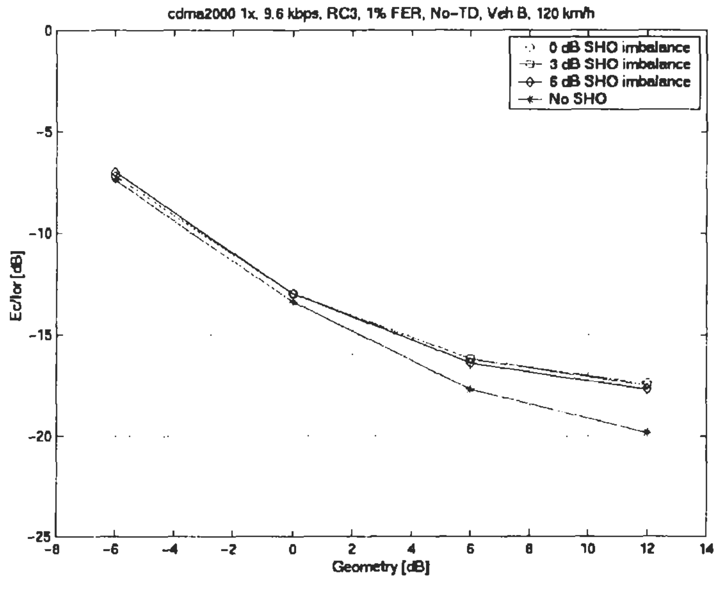

Figure 7- 7: Voice Mode, cdma2000 1X, No Transmit Diversity, Vehicular B, 120 km/h

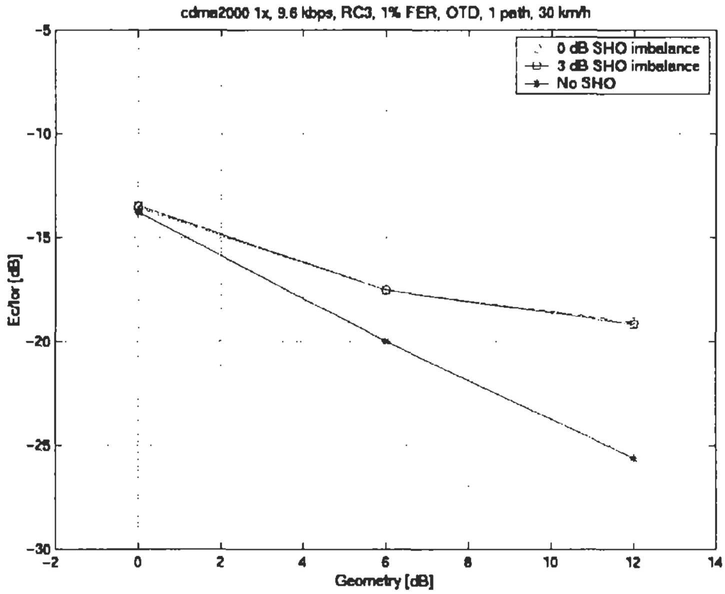

Figure 7- 8: Voice Mode, cdma2000 1X, Transmit Diversity, 1-path, 30 km/h

It should also be noted that the effects of transmit diversity are evident in the one-path, 30 km/hr results: nearly 5 dB of enhancement was evident with the cdma2000 system. One-path channels, which are also *flat fading* channels, are primarily where transmit diversity methods provide the greatest benefits.

2.2 Reverse Link Enhancements

The primary feature that improves reverse link performance when comparing cdma2000 with IS-95 is the addition of the reverse pilot channel (RPICH), which allows for coherent reception at the base station. This is particularly useful when compared to the forward link enhancement of fast forward power control, as the RPICH benefits are seen even at high mobile velocities.

Simulations were also run for the reverse link for both IS-95-B and cdma2000 1X for voice mode, to demonstrate the benefits of the reverse link enhancement to cdma2000. The simulation assumptions are given in Table 7-8.

Parameter	Value
Carrier Frequency	2 GHz
Chip Rate	1.2288 Mcps
Coding Rate	1/4 1/3
Channel Model	Pedestrian A, 3 km/h Pedestrian B, 3 km/h Vehicular B, 120 km/h 1 path, 30 km/h 2 path, 8 km/h 3 path, 100 km/h
Power Control Rate	800 Hz
Power Control Delay	1 PCG
Power Control Step Size	0.5 dB
Power Control BER	4%

Table 7-8: Reverse Link Simulation Parameters

Simulation results are provided in Figure 7-9 and Figure 7-10. It should be noted that although 1% FER is the target operating point for voice services, the results provided demonstrate performance across a range of FERs. The performance metric is E_b/N_t, which is the ratio of information bit energy to total noise. This quantity is based on the reverse fundamental channel (RFCH) for IS-95-B, but for cdma2000, this quantity includes the pilot and fundamental channel powers.

In all comparable channel models, the cdma2000 reverse link outperformed the IS-95-B reverse link. The greatest benefits were in the one-path case (nearly 3 dB), yet there was still a little over .5 dB enhancement in performance at the Vehicular B case.

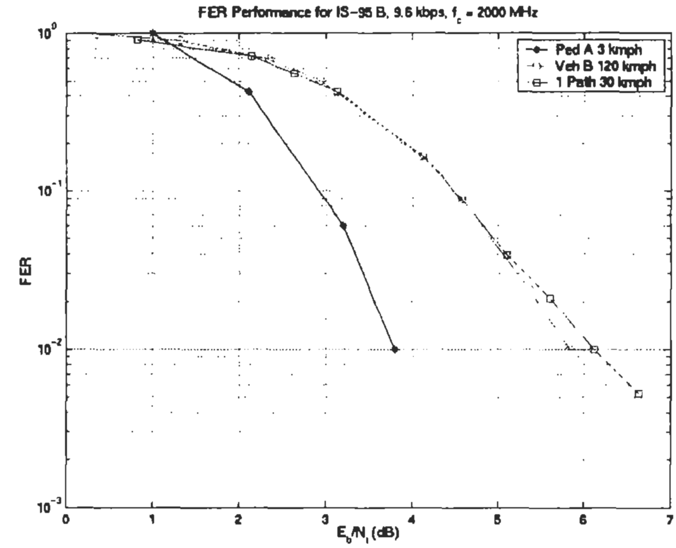

Figure 7-9: IS-95-B Reverse Link Voice Mode Results

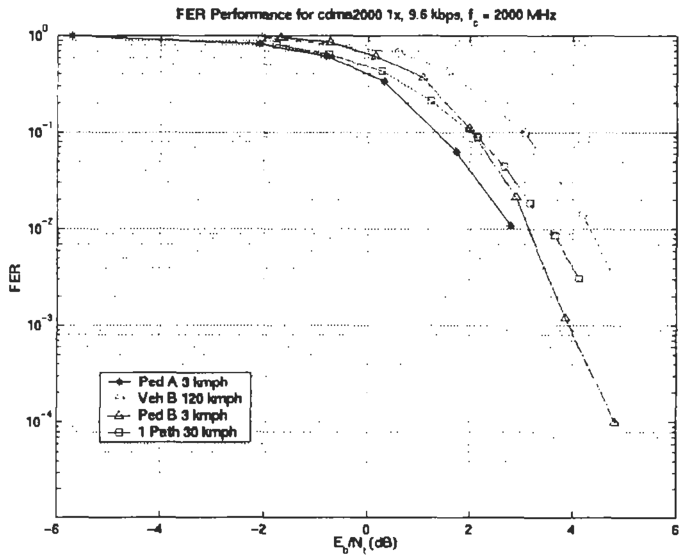

Figure 7-10: cdma2000 Reverse Link Voice Mode Results

2.3 Overall Capacity Comparisons

The common metric taken from either simulations or actual measurements of deployed equipment that characterizes the capacity of telecommunications systems is the ***erlang***. An erlang is a unit of measurement equivalent to a telecommunications channel being fully occupied for the duration of a call. For instance, a system that supports two channels occupied by two users for a quarter of the time supports .5 erlangs of traffic. In a CDMA system, the number of erlangs supported is interference-limited. This is due to the fact that all users occupy the same bandwidth at once, and for certain types of traffic (e.g., voice) a minimum quality of service must be maintained. As a result, as more and more users access a single power-limited CDMA base station, eventually users will have to be blocked from initiating calls as the base station simply cannot maintain the minimum quality of service for new calls until existing users terminate their traffic sessions. This effect can be distinguished from TDMA systems, which (disregarding co-channel interference), are erlang-limited many times by their slot structure (although this limitation can often be addressed through methods such as *slow frequency hopping*).

A simulated system with 19 hexagonal cells and full wraparound was analyzed to determine the potential capacity enhancements of cdma2000 over IS-95-B. Users were randomly distributed in the cells. The *voice activity factor*, which is the percentage of time users occupy a full-rate traffic channel, was assumed to be 50%. The *outage* criterion was that 5% of the mobile users were unable to establish a link at 1% frame error rate.

In the simulations, the cellular network is modeled by locating base stations on centers of the hexagonal grid pattern, with an assumed cell radius of 1 km. The antenna patterns at each cell site are omnidirectional. For vehicular environment, the path loss can be expressed by:

$$L = 40(1 - 4\times 10^{-3} h_b)\log_{10} R - 18\log_{10} h_b + 21\log_{10} f + 80 \tag{7-1}$$

where R is the distance between base station and mobile station (km), f is the carrier frequency of 2000 MHz and h_b is the base station antenna height of 15 m, measured from the average rooftop level. For the pedestrian and one-path environments, the path loss is expressed by:

$$L = 40\log_{10} R + 30\log_{10} f + 49 \tag{7-2}$$

The forward and reverse link capacity comparisons between IS-95-B and cdma2000 are given in Table 7-9.

System	Channel	Erlangs/Cell
IS-95B	ped A, 3 km/h	3.2
IS-95B	ped B, 3 km/h	10.7
IS-95B	veh B, 120 km/h	24.2
IS-95B	1 path, 30 km/h	20.7
cdma2k 1x, no-TD, RC3	ped A, 3 km/h	8.2
cdma2k 1x, no-TD, RC3	ped B, 3 km/h	24.2
cdma2k 1x, no-TD, RC3	veh B, 120 km/h	27.4
cdma2k 1x, OTD, RC3	1 path, 30 km/h	37.7

Table 7-9: Voice Capacity Results

The improvements of cdma2000 over IS-95-B are normally 1.5 to 2 times in terms of erlangs. The only channel conditions where improvements are negligible is the Vehicular B environment at 120 km/hr; this is not surprising, as fast forward power control becomes less effective at high velocities.

3. CDMA2000 DATA CAPACITY

The primary new feature of the cdma2000 1X physical layer (versus IS-95-B) is the potential enhancement of data rates up to 460.8 kbps. This can be analyzed using an approach similar to the system-level simulation used in Section 2.3. However, the traffic model is different, as packet data users do not need to continually occupy over-the-air resources, unlike voice users. A simple packet data model was assumed. The packet data traffic model to be used for system simulations is summarized as follows:

- The user arrival rate of users is modeled as a Poisson variable.
- The packet lengths are exponentially distributed. The mean packet length on the forward link was 12 kilobytes, and the mean packet length on the reverse link was 2.25 kilobytes.

- The users are not blocked when they arrive into the system; instead they are placed into a queue of infinite length.
- Radio link protocol (RLP) reliability is modeled as an increase in the transmission overhead of each packet.

Under these assumptions, the packet data capacity for the cdma2000 1X system is given in Table 7-10. Note that the spectrum efficiency numbers would degrade severely if the users were using circuit-switched connections rather than packet connections.

Service	**Channel Model**	**Spectrum Efficiency (kbps/MHz/cell) reverse/forward**	**Mean Delay (seconds) Reverse/forward**
Packet Data @ 76.8 kbps	Vehicular B	170.0/145.1	0.25/1.25
Packet Data @ 76.8 kbps	Pedestrian B	220.2/125.2	0.25/1.25
Packet Data @ 153.6 kbps	Vehicular B	173.2/ 98.1	0.12/0.65
Packet Data @ 460.8 kbps	Pedestrian B	147.9/ -	0.04/ -

Table 7-10: cdma2000 Packet Data Capacity

To examine the effective capacity of cdma2000 with circuit-switched data users, the system was simulated in a scenario almost identical to the voice capacity simulations in Section 2.3. Some differences are intuitive: for instance, the service is operated at 100% activity factor, and the necessary FER is not 1% but can be higher (e.g., 10%) due to RLP retransmissions. Moreover, symmetric services are not critical for data; therefore the capacity is broken down into reverse link and forward link (see Table 7-11 and Table 7-12).

Service	Channel	Erlangs/Cell
76.8 kbps, 10% FER	ped B, 3 km/h	1.8
76.8 kbps, 10% FER	1 path, 120 km/h	2.2
76.8 kbps, 10% FER	2 path 120 km/h	1.6
76.8 kbps, 10% FER	veh B, 120 km/h	2.4

Table 7-11: cdma2000 Forward Link Circuit-Switched Data Capacity

Service	Channel	Erlangs/Cell
76.8 kbps, 10% FER	ped B, 3 km/h	3.6
76.8 kbps, 10% FER	veh B, 120 km/h	3.3
76.8 kbps, 10% FER	veh B, 120 km/h	3.0
153.6 kbps, 10% FER	veh B, 120 km/h	1.2
153.6 kbps, 10% FER	veh B, 120 km/h	1.0

Table 7-12: cdma2000 Reverse Link Circuit-Switched Data Capacity

The capacity for cdma2000 1X high data rates when deployed in a suitable simulation environment seem to imply that the maximum data rates indicated by the standard are not attainable. This is not necessarily the case; conditions can exist (e.g., small or negligible propagation loss, very light loading of network) that could allow a user to achieve the peak data rates offered by cdma2000.

4. 1X-EV PERFORMANCE

The 1X-EV systems, both 1X-EV-DO and 1X-EV-DV, propose to offer enhancements of packet data throughput over cdma2000 1X systems. In terms of pure packet data performance, this is essentially the case. However, in 1X-EV-DV systems, voice services are identical to cdma2000. Moreover, packet data throughput reduces as the offered voice load increases in 1X-EV-DV (assuming the network resource allocation procedures prioritize voice users at the expense of data users).

Both 1X-EV-DO and 1X-EV-DV performance for packet data users depend on the accuracy of the link adaptation mechanisms. As the received signal quality by the mobile station changes (usually measured in terms of traffic chip-to-noise ratio E_c/N_t), the assigned modulation and coding scheme should change as well. For 1XTREME (a proposal technology for 1X-EV-DV) in a one-path Rayleigh fading channel (3 km/hr, 1960 MHz), the effective throughput (assuming two fast ARQ retransmissions and a single Walsh code used for the forward shared channel) is given in Figure 7-11. Note that there are distinct crossover points where it makes sense to change from one MCS to another — the general trend is that the MCS that maximizes throughput increases with E_c/N_t.

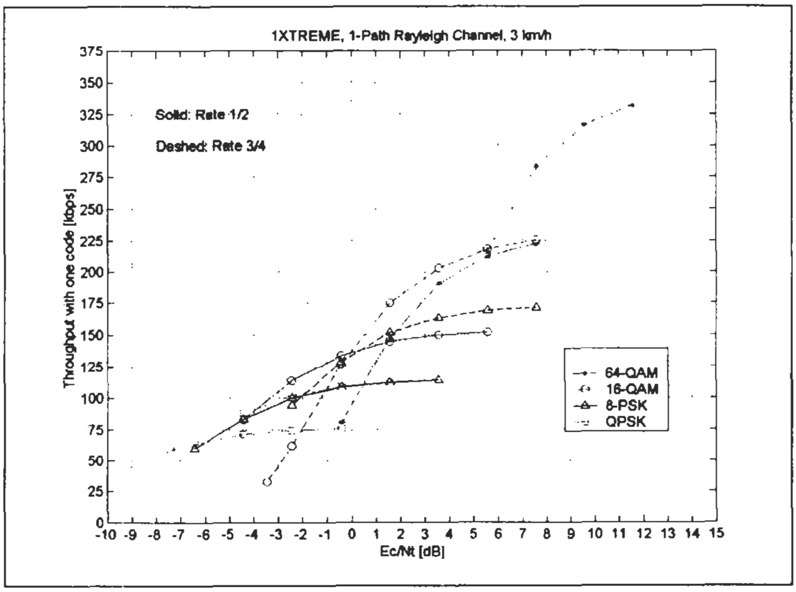

Figure 7-11: 1XTREME MCS Performance

4.1 Data-Only Performance

A simple comparison of the performance of 1X-EV-DO (based on Qualcomm's HDR system) and the 1XTREME candidate technology for 1X-EV-DV is provided in Figure 7-12. The performance numbers are provided as the effective throughput of each system as a function of the offered load (i.e., the amount of data available to deliver on the forward link at the transmitter). Ideally, the effective data throughput should be equivalent to the offered load, but due to frame errors and inaccuracies in link adaptation decisions, the two values usually do not match for any given E_c/N_t.

The 1XTREME system performs well compared to the HDR system under one-path Rayleigh conditions (as in Figure 7-12) at 3 km/hr primarily due to the use of transmit diversity methods available in cdma2000. The 1X-EV-DO system does not make use of forward link transmit diversity; however, if a 1X-EV-DO mobile station uses receive diversity, then the performance of the two systems becomes comparable.

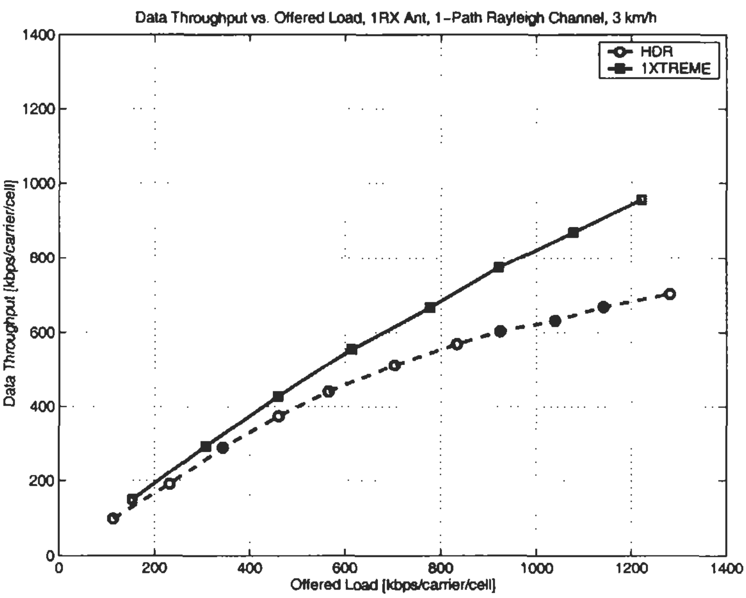

Figure 7-12: 1XTREME and 1X-EV-DO (HDR) Performance

4.2 Mixed Voice and Data Performance

The 1X-EV-DV system does in fact provide for mixed voice and data services on the same carrier. The 1XTREME system provides a means of trading off base station resources between voice users and best-effort data users. As an example [5], the throughput for best-effort data users as a function of offered load was graphed once again for 1XTREME, but for different numbers of voice users served by the cell (see Figure 7-13). The criterion for outage and FER remained the same as in the cdma2000 voice capacity analysis of Section 2.

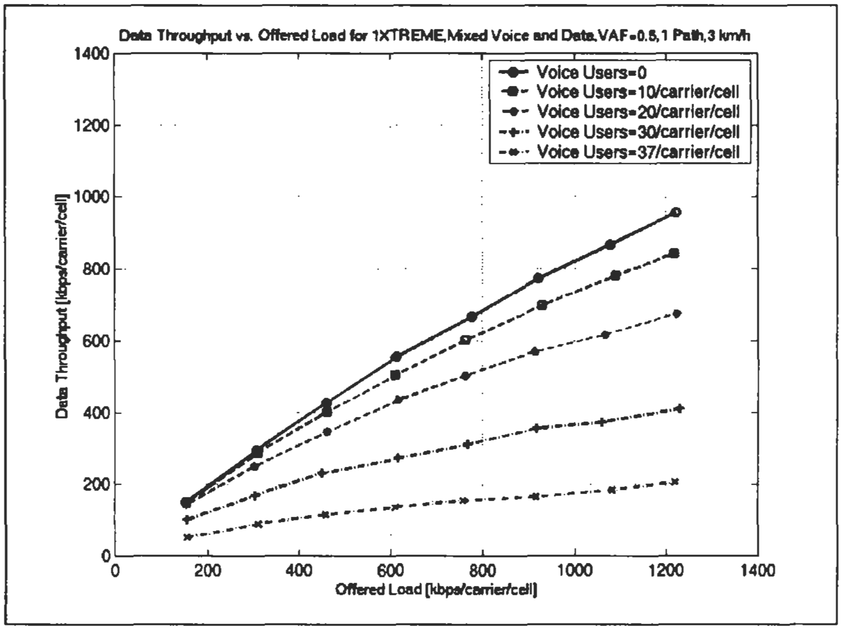

Figure 7-13: 1XTREME Throughput per Voice Users

The maximum data throughput clearly reduces as the number of voice users supported by the cell increases. The reason no more than 37 voice users were considered is that addition of more voice users in this case would have resulted in an unacceptable percentage (more than 5%) of voice users being in outage.

A closer examination of the performance of the 1XTREME system at peak offered load (1220 kbps/carrier/cell) shows how data throughput over the forward shared channel degrades with an increasing number of voice users (see Figure 7-14). However, data throughput is still considerably more than cdma2000 1X even with larger voice loading. For instance, throughput of 700 kbps is still achievable with voice loading of just under 20 users.

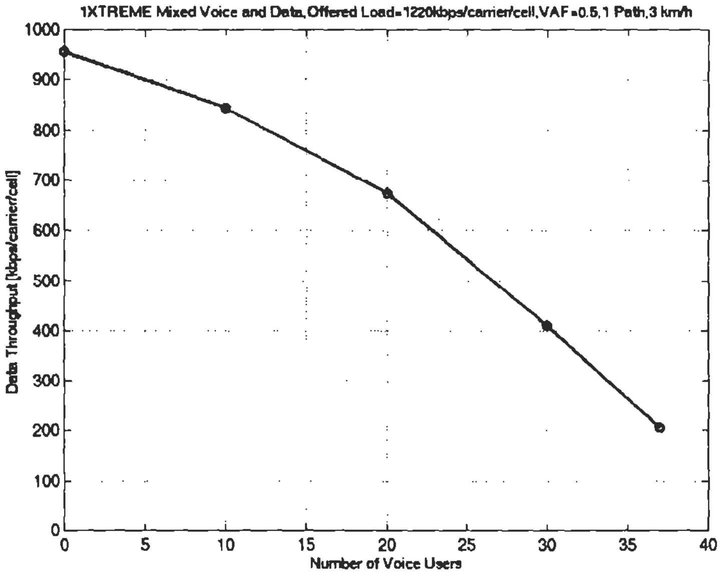

Figure 7-14: 1XTREME Data Throughput Degradation

5. WCDMA PERFORMANCE

When analyzing WCDMA with IS-95/cdma2000 systems in mind, two main reasons for performance differences arise due to the WCDMA physical layer structure:

- Fast power control rate (1500 Hz) as compared to IS-95 (800 Hz)
- Larger chipping rate (3.84 MHz) improves rake receiver performance due to the ability to separate closely spaced multipaths when compared to IS-95 chipping rate (1.2288 MHz)

These two differences become more critical in certain types of channel conditions, e.g., slow-fading or multipath channels.

The comparison between these two types of systems can be better understood by first examining WCDMA performance. For instance, depending on the offered load in a cell or sector, the capacity can be uplink-limited or downlink-limited. A primary factor in situations where the WCDMA system is uplink-limited is when the low mobile station output power becomes the dominant factor in limiting performance. However, as the number of users served by the base station on the downlink increases, then the allocated power per user on the downlink becomes small enough where system performance can become downlink-limited. This trade-off is shown in Figure 7-15, where an increasing number of 144 kbps users are added to the network at different required received E_b/N_o values (1.5 dB for the uplink and 5.5 dB for the downlink). As can be seen, as the number of users rises, the maximum allowable path loss decreases for both the uplink and downlink. This is due to the rise in interference on the uplink due to more users and the reduction in allocated power per user in the downlink. However, the offered load values where the system performance is uplink-limited are low, whereas higher offered load values result in the system being downlink-limited. This is due to the fact that at lower offered load values, the allocated power per user in the downlink is large compared to the maximum mobile transmitter power (in this example, 20 W for the base station and 125 mW for the mobile). However, at higher offered load values, the use of receiver diversity through multiple receive antennas at the base station improves performance on the uplink; similar enhancements are not made on the downlink due to the cost of implementing multiple receive antennas in handsets.

The downlink coverage may be further enhanced by taking advantage of transmit diversity techniques in WCDMA (see Figure 7-16). Similarly, the improvements in uplink coverage due to receiver diversity can be seen in Figure 7-17.

WCDMA performance for basic voice services may be analyzed at the link level in a similar manner to IS-95/cdma2000. For instance, Figure 7-18 shows the relationship between allocated E_c/I_{or} for a voice user versus frame error rate for multiple channel conditions on the downlink.

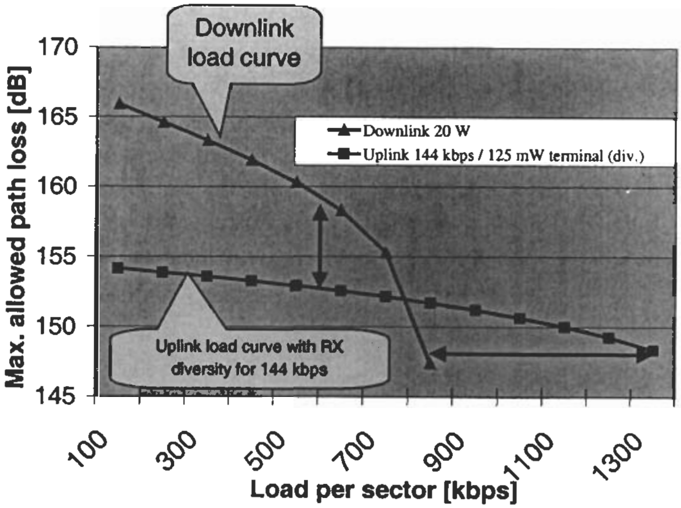

Figure 7-15: Coverage Degradation as a Function of Offered Load

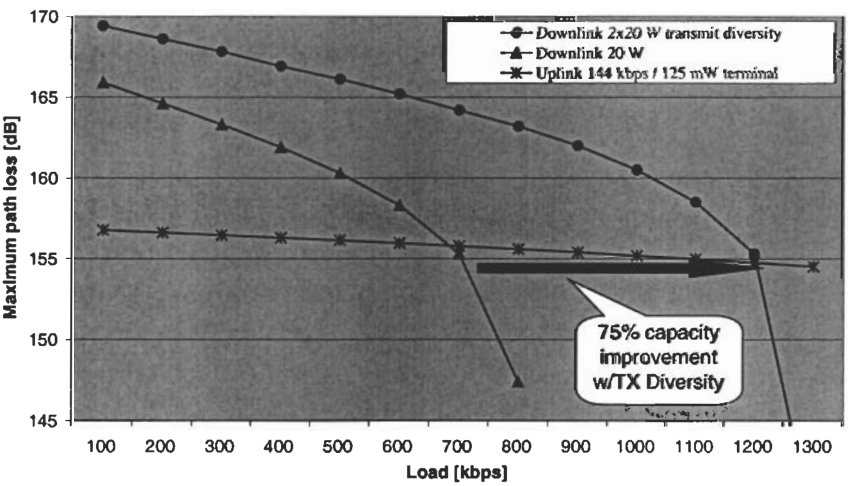

Figure 7-16: Downlink Coverage Enhancement with Transmit Diversity

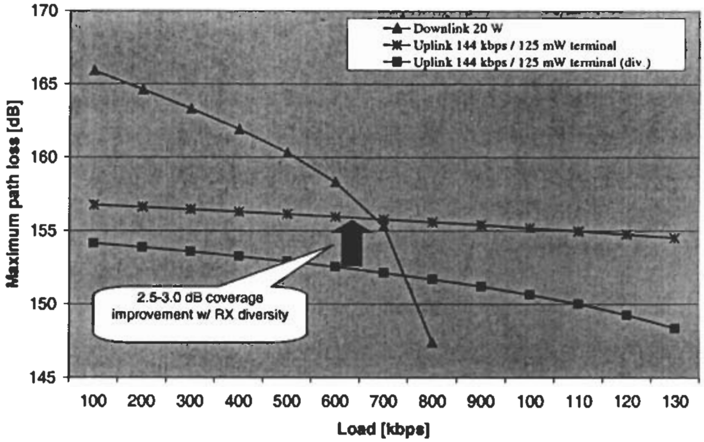

Figure 7-17: Uplink Enhancements Due to Receiver Diversity

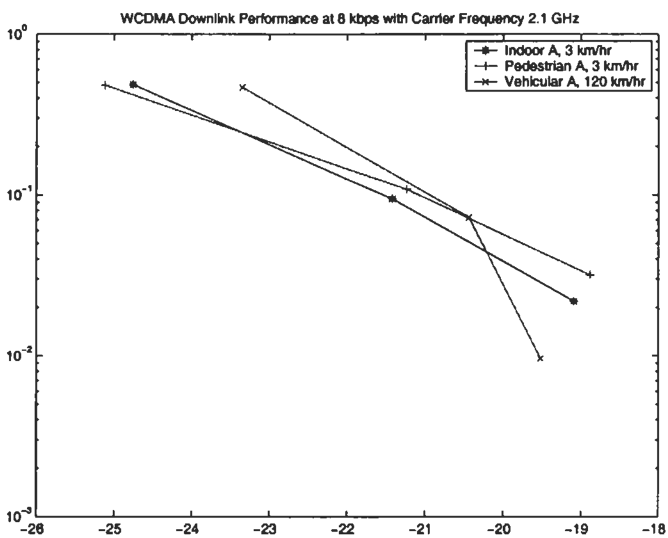

Figure 7-18: Voice Service Performance on WCDMA Downlink

5.1 WCDMA as Compared to IS-95/cdma2000

As mentioned in Section 1, a single WCDMA system cannot directly be compared to an IS-95/cdma2000 system due to the different bandwidths. Typically, three 1.25 MHz IS-95 or cdma2000 carriers can be deployed in the 5 MHz bandwidth required to deploy a single WCDMA carrier. Because the WCDMA uses a chipping rate approximately three times that of IS-95/cdma2000, one might assume that the capacities of the two deployment scenarios would be identical.

However, two differences in the downlinks of these two systems must be taken into account:

1. The common channel overhead, i.e., how much power must be allocated on the downlink to common channels such as the pilot.
2. The variance in base station transmit power actually reduces when there are larger numbers of users supported in the downlink. This results in less power back-off from the transmit power ceiling of a particular base station.

Common channel overhead affects the total coverage area of a base station, as the mobile stations in a cell must be able to demodulate these channels for channel estimation and determination of system parameters to initiate and maintain communications links. For instance, in the IS-98 testability specification for IS-95 systems [4], the three main common channels are defined to have approximate power allocations as specified in Table 7-13 for minimum performance testing of handset demodulation of the paging channel. With reference to this table, the power allocations are given in terms of chip energy per code channel E_c to total base station transmitted power I_{or}.

Table 7-13 ndicates a total common channel power overhead of 25% for IS-95 base stations. This number will most likely remain the same for cdma2000, as these particular code channels are unchanged.

Channel	E_c/I_{or} (dB)	Percentage of Total Power
Pilot	–7	20
Paging	–16	2.5
Sync	–16	2.5

Table 7-13: Power Allocation for IS-95 Common Channels

WCDMA has a different set of common channels, but its power allocation is also a function of differences in performance due to the different chipping rate. In the 3GPP mobile station transmitter and receiver performance specification [7], the mobile station (UE) receiver is qualified under the requirements given in Table 7-14. In this case, the common channel overhead is approximately 16% of the total base station transmitter power. When compared to the 25% number for IS-95/cdma2000 systems, this implies a 12% capacity enhancement of WCDMA over IS-95/cdma2000.

Channel	E_c/I_{or} (dB)	Duty Cycle	Percentage of Total Power
Common Pilot	–10	1	10
Synchronization (both Primary and Secondary)	–12	1/15	0.4
Primary Common Control	–12	14/15	5.9

Table 7-14: Power Allocation for WCDMA Common Channels

The fact that a WCDMA system can support more users due to lower common channel overhead will also lead to smaller variation in base station transmitter power. Due to fluctuations in the necessary transmitter power to each user on the downlink in a CDMA network (due to power control, user movement, changing propagation environment), the base station must ensure that the total transmitted power at any given instant in time is determined in such a way as to provide some headroom against the maximum allowed base station transmit power. A larger number of aggregate users supported by the base station reduces the level of these potential power fluctuations; intuitively, this is because increases in transmit power have a better chance of being balanced out by users who can sustain decreases in transmit power in a larger user pool. This reduction in back-off in WCDMA results in approximately 8% more available base station power; this, along with reduced overhead due to common channels, results in a capacity enhancement of 27% for WCDMA base stations versus IS-95/cdma2000 base stations in a 5 MHz deployment.

6. CONCLUSIONS

The main conclusions that one may take away from the results presented in this chapter are the following:

a. cdma2000 1X provides voice capacity advantages over IS-95-B to the extent that deployment of cdma2000 systems could actually increase voice revenues for operators.
b. cdma2000 does provide some enhancements in data capabilities over IS-95-B. However, 1X-EV systems provide even greater throughput using true connectionless links (i.e., shared channels).
c. 1X-EV-DV systems such as 1XTREME sacrifice data throughput for voice services. However, the degradation in data throughput as the offered voice load increases is graceful.
d. WCDMA has the potential to provide capacity enhancements over cdma2000 when compared with a 5 MHz deployment.

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Chapter 8

Handover in IS-95, cdma2000, 1X-EV, and WCDMA

Key Topics: Handover, handoff, soft handoff, intra-frequency handover, fast cell site selection

Overview: This chapter introduces the concept of handover in IS-95–based CDMA systems and WCDMA. The necessary procedures behind pilot set maintenance are examined, and the extensions of these concepts to WCDMA, cdma2000, and 1X-EV systems are also explored. Moreover, the concept of cell site selection in 1X-EV systems is described, along with its effects on set maintenance.

1. INTRODUCTION

The concepts of handover have already been introduced in earlier chapters. There are three types of handovers in cellular systems:

- Intra-frequency: handovers between base stations using the same carrier frequency, e.g., soft handover in CDMA systems
- Inter-frequency: handovers between base stations using different carrier frequencies, and
- Inter-system: handovers between base stations employing different air interfaces, e.g., handover between a GSM base station and a WCDMA base station.

In this chapter, intra-frequency handovers will be examined.

The concept of CDMA intra-frequency handovers entails the use of *pilot sets*. These are base station pilots (referenced by their PN offsets for IS-95–based systems or by their scrambling codes for WCDMA) that are grouped according to their functionality in the handover process. The different pilot sets are [1]:

- *Active Set* — all pilots for which there are existing dedicated channels for the mobile station; as many as 6 pilots at any given instant in time can belong to this set in an IS-95–based system, and 3 pilots can belong to this set for a WCDMA system.
- *Candidate Set* — pilots that are not in the active set but have been received with sufficient signal strength so that if the mobile can properly demodulate them, the associated dedicated channels are assigned
- *Neighbor Set* — pilots that are neither in the active nor candidate sets, but are in sufficiently close geographical vicinity to the mobile station so that they could enter into soft handover with the mobile
- *Remaining Set* — all other pilots in the system

These pilot sets are unique to each mobile station, and the network controls their membership. However, the network makes decisions to add or drop pilots from these various sets based on measurements from the mobile station. These measurements are based on the *pilot* E_c/I_o, which is the ratio of the pilot chip energy E_c to the total received in-band (over 1.25 or 3.84 MHz) power I_o.

2. IS-95 HANDOVER

The IS-95 system manages handovers through the use of the different pilot sets mentioned in the previous section. These pilot sets are updated by the network through a *Handoff Direction Message* (HDM) [2], a Layer 3 message sent by the base station to the mobile station. The mobile confirms successful reception of the HDM through the *Handoff Completion Message* (HCM). The mobile station reports pilot E_c/I_o values through the *Pilot Strength Measurement Message* (PSMM).

The base station communicates four sets of system parameters to all mobiles in its area of coverage. These parameters are usually initialized using the *System Parameters Message* or the *Extended System Parameters Message* over the forward paging channel F-PCH:

a. The static handover parameters T_ADD, T_DROP, T_COMP, and T_TDROP
b. The dynamic handover parameters SOFT_SLOPE and ADD_INTERCEPT

c. The PN offsets of the initial neighbor set
d. The search windows for the pilot sets (SRCH_WIN_A, SRCH_WIN_N, SRCH_WIN_R)

These parameters may be updated during the call using the HDM (or *Extended Handoff Direction Message* [*EHDM*] or *General Handoff Direction Message* [*GHDM*]).

2.1 Static Handover Parameters (T_ADD, T_DROP, T_COMP, T_TDROP)

The mobile station assists in the maintenance of the various pilot sets using the handover parameters. Based on these parameters, the mobile may be triggered to send PSMMs so that the base station may update the pilot sets for that mobile.

T_ADD is the threshold E_c/I_o value (in dB) at which a pilot may be added to the active or candidate sets. When a pilot that is not already in the active or candidate sets exceeds T_ADD, the mobile is triggered to send a PSMM.

T_DROP is the threshold that an active set pilot may not drop below. When an active set pilot drops below this threshold, the mobile starts the T_TDROP timer. The T_TDROP timer, which is a time duration in seconds, is the amount of time for which the pilot should remain below the T_DROP threshold before the mobile is triggered to send a PSMM. This event is used by the network to drop the pilot from the active or candidate sets. An example of T_ADD and T_DROP events is shown in Figure 8-1.

When a pilot that is in the candidate set exceeds a pilot that is in the active set by T_COMP, the mobile is also triggered to send a PSMM. This could result in a pilot being added to the active set, and possibly a pilot being dropped from the active set (due to a maximum of 6 pilots being allowed to be in the active set). This event is depicted in Figure 8-2.

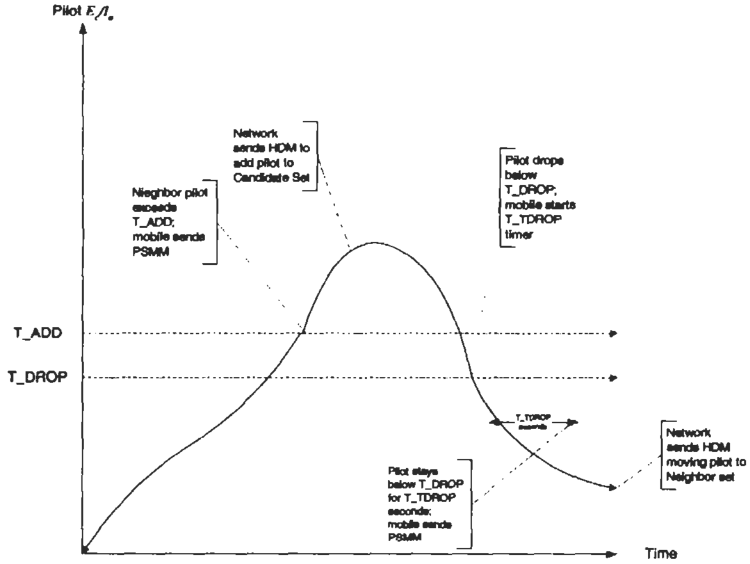

Figure 8-1: Adding and Dropping Pilots from the Candidate Set

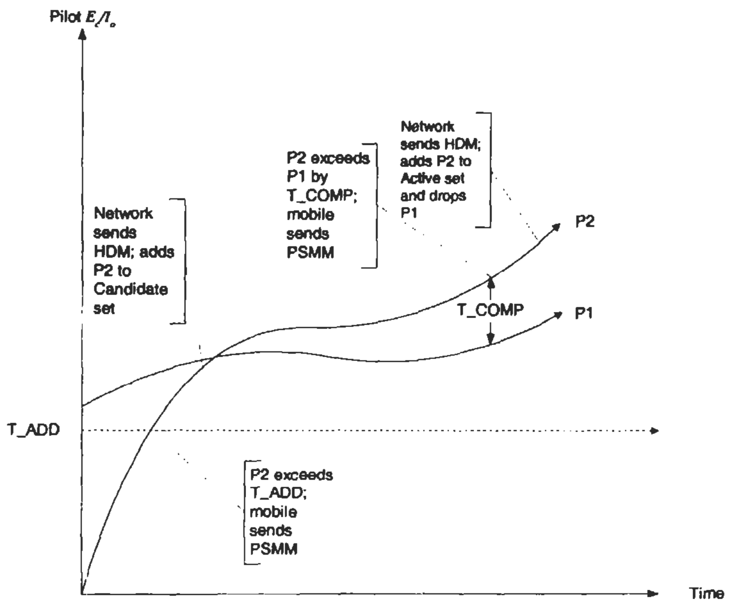

Figure 8-2: Adding and Dropping Pilots from Active Set based on T_COMP

2.2 Dynamic Handover Parameters (SOFT_SLOPE and ADD_INTERCEPT)

The dynamic handover parameters are a means of providing adjustable handover thresholds to the mobile station during operation. Use of these two parameters, SOFT_SLOPE and ADD_INTERCEPT (both in dB units), provides a means of maintaining handover sets based on the relative strength of all the pilots that the mobile is receiving. For instance, although T_ADD is still applicable for adding a pilot to the candidate set, moving a pilot j to the active set can only occur when the measured pilot strength satisfies the inequality in (8-1).

$$10\log_{10}\left(\frac{E_c}{I_0}\right)_j > \frac{\text{SOFT\_SLOPE}}{8}\left(10\log_{10}\sum_{i\in \text{ActiveSet}}\left(\frac{E_c}{I_0}\right)_i\right) + \frac{\text{ADD\_INTERCEPT}}{2} \tag{8-1}$$

Under these conditions, the mobile is triggered to send a PSMM. The network can add pilots to the active set based on this type of criteria.

The use of dynamic thresholds is easily disabled by the network by setting SOFT_SLOPE to 0 dB.

2.3 Search Windows

The mobile station receiver uses a *searcher*, which is essentially a rake receiver finger dedicated specifically to the detection and measurement of pilot multipath signals. During a phone call, the searcher is constantly examining and measuring signals to find usable multipaths. The search process during a phone call is not blind; the mobile uses predefined search windows. These are durations (specified in chips) over which the mobile shall search for multipaths around the earliest arriving multipath that the mobile is currently demodulating for data.

SRCH_WIN_A is the search window for the active and candidate sets. It is sufficiently small so that the mobile can capture all usable multipaths that arrive closely in time to the earliest arriving multipath; signals with sufficiently large signal strength usually arrive in this window.

SRCH_WIN_N and SRCH_WIN_R are the search windows for the neighbor and remaining sets. These windows are usually much larger than SRCH_WIN_A, with SRCH_WIN_R usually being even larger than SRCH_WIN_N.

2.4 Impact of Soft Handover on Power Control

The mobile station can receive reverse power control commands from multiple base stations. Each of these base stations could simultaneously send identical power control commands, or could send conflicting commands based on the signal strength each individual base station sees from the mobile.

As a result, IS-95 systems employ the "or of downs" rule, which essentially states that if a single base station for which the mobile had assigned a finger issues a "down" power control command, the mobile must step its transmit power down. This rule implies that maximal ratio combining for power control commands from *different* base stations is not possible (however, it is possible for power control commands from multipaths from the same base station).

2.5 Inter-System and Inter-Frequency Handover

Inter-system and inter-frequency handovers are possible in IS-95 systems. Inter-system handovers are mainly for handover between IS-95 and AMPS systems. This is to account for geographical areas where no IS-95 carrier is available with sufficient signal strength, but a sufficiently strong analog carrier is present. Inter-frequency handovers were introduced in IS-95-B to allow for distributing load across IS-95 systems deployed on different carrier frequencies.

The search mechanisms involve periodically leaving the serving frequency to search the candidate frequency. The candidate frequency signal strength measurement is either the pilot E_c/I_o for an IS-95 carrier or the received signal strength (the in-band received energy) for an AMPS carrier. The mobile station can make its measurements and report them as part of a one-time event or periodically. If the measurement is periodic, the mobile station has to measure the other carrier during the measurement period and report the measurement within a pre-specified time.

During the intervals when the mobile station measures candidate frequencies or systems, it usually terminates communication with the serving base station. This is due to the fact that most mobile stations do not have sufficient complexity to receive or transmit to more than one carrier frequency at a time. If these measurement intervals occur too frequently as part of the handover process, severe degradation can occur in the user traffic. This is due to the fact that reverse link and forward link frames will be missing from user traffic during the measurement intervals, and these missing frames are not guaranteed to be retransmitted (which is the case for voice applications).

3. CDMA2000 HANDOVER

cdma2000 handover manages handover lists identically to IS-95. There are some differences in power control mechanisms:

a. The network can indicate to the mobile whether reverse power control commands received from base stations in the active set are identical to one another (the PWR_COMB_IND indication). This allows the mobile to maximally ratio combine these commands.
b. The mobile station may use the nominally 800 Hz forward power control mechanism built into the reverse link to power control two code channels (by reducing the rate to 400 Hz): the forward fundamental channel (FFCH) and forward supplemental channel (FSCH).

4. 1X-EV

1X-EV systems employ a different approach to handover set maintenance. This is due to the fact that best-effort data services are not in soft handover in the conventional sense; rather, multiple base stations may be receiving the mobile station signal, but only one base station is transmitting forward traffic to the mobile.

4.1 1X-EV-DO

The concepts of active, candidate, neighbor and remaining sets remain the same for 1X-EV-DO. However, some subtle nomenclature changes were made:

- Use of dynamic thresholds are enabled or disabled by setting the system parameter DynamicThresholds ('1' enables, '0' disables).

- The cardinality of the active, candidate, and neighbor sets is determined by the system parameters $N_{RUPActive}$, $N_{RUPCandidate}$, and $N_{RUPNeighbor}$, respectively.
- PilotAdd is used instead of T_ADD. PilotDropTimer is used instead of T_TDROP.
- SearchWindowActive is used instead of SRCH_WIN_A.
- PilotCompare is used instead of T_COMP.

In addition, new features were added for handover set maintenance:

- A maximum age of a pilot in the neighbor set, NeighborMaxAge, is monitored for each neighbor pilot. The expiration of the age timer is only one criterion for removing a pilot from the neighbor list. If the cardinality of the neighbor list has not been exceeded, then "older" pilots may remain in this set.
- Each neighbor pilot may have its own unique search window size.

4.2 1X-EV-DV

The 1X-EV-DV candidate proposal, 1XTREME, proposes the use of a new handover set, the ***eligible*** set. Although maintenance of the active, candidate, neighbor, and remaining sets is identical to cdma2000, the eligible set is a set of pilots drawn from the active set.

When a message such as an HDM (in the case of 1XTREME, this is done through the *Universal Handoff Direction Message* [UHDM]) is sent to a mobile to update the handover lists, the network can also use this mechanism to update the eligible set list by indicating which pilots from the active set will also have associated forward dedicated pointer channels (FDPTRCHs) for the mobile station.

A new handover parameter, S_COMP, is also necessary for the mobile station. This parameter is used for the mobile station to select a best serving base station or sector (whose pilot PN offset is designated as SERVING_PN). The SERVING_PN may be mapped into a transmit sector indication (TSI) and sent to the network. An example of this process is depicted in Figure 8-3. Note the fact that one pilot exceeding another by S_COMP does not actually trigger a PSMM; this is due to the fact that this mechanism is designed to minimize the delays associated with layer 3 processing. Note also that the mobile only begins to monitor the FDPTRCH associated with a newly selected pilot 10 ms after initializing transmission of the new TSI.

4.2.1 Soft Handover Impacts

Base stations in the eligible set are all receiving data on the reverse link from the mobile station. However, only one base station at any given instant in time may be transmitting user traffic over the forward shared channel (FSHCH) to the mobile station. Nevertheless, when the pilots in the eligible set are assigned to a mobile station, each base station in the eligible set allocates resources for FDPTRCHs. This is particularly important to implement reverse power control, as multiple base stations may need to send power control commands to the mobile station to control reverse link interference.

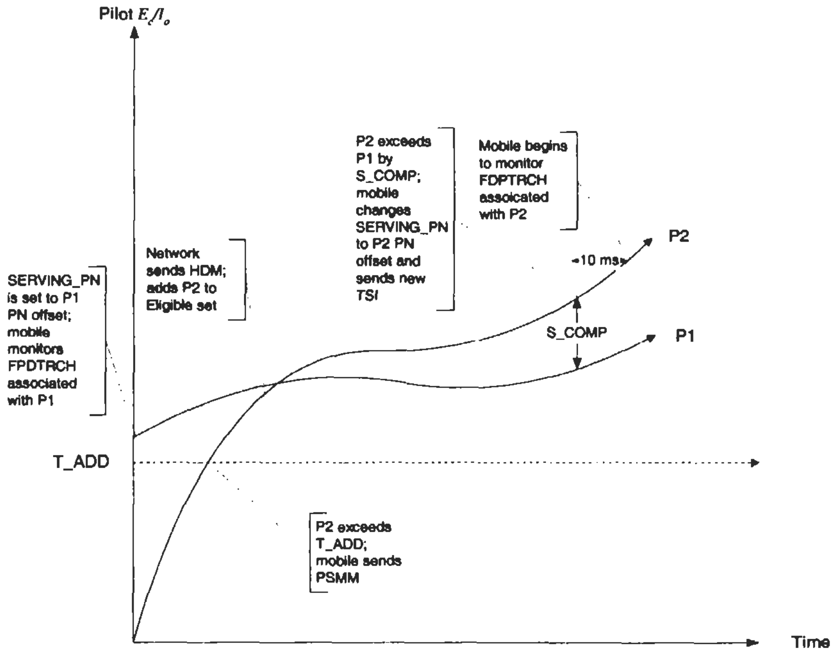

Figure 8-3: Selecting a Pilot Based on S_COMP

Each FDPTRCH could conceivably carry a pointer field (PTR) that is necessary for FSHCH assignments. Yet the mobile station can only be assigned to the FSHCH associated with the pilot it has selected. Therefore, the "unselected" pilots can just transmit the power control commands on the FDPTRCH rather than the entire payload.

4.2.2 Simultaneous Voice and Data

1X-EV-DV systems must be able to support simultaneous voice and data service for a single user. This results in somewhat different issues for handover procedures than those that would arise for a 1X-EV-DO system. As part of the prioritization of voice service at the expense of data services, a service provider may want to allocate fewer network resources for the best-effort data service to reserve more for the voice service on an individual basis. This could result in a much smaller number of eligible set base stations than active set base stations for a simultaneous voice/data user.

Recall that the active set definition is well defined for the voice service in 1XTREME, and is identical to cdma2000 voice services. However, the concept of an eligible set is not really applicable to a voice service, as voice services are in true soft handover on both the forward and reverse links. On the other hand, the concept of an active set is applicable to a data service in 1XTREME from a different level of abstraction — eligible set base stations may be derived as a subset of the active set, but the mobile station only receives forward traffic from one eligible set base station at a time.

This discrepancy may be resolved if the eligible set is assigned with identical pilots to the active set. Then the voice service and data service are associated with the same pilots, and almost all procedures with respect to handover set maintenance and power control become identical to cdma2000, except for one glaring exception: the mobile must still set a TSI (based on S_COMP) to choose a best serving cell or sector from which the mobile station may receive forward traffic on the FSHCH for the data service. The voice service uses an identical cdma2000 fundamental channel and can follow identical cdma2000 power control procedures (including "or of downs" and the use of PWR_COMB_IND).

The power control subchannel in cdma2000 may appear either on the forward fundamental channel (FFCH) or forward dedicated control channel (FDCCH) based on a layer 3 indication (FPC_PRI_CHAN). For simultaneous voice/data, this same indicator can be used to let the mobile know whether the power control subchannel appears on the FDPTRCH or FFCH (in 1XTREME simultaneous voice/data the power control subchannel cannot appear on a FDCCH). The location of this power control subchannel actually implies power control procedures for the mobile station. For instance, if the power control subchannel appears on the FFCH, the mobile station is triggered to fall back to cdma2000 power control procedures for the active set. This entails:

- The "or of downs" rule applies as in cdma2000 to active set base stations.
- Forward power control commands generated by the mobile station are generated to control the FFCH.

However, if the power control subchannel appears on the FDPTRCH, then the mobile station must follow procedures according to the *eligible set*, implying

- The "or of downs" rule only applies to eligible set base stations. Active set base stations may be ignored unless they are also in the eligible set.
- Forward power control commands generated by the mobile station are generated to control only the FDPTRCH associated with the SERVING_PN.

Under these conditions, the voice service could potentially suffer due to the prioritization of power control mechanisms to the data service. Therefore, the preferred mode of operation is normally making the eligible set identical to the active set, thus eliminating the need to prioritize to one service or the other.

5. WCDMA HANDOVER

The handover mechanisms supported in WCDMA are very similar to those supported in IS-95. A mobile station measures E_c/I_o for the common pilot channel (CPICH) supported by a particular base station and reports it using the *Measurement Report* message. The base station in turn can update the active set using the *Active Set Update* message. Pilots in WCDMA are identified in the messaging through their primary scrambling code index (of which there are 512 possible values); this contrasts with IS-95's method of identifying pilots through PN offsets (although there are also 512 possible PN offset values).

An example of the procedure for active set updating in WCDMA is given in Figure 8-4. Three events (as defined in [3]) are depicted:

- Event 1A: A primary CPICH enters reporting range
- Event 1B: A primary CPICH leaves the reporting range

- Event 1C: A non-active primary CPICH is received with greater signal strength than an active primary CPICH

Each of these events occurs with respect to a network-defined reporting range, which may also be adjusted by a *hysteresis* parameter. In addition, a removal event (such as Event 1B) may be enhanced with a drop timer.

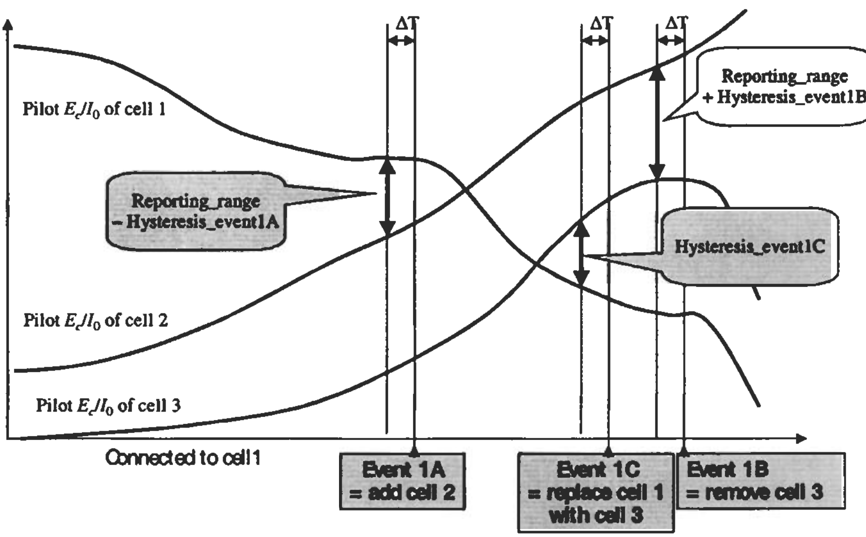

Figure 8-4: WCDMA Handover Events

5.1 Inter-System and Inter-Frequency Handovers

Inter-system and inter-frequency handovers are extremely important in the deployment of WCDMA networks. Inter-system handovers are handovers between WCDMA and another technology such as GSM or EDGE. Inter-frequency handovers occur between WCDMA systems that are deployed using different carrier frequencies. Inter-system handovers are useful in the gradual deployment of WCDMA networks, as these systems can be brought up incrementally with legacy systems such as GSM supporting the remaining users. Dual-mode (or even multimode) handsets will provide seamless roaming between legacy systems and WCDMA networks, which will be necessary if there exist large areas of coverage only served by legacy systems. Inter-frequency handovers, on the other hand, are useful for providing congestion relief in highly loaded areas. Presumably, users may be accommodated on two or more frequencies, thus balancing the load across multiple WCDMA carriers.

Both of these types of handovers are facilitated by use of the *compressed mode* of operation (see Chapter 6), wherein a mobile and base station transmit and receive over the serving frequency for only part of the 10 ms frame duration. This process is initiated with a *Measurement Control* message [4] sent to the mobile station, providing searching parameters to the mobile station. The mobile station will then proceed to measure signal quality on the other frequency or system. The signal quality measurement can be one of the following [5]:

a. Received signal code power (RSCP): the power per chip received on a common pilot channel in a WCDMA system.
b. Received signal strength indication (RSSI): the received power within the system bandwidth. This measure is applicable to either a GSM or WCDMA system.
c. E_c/I_o: the chip-to-noise energy ratio for a WCDMA common pilot channel. This value is identical to RSCP/RSSI.

As an example, let us assume a handover scenario where the mobile station has been ordered to operate in compressed mode every third frame to make measurements of a GSM carrier. In this case, the mobile is on the serving frequency for 7 of the 15 slots in a frame, and uses the remaining 8 for handover measurements on the GSM frequency. If the mobile can make 6 RSSI measurements during this time and the mobile station is required to measure 10 GSM neighbors with 4 RSSI measurement per neighbor, then the mobile will require 7 compressed mode frames to perform this operation (40 total measurements, with 6 measurements per frame). Since every third frame is allocated for measurements, the process takes 210 ms. If we assume that the time it takes for the base station to order the mobile to move to one of the GSM carriers after receiving measurements from the mobile station is bounded by some value that is a function of the network, say 1.5 seconds, then the entire process takes at least 1.71 seconds.

6. CONCLUSIONS

The handover mechanisms for IS-95 have generally carried over to cdma2000 with small changes. Moreover, the handover mechanisms in

WCDMA have leveraged considerably from IS-95–based systems. However, the handover mechanisms have had to undergo a substantial change for 1X-EV systems, as these systems do not employ symmetric soft handover. Similar considerations will also have to be taken into account for WCDMA evolution modes such as HSDPA. As 1X-EV and HSDPA systems are deployed in the near future, it will be interesting to observe how accurate the mobility mechanisms in these systems are with respect to ensuring seamless handovers between cells.

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Appendix

CDMA Transceivers

Key Topics: CDMA transceivers, RF front-ends, analog-to-digital conversion, digital-to-analog conversion, power amplifiers

Overview: The basic components of a CDMA transceiver are examined using the specific case of an IS-95 handset. IS-95 CDMA handsets are of particular interest due to the fact that they must typically maintain linear behavior over a large dynamic range when compared to other public wireless systems. The effects of analog-to-digital and digital-to-analog conversion, intermodulation distortion, and other effects of analog front-ends on CDMA performance are explored.

1. INTRODUCTION

CDMA transceivers operate under several special conditions that introduce design considerations not normally brought into play when designing GSM, TDMA, or AMPS transceivers. Among the major differences are:

a. Dynamic range of operation must be large for fast power control mechanisms.
b. Operation under adjacent channel interference from narrowband systems (e.g., AMPS, TDMA) is difficult due to the large power difference between CDMA signals and narrowband signals.
c. Wider bandwidth CDMA signals can suffer from in-band distortion products that may not affect narrowband systems.

Of critical importance is CDMA handset design under these conditions. In general, handsets are designed to be compact, inexpensive, and power efficient. However, addressing problems such as intermodulation distortion, adjacent channel interference, and emissions reduction can result in increases in size, cost, and power consumption.

In this chapter, CDMA handset transceiver design is examined. In particular, the case of IS-95/cdma2000 handset design will be investigated. Much of the analysis may also be applied to base station transceiver design; however, much of the optimization necessary to produce a competitive mobile phone product is not so critical for base station radios.

A typical wireless terminal receiver for a full-duplex BPSK/QPSK CDMA system (pictured in Figure A-1) consists of several components; namely, the duplexer, the low-noise amplifier (LNA), the first mixer or IF mixer, a channel selection filter (often implemented as a surface acoustic wave [SAW] filter at an intermediate frequency [IF]), the automatic gain control (AGC) amplifier, the I-Q demodulator, the baseband antialiasing (AAF) filters, the analog-to-digital (ADC, or A/D) converters, and the digital section.

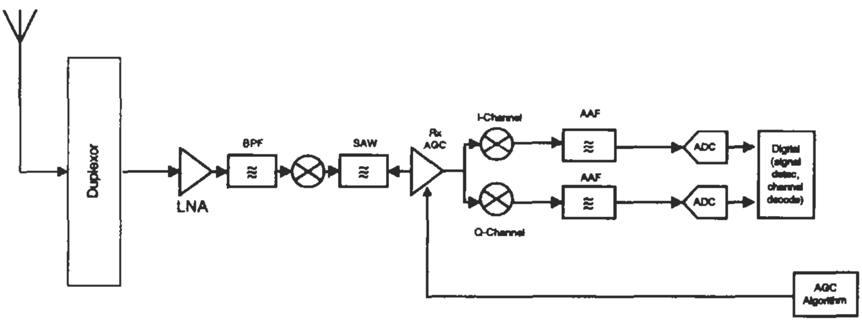

Figure A-1: Receiver Chain

Partitioning this functionality into three sections of the handset receiver is usually desirable. There three sections are the RF (radio frequency) front-end, which encompasses all circuitry between the antenna and the antialiasing filters (AAFs); the mixed-signal section, which includes the AAF filters and A/D converters; and finally the digital section.

An overview of the CDMA transmitter chain is shown in Figure A-2.

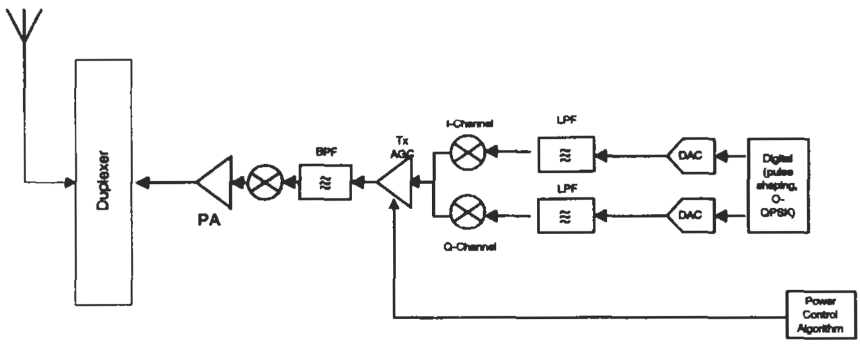

Figure A-2: CDMA Transmitter Chain

Working from right to left in Figure A-2, after offset-QPSK modulation (in the IS-95 case) and pulse shaping, the I and Q signals respectively drive two fixed-resolution digital-to-analog converters (DACs). These DAC outputs in turn are filtered to reject sampling images and to meet electromagnetic compatibility requirements. These signals are then converted to an IF frequency by modulation with orthogonal waveforms (IQ modulation) and passed into a transmit AGC (Tx AGC) amplifier. This signal in turn is filtered for rejection of spurious products and upconverted to the desired carrier frequency. The signal is passed into a power amplifier (PA) and then through the duplexer to the antenna.

2. INTERMODULATION

Intermodulation distortion is a phenomenon that occurs in wireless systems, and can be detrimental to wireless transceiver performance. This effect can impact both the receiver and transmitter in a wireless system. As a result, there exists much concern over reducing the degradation caused by intermodulation.

Intermodulation occurs as a result of the use of nonlinear components in typical wireless transceivers. When spurious interference is provided as input to nonlinear elements, this interference tends to appear in the output of these nonlinear elements as a linear interference term and several nonlinear interference terms. The nonlinear interference terms are often modeled as the weighted sum of powers of the input interference term, and therefore can be troublesome. Problems particularly arise when the nonlinear interference terms increase in power as a function of the input power levels to nonlinear components.

The effect of intermodulation is problematic in coherent-detection CDMA (code division multiple access systems). CDMA transceivers must normally operate over a large dynamic range, due to the need for feedback power control and pulse shaping. The most widely deployed CDMA system for public use today is the IS-95 system [4], which is implemented in many parts of North America, Korea, and Japan. The effect of intermodulation on this system is of particular interest for handset transceivers, as handset costs are impacted by linearity requirements in radio frequency (RF) components.

2.1 Intermodulation Theory

Ideal transceiver components will have a desired linear response to an infinite range of input signal levels. However, nonideal components may have a voltage output response given by

$$V_{out} = \sum_{n=0}^{\infty} a_n V_{in}^{\ n} \tag{A-1}$$

where V_{in} is the input signal voltage level, a_n is a scalar coefficient, and V_{out} is the output voltage level. In a communications system, this type of response results in undesired products. For instance, consider the case where the input signal is

$$V_{in} = A\cos(\omega_1 t) + B\cos(\omega_2 t) \tag{A-2}$$

In (A-2), assume that the sinusoid at frequency ω_1 is the desired signal and the sinusoid at ω_2 is an undesired tone with sufficiently large spectral separation from the desired signal. The first two terms of the expansion of (A-1) will be linear terms: a_0, and $a_1[A\cos(\omega_1 t)+B\cos(\omega_2 t)]$. a_0, or the DC term, is often considered negligible. a_1 is the device small-signal, or linear, gain. However, the second-order intermodulation product is

$$V_{out}^{(2)} = a_2 \left\{ \begin{matrix} \frac{A^2}{2}(1+\cos(2\omega_1 t)) + \frac{B^2}{2}(1+\cos(2\omega_2 t)) + \\ AB[\cos((\omega_1 - \omega_2)t) + \cos((\omega_1 + \omega_2)t)] \end{matrix} \right\} \tag{A-3}$$

This second-order product contains a spectral component at ω_1-ω_2, which increases exponentially with the input signal by a power of 2. Moreover, the third-order term in the expansion is:

(A-4)

$$V_{out}^{(3)} = a_3 \left\{ \begin{array}{l} (\frac{3A^3}{4} + \frac{3AB^2}{4})\cos(\omega_1 t) + \\ (\frac{3B^3}{4} + \frac{3A^2B}{4})\cos(\omega_2 t) + \\ \frac{A^3}{4}\cos(3\omega_1 t) + \frac{B^3}{4}\cos(3\omega_2 t) + \\ \frac{3AB^2}{4}[\cos((\omega_1 - 2\omega_2)t) + \cos((\omega_1 + 2\omega_2)t)] + \\ \frac{3A^2B}{4}[\cos((2\omega_1 - \omega_2)t) + \cos((2\omega_1 + \omega_2)t)] \end{array} \right\}$$

This third-order product is particularly troublesome in transceiver design because the signals at frequencies $2\omega_1$-ω_2 and ω_1-$2\omega_2$ lie very close to the signal of interest in the spectral domain; in fact, many times this signal will fall on top of the desired signal bandwidth. Moreover, this term increases exponentially in magnitude by a power of 3 when compared to the desired linear term.

An intercept point is defined as the input signal power level at which an undesired higher-order output product is equal to the desired linear output signal. A device in transceiver design is often characterized by its intercept points. Of particular interest are the second- and third-order intercept points, which correspond to the terms in (A-3) and (A-4), respectively. Although the second-order intercept point is important, the third-order intercept point is considered more critical due to this product's proximity to the desired signal band. Nevertheless, the *n*th-order intercept point can be defined as

(A-5)

$$IPn = A + \frac{\Delta}{n-1}$$

where A is the input signal level in dBm and Δ is the difference in dB between the desired signal level and undesired product level. This concept is illustrated in Figure A-3.

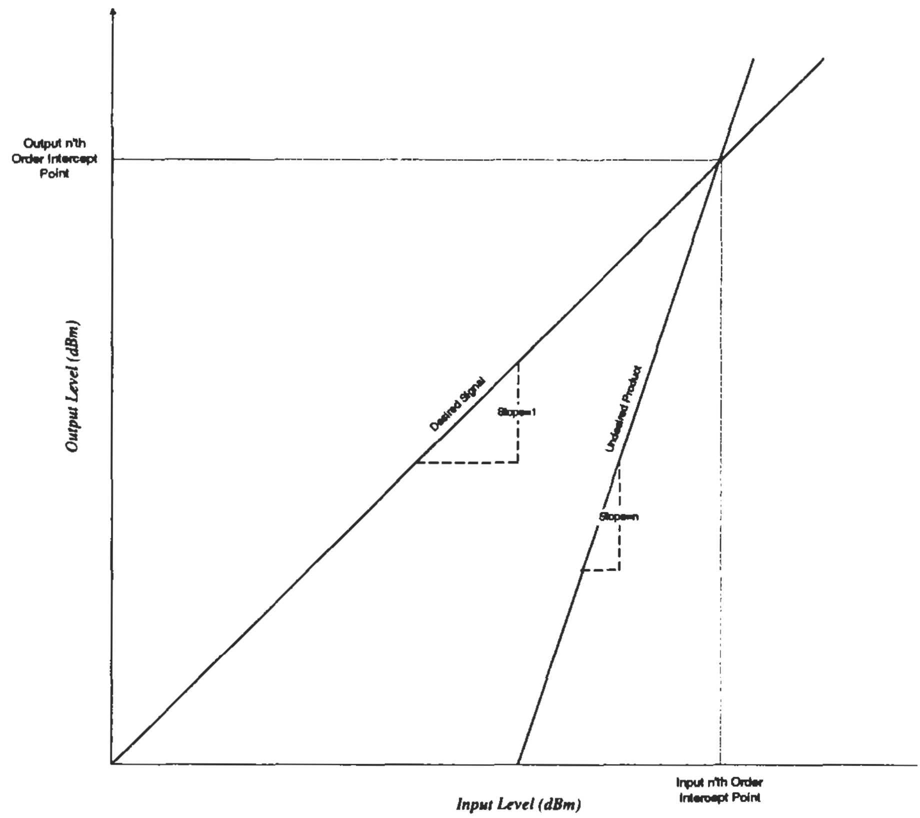

Figure A-3: Intercept Point

Using the second-order intercept point, a_2 can be found by setting $A = Vip_2$ and $B = 0$ in (A-2), inferring that $V_{out}^{(2)} = V_{out}^{(1)}$ when

$$a_2 = \frac{a_1}{Vip_2} \tag{A-6}$$

Similarly, a_3 may be found by setting $A = Vip_3$:

$$a_3 = \frac{4a_1}{3Vip_3^{\,2}} \tag{A-7}$$

Since the third-order product increases cubically compared to the desired linear product, it is necessary to have adequate filtering and gain control to reduce spurious products that may dominate at the input stages to nonlinear elements in the transceiver.

The intercept points of a system are useful for determining system performance; however, the intercept points are *fictional* in that the system or individual device will reach saturation well before the point when input signal levels are such that the intercept points are achieved. As a result, the intercept point is a useful measure for *small signal* analysis. *Large signal* analysis must take into account the compression point in addition to the intercept point. The 1-dB compression point for a system is defined as the input signal level at which the output voltage has decreased 1 dB with respect to the linear gain level. This value is closely related to the intercept point; using the previous analysis, we can see exactly how they are related. Returning to (A-2), if the undesired term is excluded (i.e., $B = 0$), then the only contributions to the output are the linear term and the third-order term. Then when $A = V_{IP}$, the ratio of linear gain to actual gain becomes

$$\frac{a_1 V_{cp}}{a_1 V_{cp} + a_3 \frac{3}{4} V_{cp}^{\ 3}} = 10^{\frac{1}{20}} \tag{A-8}$$

If the value for a_3 derived in (A-7) is substituted in (A-8), then a closed-form solution for the relationship between the 1-dB compression point and the third-order intercept point may be derived:

$$10\log_{10}\left(\frac{V_{ip3}^{\ 2}}{V_{cp}^{\ 2}}\right) = 10\log_{10}\left(\frac{1}{1-10^{\frac{-1}{20}}}\right) \tag{A-9}$$

This implies a 9.638 dB difference between a 1-dB compression point and third-order intercept point. This relationship is a good rule of thumb, but cannot be guaranteed in practice.

2.2 Receiver Intermodulation

Receiver intermodulation usually results from the presence of interfering signals in the proximity of the receiver. These interfering signals induce intermodulation distortion in the nonlinear components present in a typical wireless receiver. For instance, in typical IS-95 CDMA systems operating in the cellular band (806–890 MHz) in North America, handsets often experience intermodulation distortion from AMPS (Advanced Mobile Phone

Service) base station transmitters ([2],[3]). Since AMPS is a frequency-modulated narrowband system, the required carrier-to-interference ratios for these systems are much higher than IS-95 systems, and therefore inter-modulation distortion in IS-95 receivers can degrade performance completely [4]. The sources of degradation can be more clearly evaluated when examining the receiver chain being used and isolating all nonlinear elements in this chain.

2.2.1 Wireless Receivers

Referring to Figure A-1, the elements that will be subject to the greatest intermodulation distortion are the active components that operate at high frequencies. LNAs and other preamplifiers (e.g., AGC amplifiers), and mixers are examples of such devices. Passive elements, such as SAW filters, typically have a sufficiently large range of linearity so that they are often assumed to be infinitely linear.

In order to determine if a typical receiver chain can achieve the required intercept point, one needs to understand and evaluate this intercept point based on the individual components' intercept points, passband gain, and selectivity. This can be found with the following formula:

$$\frac{1}{IP3} = \frac{1}{IP3_{Duplexor}} + \frac{G_{duplexor}}{IP3_{LNA}} + \frac{G_{duplexor}G_{LNA}}{IP3_{SAW}} + \dots \tag{A-10}$$

Passive elements such as a SAW filter display linear behavior over a large input range, and thus their individual $IP3$s are assumed to be infinite. However, mixer and amplifier stages do not display such behavior. Another interesting aspect of the formula in (A-10) is that, unlike the noise figure, the second-to-last stage in the receiver chain has a large impact on system $IP3$. Finally, the cumulative gain in each stage must take into account selectivity, as the difference between the interference signal level and the desired signal level for the interference can decrease after every filtering stage. Thus, (A-10) may be modified to take this into account:

$$\frac{1}{IP3} = \frac{1}{IP3_{Duplexor}} + \frac{G_{duplexor}{}^{IM} / G_{duplexor}}{IP3_{LNA}} + \frac{G_{duplexor}{}^{IM} G_{LNA}{}^{IM} / \left(G_{duplexor} G_{LNA}\right)}{IP3_{SAW}} + \ldots \tag{A-11}$$

where $G_{(\ )}{}^{IM}$ is the gain at the intermodulation product frequency. Note that these equations only provide a first-pass analysis and that system $IP3$ calculation cannot be accurately determined without actual testing of a radio implementation.

2.3 Transmitter Intermodulation

Transmitter-induced intermodulation can be just as troubling as receiver intermodulation, due to the fact that the required modulation becomes compromised as a result of internal intermodulation products. This effect in turn makes it difficult to manufacture radios that meet mandated electromagnetic compatibility requirements in addition to maintaining required modulation accuracy.

Referring to Figure A-2, the Tx AGC amplifier along with the PA provides the necessary dynamic range for CDMA power control to operate. This requires a suitable range of linearity (at least 80 dB, according to the minimum performance specifications for IS-95 terminals IS-98-A [5]). These two amplifiers are also subject to intermodulation distortion; however, the sources of the interference are internal, and are usually provided by the upconversion stages present in the transmitter.

2.3.1 Sources of Transmitter Internal Interference

There are two primary sources of interference that can result in intermodulation in the transmitter chain pictured in Figure A-2: the IQ-modulation and the upconverter.

2.3.1.1 IQ-Modulator

The IQ modulator takes baseband input from both the I and Q paths and converts it to an IF frequency by modulating each respective path by mutually orthogonal waveforms (usually phase-synchronous sine and cosine waves). In addition to the desired waveform, the IQ modulator also provides undesired interference terms that can result in intermodulation within the Tx AGC amplifier.

The undesired interference terms can be traced back to gain and phase mismatch between the I and Q paths. The mismatch in gain and phase between the I and Q paths manifests itself in the output of the IQ mixer as two effects: sideband leakage and carrier leakage. In order to understand these effects, a mathematical model for the IQ mixer may be used. A typical IQ mixer is shown in Figure A-4.

Typically, the local oscillator frequency is referred to as ω_{LO}. The 90-degree phase shift results in orthogonality between the modulating signals; this results in quadrature modulation. The signals coming in on the I and Q channels may be represented as

$$
\begin{aligned}
i(t) &= A(t)\cos(\omega_m t) \\
q(t) &= B(t)\sin(\omega_m t)
\end{aligned}
\tag{A-12}
$$

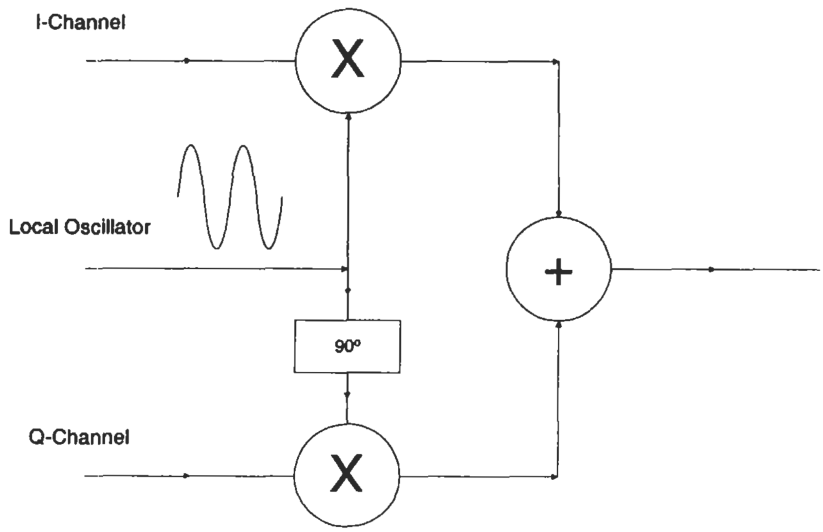

Figure A-4: IQ-Mixer

where $A(t)$ and $B(t)$ are the I and Q information signals and ω_m is the modulation frequency. The ideal output of the IQ mixer would be [6]

$$y(t) = i(t)\cos(\omega_{LO}t) + q(t)\sin(\omega_{LO}t) \tag{A-13}$$

However, due to amplitude and phase imbalances in the modulating signals and amplitude and phase imbalances within the mixer, this ideal output does not occur in reality. For the sake of this analysis, it is assumed that both the amplitude and phase imbalances of the modulating signal and the mixer can be combined into one amplitude imbalance and one phase imbalance term, with both of these effects showing up on the Q channel. Moreover, one must take into account the individual DC offset terms, which will be present in the modulating signal and the mixer. IQ mixer output actually is

$$\begin{aligned} y(t) = {} & [i(t) + dc_m][\cos(\omega_{LO}t) + dc_{LO}] + \\ & K[q(t) + dc_m][\sin(\omega_{LO}t + \phi) + dc_{LO}] \end{aligned} \tag{A-14}$$

where dc_m and dc_{LO} are the DC offsets of the modulating signal and mixer, respectively; K is the amplitude imbalance; and ϕ is the phase imbalance. The I and Q channel modulating signals are expanded to result in

$$\begin{aligned} y(t) = {} & [A(t)\cos(\omega_m t) + dc_m][\cos(\omega_{LO}t) + dc_{LO}] + \\ & K[B(t)\sin(\omega_m t) + dc_m][\sin(\omega_{LO}t + \phi) + dc_{LO}] \end{aligned} \tag{A-15}$$

This can in turn be expanded to

$$\begin{aligned} y(t) = {} & \frac{1}{2}\cos((\omega_m - \omega_{LO})t)[A(t) + KB(t)\cos(\phi)] + \\ & \frac{1}{2}B(t)\sin((\omega_m - \omega_{LO})t)[KB(t)\sin(\phi)] + \\ & \frac{1}{2}\cos((\omega_m + \omega_{LO})t)[A(t) - KB(t)\cos(\phi)] + \\ & \frac{1}{2}B(t)\sin((\omega_m + \omega_{LO})t)[KB(t)\sin(\phi)] + \\ & dc_m\cos(\omega_{LO}t) + Kdc_m\sin(\omega_{LO}t)\cos(\phi) + \\ & Kdc_m\cos(\omega_{LO}t)\sin(\phi) + A(t)dc_{LO}\cos(\omega_m t) + \\ & KB(t)dc_{LO}\sin(\omega_m t) + dc_m dc_{LO} + Kdc_m dc_{LO} \end{aligned} \tag{A-16}$$

The desired terms appear at the frequency $\omega_m - \omega_{LO}$. The other terms are undesired and can be problematic. Usually, the DC terms and the terms corresponding to the modulation frequency ω_m are easy to filter if the desired term is sufficiently high in frequency. However, two extremely troublesome products are the sideband products, corresponding to the frequency $\omega_m + \omega_{LO}$, and the carrier products, corresponding to the frequency ω_{LO}. For simplification purposes, let as assume the $A(t)$ and $B(t)$ are power normalized such that their power is always equal to 1. The ratio of the power of the sideband products to the desired products is known as *sideband suppression*, and can be evaluated as

$$\textit{Sideband suppression}(dB) = 10\log_{10}\left(\frac{1-2K\cos(\phi)+K^2}{1+2K\cos(\phi)+K^2}\right) \tag{A-17}$$

The ratio of the power of the carrier frequency, or LO frequency, products to the desired products is known as the *carrier suppression*, and may be evaluated as

$$\textit{Carrier suppression}(dB) = 10\log_{10}\left(\frac{dc_m{}^2+2Kdc_m dc_{LO}\sin(\phi)+K^2 dc_m{}^2}{\frac{1}{4}\left(1+2K\cos(\phi)+K^2\right)}\right) \tag{A-18}$$

In a transmit IQ mixer, the carrier terms and sideband terms tend to fall extremely close to, if not on top of, the desired signal in the spectral domain. This is due to the modulating frequency being at or near baseband. Therefore, one must take care in baseband design to allow sufficient margin for the distortion an IQ mixer may provide. However, one may alleviate the problems of carrier and sideband leakage through DC-offset correction, amplitude balancing, and phase balancing in the input stages to the IQ modulator.

2.3.1.2 Upconverter

The upconverter serves to translate a signal at an IF frequency to the final carrier frequency. The interference that it produces will in turn drive the PA and can result in unwanted products in the transmitted signal. In this section, an image-rejecting upconverter is analyzed; this type of upconverter has inherent rejection of the undesired sideband. The upconverter is a high-frequency device, whose modeling is similar to the IQ modulator. A common method to implement upconversion by means of the image-reject mixer is shown in Figure A-5.

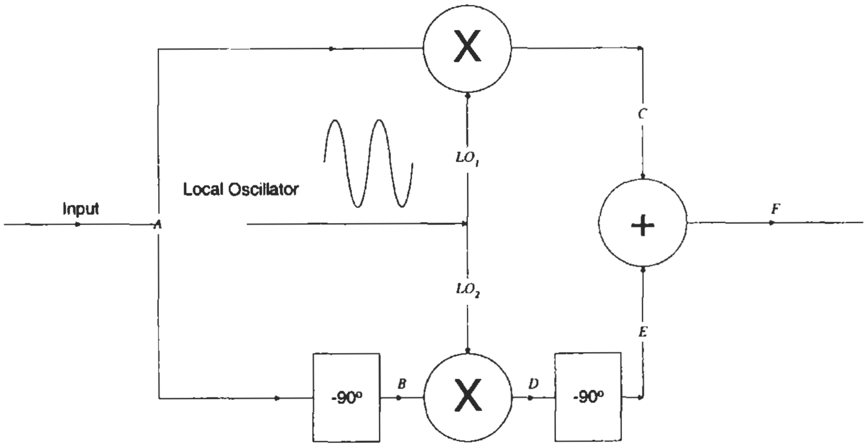

Figure A-5: Image-Reject Mixer

This mixer has the advantage of suppressing the undesired sideband. This can be seen by modeling the response of this mixer to a sinusoid. Assuming that an input signal at frequency ω_{IF} is to be upconverted to a frequency at $\omega_{IF} + \omega_{LO}$, then two input sinusoids are required at frequencies ω_{IF} and ω_{LO}. Referring back to Figure A-5, then the input signal at point A is given by

$$A = I\cos(2\pi\omega_{IF}t) + dc_{IF} \tag{A-19}$$

where I is the amplitude of the input, and dc_{IF} is a DC offset. This signal is passed through a 90-degree hybrid coupler [7] to produce the signal at point B:

$$B = I\cos\left(2\pi\omega_{IF}t + \frac{2\pi}{4(\omega_{LO} - \omega_{IF})}\right) + dc_{IF} \tag{A-20}$$

The signal at points B and A is now to be modulated with a sinusoid at frequency ω_{LO} to produce the signals at points C and D. However, much like the IQ mixer, this local oscillator signal will be subject to gain and phase imbalances. Thus the signals at LO_1 and LO_2 may be represented as:

$$LO_1 = Lc\cos(2\pi\omega_{LO}t + \theta) + dc_{LO}$$
$$LO_2 = L\cos(2\pi\omega_{LO}t) + dc_{LO} \tag{A-21}$$

where L is the nominal amplitude, c is the gain mismatch scaling constant, θ is the phase mismatch, and dc_{LO} is the DC offset of the LO signal. Thus the signal at point C can be represented as

$$\begin{gathered} C = A \bullet LO_1 = \\ \frac{ILc}{2}\cos(2\pi(\omega_{LO} + \omega_{IF})t + \theta) + \\ \frac{ILc}{2}\cos(2\pi(\omega_{LO} - \omega_{IF})t + \theta) + \\ Lcdc_{IF}\cos(2\pi\omega_{LO}t + \theta) + Idc_{LO}\cos(2\pi\omega_{IF}t) + dc_{IF}dc_{LO} \end{gathered} \tag{A-22}$$

One must take note of the fact that in the above expansion, an IF leakage term has surfaced. C may be expanded even further to yield sideband scaling and LO leakage terms:

$$\begin{gathered} C = \frac{\cos(\theta)ILc}{2}\cos(2\pi(\omega_{LO} + \omega_{IF})t) - \\ \frac{\sin(\theta)ILc}{2}\sin(2\pi(\omega_{LO} + \omega_{IF})t) + \\ \frac{\cos(\theta)ILc}{2}\cos(2\pi(\omega_{LO} - \omega_{IF})t) - \\ \frac{\sin(\theta)ILc}{2}\sin(2\pi(\omega_{LO} - \omega_{IF})t) + \\ Lcdc_{IF}\cos(\theta)\cos(2\pi\omega_{LO}t) - Lcdc_{IF}\sin(\theta)\sin(2\pi\omega_{LO}t) + \\ Idc_{LO}\cos(2\pi\omega_{IF}t) + dc_{IF}dc_{LO} \end{gathered} \tag{A-23}$$

The signal at point D results from the modulation of the signal at point A passed through the 90-degree hybrid. This signal is:

(A-24)

$$D = B \bullet LO_2 =$$

$$\frac{IL}{2}\cos\left(2\pi(\omega_{LO} - \omega_{IF})t - \frac{2\pi}{4(\omega_{LO} - \omega_{IF})}\right) +$$

$$\frac{IL}{2}\cos\left(2\pi(\omega_{LO} + \omega_{IF})t + \frac{2\pi}{4(\omega_{LO} - \omega_{IF})}\right) +$$

$$Ldc_{IF}\cos(2\pi\omega_{LO}t) + Idc_{LO}\cos\left(2\pi\omega_{IF}t + \frac{2\pi}{4(\omega_{LO} - \omega_{IF})}\right) + dc_{IF}dc_{LO}$$

Once again, an IF leakage term results from an expansion of D:

(A-25)

$$D = \frac{IL}{2}\cos\left(2\pi(\omega_{LO} - \omega_{IF})t - \frac{2\pi}{4(\omega_{LO} - \omega_{IF})}\right) +$$

$$\frac{IL}{2}\cos\left(2\pi(\omega_{LO} + \omega_{IF})t + \frac{2\pi}{4(\omega_{LO} - \omega_{IF})}\right) +$$

$$Ldc_{IF}\cos(2\pi\omega_{LO}t) + Idc_{LO}\cos\left(\frac{2\pi}{4(\omega_{LO} - \omega_{IF})}\right)\cos(2\pi\omega_{IF}t) -$$

$$Idc_{LO}\sin\left(\frac{2\pi}{4(\omega_{LO} - \omega_{IF})}\right)\sin(2\pi\omega_{IF}t) + dc_{IF}dc_{LO}$$

The signal at point D is now passed through another 90-degree hybrid to yield

(A-26)

$$E = \frac{IL}{2}\cos\left(2\pi(\omega_{LO} - \omega_{IF})t - \frac{\pi}{(\omega_{LO} - \omega_{IF})}\right) +$$

$$\frac{IL}{2}\cos(2\pi(\omega_{LO} + \omega_{IF})t) +$$

$$Ldc_{IF}\cos\left(2\pi\omega_{LO}t - \frac{2\pi}{4(\omega_{LO} - \omega_{IF})}\right) +$$

$$Idc_{LO}\cos\left(2\pi\omega_{IF}t - \frac{2\pi}{4(\omega_{LO} - \omega_{IF})}\right) + dc_{IF}dc_{LO}$$

Referring back to (A-23), one can see the terms corresponding to the undesired sideband at $\omega_{LO} - \omega_{IF}$ are eliminated under ideal conditions when the signals at points E and C are summed together to yield F. This is due to the fact that when the phase mismatch θ is 0 degrees, the term in C corresponding to $\sin(2\pi(\omega_{LO}-\omega_{IF})t)$ disappears and the term in C corresponding to $\cos(2\pi(\omega_{LO}-\omega_{IF})t)$ is 180 degrees out of phase with the $\cos(2\pi(\omega_{LO}-\omega_{IF})t)$ term in E.

Of course, in reality the signal at F will have many undesired products, as indicated by the equations above. The four primary products present at the output of the image-reject mixer are the desired upper sideband, the lower sideband leakage, the IF leakage, and the LO leakage. The desired signal power may be expressed as

(A-27)

$$Desired = \frac{1}{2}\left(\frac{\cos(\theta)ILc}{2}\right)^2 + \frac{1}{2}\left(\frac{IL}{2}\right)^2$$

The lower sideband, LO, and IF powers may be expressed as

(A-28)

$$Sideband = \frac{1}{2}\left(\frac{\cos(\theta)ILc}{2} - \frac{IL}{2}\right)^2 + \frac{1}{2}\left(\frac{\sin(\theta)ILc}{2}\right)^2$$

(A-29)

$$IF = \frac{1}{2}(L^2c^2dc_{IF}{}^2 + I^2dc_{LO}{}^2)$$

(A-30)

$$LO = \frac{1}{2}(I^2dc_{LO}{}^2 + L^2dc_{IF}{}^2)$$

The DC terms at point F are usually filtered by a DC block and are not generally considered of consequence. However, the lower sideband term, the LO term, and the IF term may not be very easy to filter out. Of particular interest are the LO and IF isolation values, defined as

(A-31)

$$LO\,Isolation(dB) = 10\log_{10}(Desired/LO)$$

$$IF\,Isolation(dB) = 10\log_{10}(Desired/IF) \tag{A-32}$$

The analysis presented above assumed the gain and phase mismatch to be manifest in the LO signal to simplify analysis. However 90-degree transformer hybrids usually suffer from gain imbalances and phase mismatch. If one chooses another topology to accomplish the phase shifting, gain and phase matching still will be critical for performance. It should also be noted that the summing amplifier needed for summing C and E together to produce F may behave nonlinearly over much of the upconverter's dynamic range. However, advanced circuit topologies may be able to alleviate this problem.

2.3.2 Transmitter Components Subject to Intermodulation

The Tx AGC amplifier and PA are nonlinear elements in a CDMA system that must maintain a suitably large range of linearity.

2.3.2.1 Transmit AGC Amplifier

The transmit AGC amplifier is used primarily for power controlling the handset. These amplifiers typically support at least 80 dB of dynamic range, as per the requirements of IS-98-A [5]. Typically, the third-order polynomial model given in Section 2.1 is appropriate for modeling such amplifiers. However, an AGC amplifier is variable gain, meaning that its intercept point changes with each gain setting. The worst-case intercept point occurs when the amplifier is at maximum gain; therefore, it makes sense to verify operation at the maximum gain setting of the AGC amplifier. For instance, the component described in [8] provides a -18 dBm input third-order intercept point at 35 dB of gain setting. At gain settings greater than 35 dB, the intercept point reduces quickly to -27 dBm at 45 dB of gain setting. The component described in [9] provides -26 dBm at 39 dB maximum gain setting.

2.3.2.2 Power Amplifier

CDMA PA's typically must operate over a large dynamic range. As given in [10], the input and output voltages for the power amplifier may be given as:

(A-33)

$$v_i(t) = \mathrm{Re}\{\rho e^{j2\pi f_c t}\}$$
$$v_o(t) = \mathrm{Re}\{A(|\rho|)e^{j2\pi f_c t + j\theta(|\rho|)}\}$$

where f_c is the carrier frequency, ρ is the input complex magnitude, and $A(\ )$ and $\theta(\ )$ represent the gain and phase characteristics of the PA. The gain distortion term represents the amplitude-to-amplitude modulation (AM-to-AM) and the phase distortion term represents the amplitude-to-phase modulation (AM-to-PM) effects seen in many nonlinear devices. These functions can normally be derived from a single-tone test, where the effects of PA distortion on a singular frequency are captured. In [10], the authors demonstrate that by using this model, they were able to effectively simulate their two-stage gallium-arsenide MESFET amplifier and demonstrate the device's compliance with emission requirements for PCS CDMA.

Due to gain and phase distortion, spectral regrowth results in the output of the PA. This is troublesome due to the need to meet specific electromagnetic compatibility requirements. Moreover, input stages to the PA that provide spurious products will contribute difficulty in meeting the required spectral emissions mask.

3. NOISE FIGURE

The receiver noise figure is the factor by which thermal noise is scaled in the receiver circuitry. This value particularly affects *receiver sensitivity*, i.e., the minimum in-band received signal power at which the receiver can successfully demodulate the incoming signal. Given the individual noise figures for the duplexer, LNA, etc., the overall noise figure may be calculated by Friss's formula as

(A-34)

$$NF = NF_{duplexor} + \frac{NF_{LNA}}{Gain_{duplexor}} + \frac{NF_{BPF}}{Gain_{duplexor}Gain_{LNA}} + \ldots$$

This formula indicates that the noise figure is dominated by the initial stage of the receiver.

The noise figure results in an increase in the effective noise floor that the signal of interest sees by the time it arrives into the digital section. Digital circuitry typically has a sufficiently small noise figure to be assumed negligible. A typical partitioning is shown in Table A-1.

Parameter	Duplexer	LNA	BPF	Mixer	Saw	Rx AGC	IQ-Mixer	AAF
Gain (dB)	–4	16	–3	13	–7	45	20	30
Gain (lin)	0.398107	39.81072	0.501187	19.95262	0.199526	31622.78	100	1000
Noise Figure	4	1.5	3	9	0	5	21	0
Noise Factor	2.511886	1.412538	1.995262	7.943282	1	3.162278	125.8925	1
Cumulative Gain	–4	12	9	22	15	60	80	110
F Composite	2.511886	3.548134	3.610931	4.485038	4.485038	4.553415	4.55354	4.55354
NF Composite	4	5.5	5.576192	6.517661	6.517661	6.583373	6.583492	6.583492

Table A-1: Noise Figure Partitioning

The impact of the CDMA noise figure is felt particularly at low signal levels, where the thermal noise floor accounts for most of the interference in the 1.25 MHz CDMA signal bandwidth. According to J-STD-018 [11], the equivalent of IS-95-A for the 1.9 GHz PCS band, the receiver must be able to demodulate a CDMA signal whose in-band received signal energy is as little as I_{orx} = -104. Given that the desired traffic channel in this test is -15.6 dB down from the total in-band signal power (i.e., Traffic E_c/I_{or} = -15.6 dB), the effective bit energy E_b is -104 dBm - 15.6 dB + 21.1 dB = -98.5 dBm. The required frame error rate under this test is .5%; given a lossless receiver design, this can be achieved with an SNR of as little as E_b/N_t = 2.8 dB [12]. This means that the noise floor cannot be larger than -98.5 - 2.8 = -101.3 dBm. Given that thermal noise at room temperature over a 1.25 MHz bandwidth is -113 dBm, this infers an 11.7 dB maximum noise figure for the receiver.

The noise figure of the transmitter may also be calculated. However, this noise figure is usually not so critical as the initial transmitter stages have relatively low noise figures versus their corresponding gain, and the transmitter output power is usually sufficiently large to make the contribution of thermal noise negligible at the antenna output.

4. ANALOG/DIGITAL CONVERSION

Analog-to-digital converters (ADCs or A/Ds) and digital-to-analog converters (DACs or D/As) are critical for the receiver and transmitter chains, respectively. The impacts of each are analyzed in this section.

4.1 Analog-to-Digital Conversion

The A/D converter can be of particular importance in a CDMA system due to a desirable aspect of CDMA systems: the processing, or spreading, gain. This results from the energy of each information bit being spread over no less than 64 chips in IS-95 or cdma2000 spread spectrum transmission of voice. This type of modulation relaxes the requirements for dynamic range in the A/D converter; however, this is made possible with careful consideration for the automatic gain control (AGC) setpoint and possible DC offset present at the converter.

The analysis may be accomplished by making some reasonable assumptions, the primary of which is that the received signal is approximately Gaussian in nature. Although the received signal envelope may be Rayleigh distributed, two separate quantizers are used for both the real and complex parts of the signal; both components should be Gaussian to achieve a Rayleigh-distributed envelope. Although nonuniform quantizers have been proven to provide superior performance to uniform quantizers for Gaussian random processes [13], existing technology does not make nonuniform quantization feasible in hardware. Therefore, we only consider the case of uniform quantization of Gaussian random variables. Given a received continuous signal x, it is necessary to break x up into a finite number of intervals and map them to "bins" in order to quantize them. If we number the interval endpoints as $x_1, x_2, \ldots x_{L+1}$, then obviously $x_1=-\infty$ and $x_{L+1}=\infty$. If $x_2, \ldots x_L$ are placed at uniform spacings of Δ apart, then a uniform quantization can take place. The points x_2 and x_L are equal to $-x_{ol}$ and x_{ol} respectively, where x_{ol} is the quantizer setpoint, or the overload point. The output levels of the quantizer $y_1, \ldots, y_L$ are placed at the midpoints of the intervals $[x_k, x_{k+1}]$.

The quantization error for different quantizer setpoints must be analyzed to determine in which range the AGC operation must keep the incoming received power in order to achieve minimal distortion. In order to do this, the received signal must be analyzed. The received signal at the antenna connector can be written as

(A-35)

$$r(t) = \sum_{l=1}^{L}\sum_{m=1}^{M} P_m(t)\alpha_l e^{j\theta_l} * \delta(t-\tau_l) +$$
$$\sum_{b=1}^{B}\sum_{l=1}^{L(b)}\sum_{m=1}^{M(b)} P_{m,b}(t)\alpha_{l,b} e^{j\theta_{l,b}} * \delta(t-\tau_{l,b}) + n(t)$$

where L is the number of paths in the channel between the desired base station and the mobile, M is the number of mobiles being served by the desired base station, $P_m(t)$ is the signal from mobile m, α_l is the attenuation on path l, * is a convolution operator, τ_l is the delay of path l, B is the number of other base stations, and $n(t)$ is background (thermal) noise. Background noise is normally assumed to be Gaussian in nature. The second term in the sum is out-of-cell interference, which is sometimes assumed to be Gaussian (although this is not necessarily the case). The in-cell received power is composed of the signal intended for the mobile, all other signals for other mobiles, and the multipath that results from the channel. This type of multiple access interference may also be assumed to be Gaussian in nature. Therefore the received signal at the mobile is composed of the mobile signal and additive Gaussian noise. We represent this signal as

(A-36)

$$r(t) = v(t)\sqrt{E_m}\,c(t) + n(t)$$

where $v(t)$ is the voice activity factor, $c(t)$ is the data from the alphabet {-1,+1}, and E_m is the transmission energy allocated to mobile m. The data is formed as a result of source coding, channel coding, and spreading. Although the distribution is not known a priori, the data can be assumed to be either +1 or -1 with equal probabilities of .5. Moreover, although the voice activity is normally assumed to be a binary random variable [14], it will be assumed to vary slowly with respect to the AGC operation, and therefore for analysis purposes can be assumed to be constant. Therefore, we finally represent the received signal at the mobile as

(A-37)

$$r(t) = \sqrt{E_{mv}}\,c(t) + n(t)$$

The probability distribution of data in $r(t)$ is

(A-38)

$$p(c) = \frac{1}{2}\delta(t-1) + \frac{1}{2}\delta(t+1)$$

The probability distribution of $n(t)$ is already assumed to be Gaussian; since the noise and data are assumed to be independent, the probability distribution of the received signal can be derived:

(A-39)

$$p_r = p_c * p_n$$

Assuming the noise term has mean μ and standard deviation σ, the effective probability distribution becomes

(A-40)

$$p_r = \frac{1}{2}\frac{1}{\sqrt{2\pi}\sigma}e^{-\frac{(x-\mu-\sqrt{E_{mv}})^2}{2\sigma^2}} + \frac{1}{2}\frac{1}{\sqrt{2\pi}\sigma}e^{-\frac{(x-\mu+\sqrt{E_{mv}})^2}{2\sigma^2}}$$

This distribution begins to look more and more Gaussian as the standard deviation of the noise signal becomes much larger than the transmitted power of the data. In the case where the noise is zero-mean and its standard deviation is assumed to be unity, Figure A-6 shows p_r (zero-mean, unity standard deviation noise) for several different values of E_{mv}.

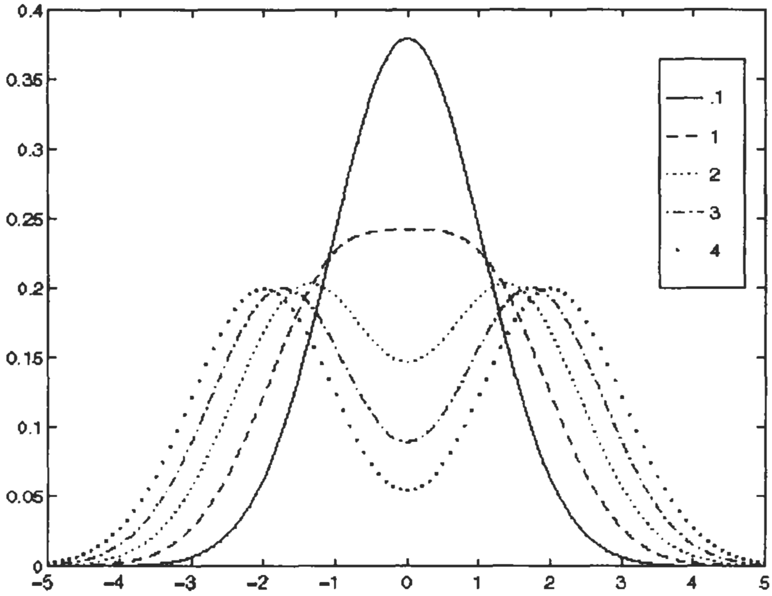

Figure A-6: Data-Modulated Signal Distribution

In the CDMA environment prior to despreading, the data modulated signal (i.e., the traffic, paging, or sync signal) is of much lower power than the interference; therefore, it is reasonable to assume that the distribution of the received signal follows the distribution of the noise. Assuming a uniform quantizer as before with L output levels and input transition points $r_1, \ldots r_{L+1}$

with overload point $r_L = r_{ol}$, we can now define the quantization error in the mean-squared sense:

(A-41)

$$E\{e^2\} = E\{(r - \bar{r})^2\}$$
$$= \sum_{k=2}^{L-1} \int_{r_k}^{r_{k+1}} (r - y_k)^2 p_r(r)dr + 2\int_{r_L}^{\infty} (r - y_L)^2 p_r(r)dr$$

The first term in the sum is known as the *granularity error*; this is the error due to quantization for input data lying in the range between the quantizer setpoints. The second term in the sum is the *overload error*; this is the error that results from clipping input data that lies outside the range given by the quantizer setpoints. The mean squared error can now be calculated as a function of the ratio of the overload point r_{ol} to the standard deviation of the input σ_r for different levels of quantization L. In Figure A-7, the effective SNR (assuming the input standard deviation is 1) is graphed for different levels of quantization.

These plots illustrate two very important concepts: (1) optimal quantizer setpoints must be satisfied to minimize mean-squared error for different levels of quantization, and (2) overload error increases more severely than granularity error as the number of quantization levels increases. This implies not only that the AGC must keep output power constant to minimize mean-squared error with respect to fixed quantizer setpoints, but that the AGC operation has more margin to take care of overload than granularity at quantization levels higher than 3 bits. In other words, the AGC should respond more quickly to input energy increases than decreases. Note: Normally the dynamic range of a quantizer is measured by the peak-to-peak input voltage level of the quantizer, known as V_{pp}, and the input signal standard deviation is taken as an RMS voltage measurement V_{rms}. Therefore r_{ol} is $V_{pp}/2$ and σ_r is V_{rms}.

To see how imperfect AGC operation can degrade performance, one can simply look at the effective frame error rate bound as an indicator. The probability of bit error (*uncoded*) for a BPSK system is normally given by [15]:

(A-42)

$$P_b = Q\left(\sqrt{\frac{E_b}{N_t/2}}\right)$$

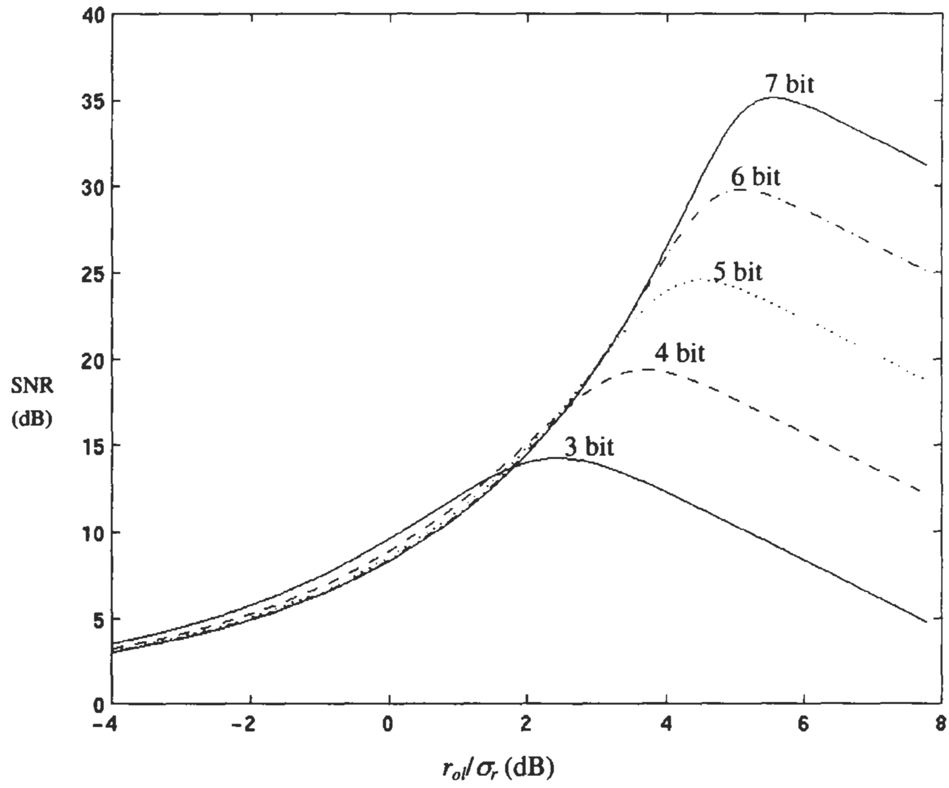

Figure A-7: Quantization Effects of Gaussian Signal

The mean-squared quantization error had previously been derived; an acceptable method of modeling this error is to assume it is Gaussian [16]. As a result, the quantization error variance (i.e., mean-squared error), denoted now by σ_e^2, can be incorporated into bit-error probability by adjusting the effective SNR as

$$P_b = Q\left(\sqrt{\frac{E_b}{N_t/2+\sigma_e^{\,2}}}\right) \tag{A-43}$$

The probability of a coded bit error at coding rate R is [17]:

$$P_b = \sum_{d=d_f}^{\infty} b_d D^d \Bigg|_{D=e^{-\frac{E_b}{N_0/2+\sigma_e^2}R}} \times \\ Q\left(\sqrt{\frac{E_b}{N_t/2+\sigma_e^2}Rd_f}\right)e^{\frac{E_b}{N_t/2+\sigma_e^2}Rd_f} \tag{A-44}$$

The value d_f in (A-44) is the minimum free distance, i.e., the Hamming distance between the all-zero path on the trellis and the path that merges with the all-zero path in the least amount of stages. This value indicates an upper bound on the coding gain. The summation indicates all other paths that merge with the all-zero path. This relation only provides an upper bound on the coding gain; however, this upper bound has been shown to be tight. This translates to frame error rates by the equation

$$P_f = 1-(1-P_b)^N \tag{A-45}$$

where N is the length of the frame in bits. For 192 bit frames (such as those that can be seen in the IS-95/cdma2000 forward paging channel), frame error rates for different levels of quantization are shown in Figure A-8 ("X"s are used to denote frame error rates under the assumption of infinite quantization).

It should be noted that the frame error rates given are ideal; IS-95-A describes a random/deterministic puncturing method that degrades the performance of the code.

The AGC setpoint can now be analyzed with respect to frame error. For instance, in the mobile environment, instantaneous power increases can occur when a mobile passes close to a base station or, as is common in cellular band operation, a mobile is near the edge of an AMPS or TDMA cell (this problem can be alleviated, of course, by bandpass filters). This is important in that Figure A-8 only shows optimal FERs for different numbers of levels used for quantization and implies that one can achieve nearly optimal performance with as little as 4 bits of quantization. However, this may not necessarily be true under suboptimal conditions where margin may be required for instantaneous incoming power changes that are too quick to be tracked by an AGC operation.

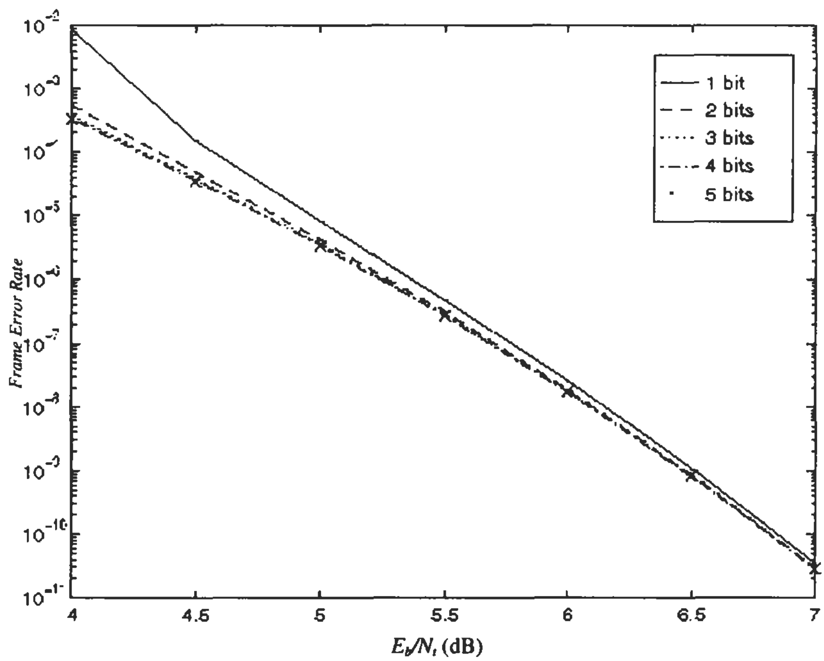

Figure A-8: Frame Error Rates for Different Levels of Quantization

As an example, the theoretical performance of an IS-98-A AWGN test for an E_b/N_t of 2.3 dB and 3.3 dB was examined with respect to different setpoints (represented as the ratio of one half of the A/D converter peak-to-peak voltage $V_{pp}/2$ to the RMS voltage V_{rms}) for different levels of quantization. The results are given in Figures A-9 and A-10 (optimal value is derived from simulation).

Five bits and six bits of quantization can provide quite a bit more margin for instantaneous power increases than 4 bits of quantization, given an AGC setpoint that is moved to the right of the curve. However, it is always preferable from a cost and power consumption perspective to keep the complexity of the quantizer at as low a level as possible.

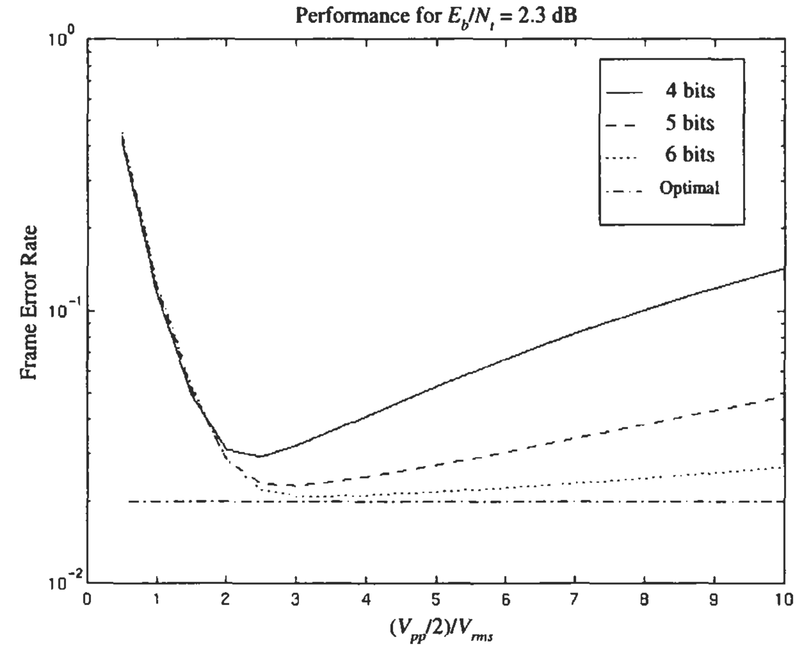

Figure A-9: Fading Margin for Different Levels of Quantization - 3.3 dB Case

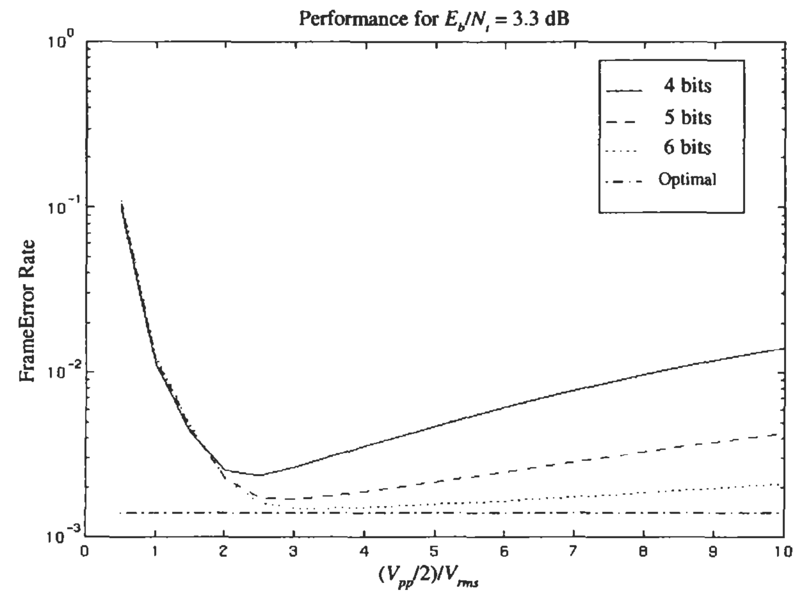

Figure A-10: Fading Margin for Different Levels of Quantization - 2.3 dB Case

4.1.1 Quantizer Nonlinearity

Practical A/D quantizers suffer from many nonidealities, primary of which are integral nonlinearity, differential nonlinearity, and DC offset. Integral nonlinearity (INL) is the peak deviation from the ideal linear quantizer mapping. This value is usually measured in terms of the number of least significant bits (LSBs). This concept is best illustrated in Figure A-11.

The deviation from the linear mapping increases quantization noise; however, A/D converters may be difficult to design if the INL requirements are very small (much less than 1 bit for most typical converters).

Differential nonlinearity is another nonideality that impacts receiver design with respect to the A/D converters. Differential nonlinearity refers to the maximum deviation from uniformly spaced quantizer bins. This value is also typically given in terms of number of LSBs. An illustration of quarter-bit DNL is shown in Figure A-12.

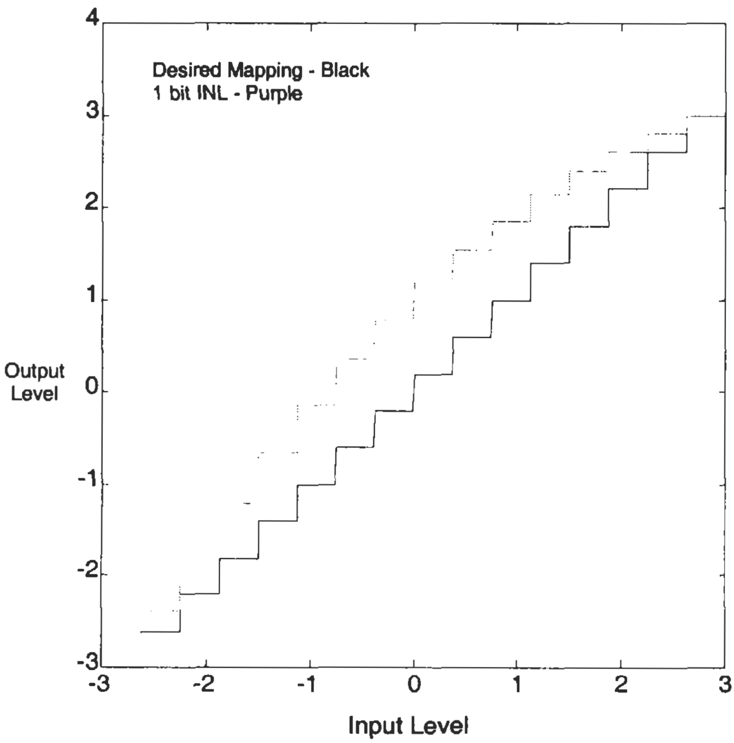

Figure A-11: Integral Nonlinearity

Although the diagram only shows positive bin size deviation, the bin size deviation can in fact be negative. For instance, for two different input values with positive difference, their corresponding output transition may have negative difference. This effect is known as *non-monotonicity*, which results in difficult system design. Therefore, a common requirement on A/D converters is monotonicity.

The A/D design will determine whether monotonicity is met. For instance, in a *successive approximation* A/D converter [18] the output is driven one bit at a time by level comparison. If the analog level of 4.5 is presented to a 3-bit A/D, the procedure would be:

1. Is 4.5 > 4? Yes, output 1 for MSB.
2. Is 4.5 > 6? No, output 0 for next bit.
3. Is 4.5 > 5? No, output 0 for last bit.

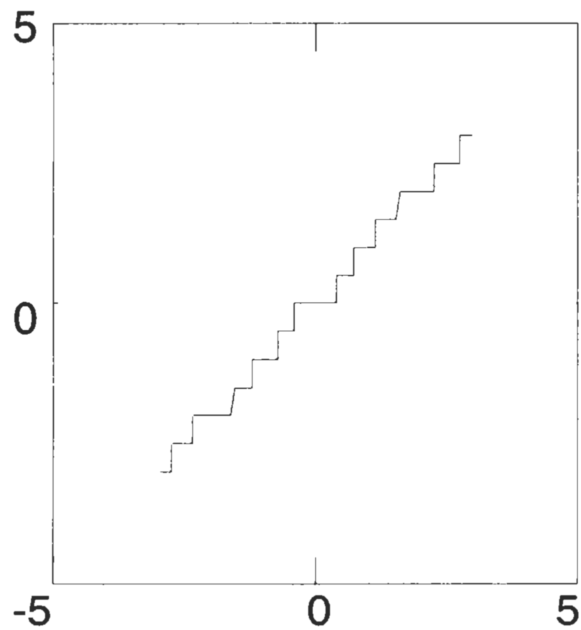

Figure A-12: 1/4-bit DNL

This type of conversion cannot be ensured to have a monotonic characteristic. An alternative design would be *word-at-a-time* output, which may be accomplished by parallel circuits all accomplishing threshold comparison. This increases the level of complexity of the A/D converter design, but provides for fast operation and ensures monotonicity.

4.1.2 DC-Offset Impact

DC offset severely impacts receiver performance as it restricts the dynamic range of the A/D quantizers. This effect may be analyzed in much the same manner as we analyzed the impact of excess input signal power to the quantizer with respect to its setpoint. Under these assumptions, the impact of DC offset power may be explored with regards to the case of E_b/N_t of 3.3 dB. Figure A-13 and Figure A-14 illustrate the effects of ½-bit and 1-bit DC offset, respectively, on performance.

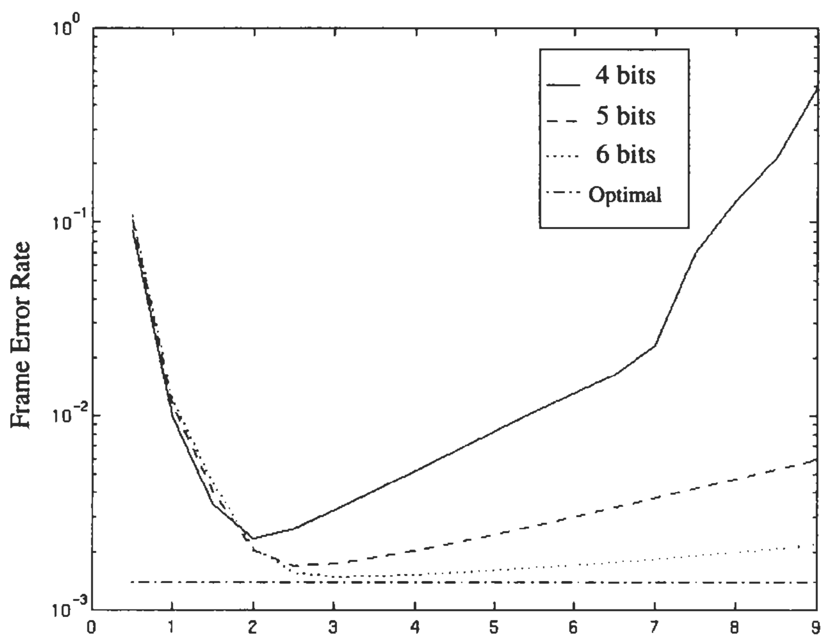

Figure A-13: 1/2-bit DC Offset

As can be seen, even a 5-bit quantizer can suffer performance degradation with 1 bit of DC offset. The 4-bit quantizer is even more sensitive. The 6-bit quantizer is the most resilient with regards to DC offset, but its performance superiority is mainly seen when the $(V_{pp}/2)/V_{rms}$ levels are greater than 6.

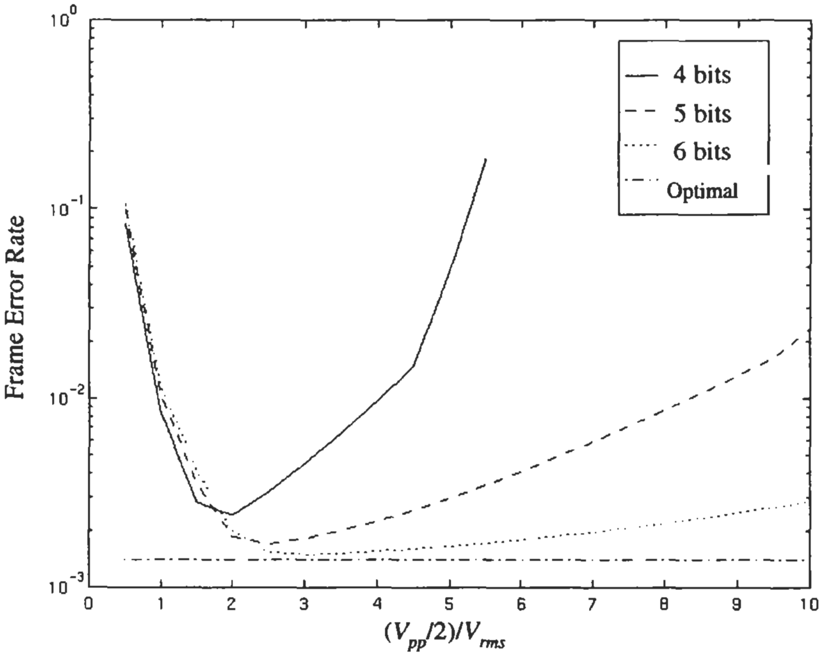

Figure A-14: 1-bit DC Offset

4.2 Digital-to-Analog Conversion

The digital-to-analog converters (DACs) can impact transmitter performance in several ways. Just like A/D converters, DACs also suffer from nonlinearity and can rarely meet 0-bit INL or DNL. Moreover, DACs are affected by DC offset in a slightly different manner from A/D converters: DC offset restricts the available dynamic range of the converter. This is due to the need to compensate for the DC offset before it reaches other parts of the transmit chain that will be sensitive to extra DC, in particular the IQ modulator. Restricting the dynamic range means scaling in the digital section, which can result in an increase in quantization noise for low-amplitude signal components; for such signal levels, the result is that the quantization noise looks white. However, restricting the dynamic range results in truncation errors, which not only become manifest in waveform quality but emissions fluctuations as well, since the quantization noise in this case is not Gaussian.

The DAC resolution and sampling rate must be chosen to maximize the *spurious free dynamic range* (SFDR). SFDR is the ratio of power between the desired frequency and the largest undesired product. It is usually determined by a single-tone test. For IS-95 systems, the convention is to use 8-bit DACs operating at 4 times the chipping frequency, i.e., 4.9152 MHz. The resolution has been shown [19] to provide sufficient SFDR without increasing complexity unnecessarily; the frequency is derived from the baseband pulse-shaping filter requirements of ¼-chip tap spacing.

5. FILTERING

The use of channel filtering in the transmit and receive chains ultimately serves to isolate the CDMA signal and reject unnecessary out-of-band signal components. However, the source of the out-of-band spectral components targeted by the receiver filter chain is different from that targeted by the transmitter filter chain.

5.1 Receiver Filter Chain

Referring to Figure A-1, several filtering chains are present, including an initial bandpass filter (BPF), a surface acoustic wave (SAW) filter to intermediate frequencies (IF), and an antialiasing filter (AAF). The BPF serves to reject images that may leak into the received signal as a result of downconversion. The SAW filter (which may be substituted with any other IF filter) serves to isolate the CDMA signal and reject some of the out-of-band components that may have resulted due to intermodulation.

The antialiasing filter serves two primary purposes: (1) to band-limit the desired signal appropriately to satisfy the Nyquist sampling theorem, and (2) to limit out-of-band interference to below the quantizer noise floor. These two purposes may be thought of as synonymous. However, CDMA poses unusual requirements for operation at low C/I ratios in the presence of large narrowband interference. Therefore, the AAFs in a CDMA system must provide sufficient channel isolation to prevent the quantizer being "swamped" by a jammer.

The filtering stages of an IS-95 system in reality consist of every single component in the transmitter and receiver, due to the fact that every device does exhibit some amount of frequency selectivity. However, the filtering stages of interest are pictured in more detail in Figure A-15.

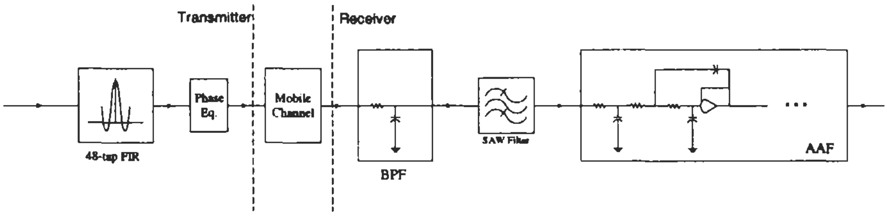

Figure A-15: Filtering Stages

The duplexer normally has a large passband as its main purpose is to isolate the transmitter and receiver, so it is not pictured. The first bandpass filter in the receiver has a large bandwidth as well, and may be implemented as a lowpass filter, therefore it is not of interest (although it is pictured). However, the SAW and AAF in the receiver are of particular interest due to the isolation they provide for the in-band signal.

The sample-and-hold functionality ($h_0(t) = 1$ for $0 < t < T$, $h_0(t) = 0$ otherwise for sampling period T) of the A/D converter may be thought of as providing some filtering; however, its frequency response is given by

$$H_0(j\omega) = \frac{\sin\left(\frac{\omega T}{2}\right)}{\frac{\omega}{2}} e^{-\frac{j\omega T}{2}} \tag{A-46}$$

Given that CDMA converters are usually oversampled by a factor of 8 times the Nyquist rate (for code tracking purposes), the sample-and-hold does not provide much filtering for interference within 1 MHz from center frequency. Therefore, the AAF is required to provide much of the filtering necessary for channel selection. Two types of filters that are typically considered for achieving this sharp selectivity are elliptic and Tchebychev filters.

Usually, AAFs are analog filters (such as Tchebychev or elliptic filters) that have poor phase responses. However, IS-95 systems require the use of a *phase equalizer* at the base station transmitter. This results in greater linearization of the phase in the passband of filters designed at the mobile receiver; however, some filter implementations will benefit more than others. The phase equalizer (Section 7.1.3.1.10.2, [4]) is given by the following transfer function:

(A-47)

$$H(\omega) = \frac{\omega^2 + ja\omega\omega_0 - \omega_0^{\,2}}{\omega^2 - ja\omega\omega_0 - \omega_0^{\,2}}$$

where a is 1.36 and ω_0 is $2\pi \times 3.15 \times 10^5$. The seventh-order Tchebychev filter has been proposed for CDMA baseband implementation [20], while the fifth-order elliptic filter was proposed in [19].

If one seeks to reduce the complexity of the antialiasing filter by increasing the resolution of the A/D, then one should be aware of the increasing complexity of the A/D converters. A simple rule of thumb yields 6 dB of relief for the antialiasing filter for every bit added to the A/D converter. Moreover, given the in-band signal power P_{in} and narrowband jammer power after antialiasing filtering P_{jam}, the RMS power level that the A/D converter sees is

(A-48)

$$P_{rms}^{\;2} = P_{jam} + P_{in}$$

Knowing this, the effective quantization noise of the in-band signal changes. This is because the AGC setpoint is arrived at due to a desired headroom for a slow AGC operation. Therefore, the AGC setpoint is suboptimal, and the quantization noise for the in-band signal worsens when the power measurement the AGC operation is using as input is dominated by the jammer. Starting with 5 bits of quantization, assuming desired channel selection from the antialiasing filter, Table A-2 lists the filter requirements for different levels of quantization in order to achieve equivalent SNR (assuming AGC setpoint of 6).

Number of bits	Suppression of tone (dB)
5 bits	43.00
6 bits	33.74
7 bits	26.60
8 bits	20.39
9 bits	14.29
10 bits	8.24
11 bits	2.21

Table A-2: Suppression Requirements for Single-Tone Interference

Keep in mind that increasing the number of bits available at the quantizer reduces the complexity of the AAF, but does not eliminate the need for additional digital filtering.

5.2 Transmitter Filter Chain

The critical filtering stages in the transmitter chain of Figure A-2 are the lowpass filters (LPFs) following the DACs. These filters are necessary to reject the spectral images that result from the digital-to-analog conversion operation. Once again, the spectral response of the sample-and-hold attenuates these images somewhat, but not to the extent necessary for ITU or FCC emissions compliance.

The first image is the most critical, as this will have the least amount of attenuation of all spectral images. This image falls at 4.9152 MHz from center and can be thought of as an interfering wideband signal that could possibly show up as an intermodulation product at the output of any active stages such as the mixers or amplifiers.

In addition, many transmitters make use of an IF SAW filter or some other sort of filter to further isolate the signal prior to transmission. Once again, intermodulation can be a problem with the IQ mixer and upconverter, so this filtering stage can help in knocking down unwanted spectral components.

6. CONCLUSIONS

A typical CDMA handset transceiver has been analyzed in this chapter, with particular attention paid to the effects of intermodulation, noise figure, and analog/digital conversion. Several other effects such as filter mismatch, IQ gain and phase imbalance, and oscillator noise are also typical transceiver effects that must be considered in the design of a CDMA system. However, using the information provided in this chapter, one can begin to analyze a typical CDMA transceiver chain.

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third-generation cdma systems for enhanced data services

Giridhar Mandyam and Jersey Lai
Nokia Research Center, Irving, Texas

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Edited by Jerry D. Gibson, Southern Methodist University

Many advances have been made in the development of third-generation wireless services in recent years. The two primary technologies that have emerged for third-generation technology are based on code division multiple access (CDMA) theory, namely Wideband CDMA (WCDMA) and cdma2000. This text provides a side-by-side description of both of these technologies so as to provide a wireless technologist, telecommunications sales or marketing specialist, or the interested reader knowledge of the similarities and differences between these two technologies. Moreover, this text provides a description of the new evolution modes of WCDMA and cdma2000, which promise to enhance maximum data throughput over the first WCDMA and cdma2000 systems that are deployed.

- Comparison of cdma2000 and WCDMA in the same text, including performance curves
- Description of 1X-EV, the evolution mode of cdma2000
- Description of IS-95, therein providing a description of all 3 major CDMA technologies in the same text
- Appendix on CDMA transceiver design, which has not been provided in any CDMA technology text to this point

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