ADDIS ABABA UNIVERSITY SCHOOL OF GRADUATE STUDIES

FACULTY OF TECHNOLOGY

DEPARTMENT OF ELECTRICAL AND COMPUTER ENGINEERING

STUDY OF LOW COMPLEXITY MULTIUSER RECEIVER DESIGN FOR SPACE-TIME CODED CODE DIVISION MULTIPLE ACCESS SYSTEMS

A thesis submitted to the school of graduate studies of Addis Ababa University

in partial fulfillment of the requirements for the degree of Masters of Science

in Electrical and Computer Engineering.

By: Girma Andargie

Advisor: Dr.Ing. Hailu Ayele

July, 2007

Addis Ababa, Ethiopia

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Declaration

I, the undersigned, declare that this thesis work is my original work, has not been presented for a degree in this or any other universities, and all sources of materials used for the thesis work have been fully acknowledged.

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Place: Addis Ababa

Date of submission:

This thesis has been submitted for examination with my approval as a university advisor.

Dr. Ing. Hailu Ayele   Signature: ____________

Advisor's Name

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List of Abbreviations

AWGN	Additive White Gaussian Noise
BPF	Band Pass Filter
CMDA	Code Division Multiple Access
DS-CDMA	Direct Sequence Code Division Multiple Access
DSSS	Direct Sequence Spread Spectrum
FDMA	Frequency Division Multiple Access
FHSS	Frequency Hopping Spread Spectrum
BM	Branch Metric
GSM	Global System Mobile
ISI	Inter Symbol Interference
MAI	Multiple Access Interference
Mbps	Mega bits per second
MC	Multi-Carrier
MLSE	Maximum Likelihood Sequence Estimator
PDMA	Polarization Division Multiple Access
PG	Processing Gain
PN	Pseudo Noise Sequence
QPSK	Quadrature Phase Shift Keying
SDMA	Space Division Multiple Access
SNR	Signal to Noise Ratio
STBC	Space-Time Block Coding
STC DS-CDMA	Space-Time Coded Direct Sequence Code Division Multiple Access
STC	Space-Time Coding
STTC	Space-Time Trellis Coding
STTC DS-CDMA	Space-Time trellis Coded Direct Sequence Code Division Multiple Access
TDMA	Time Division Multiple Access
THSS	Time Hopping Spread Spectrum

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STC	Space-Time Coding
STTC	Space-Time Trellis Coding
STTC DS-CDMA	Space-Time trellis Coded Direct Sequence Code Division Multiple Access
TDMA	Time Division Multiple Access
THSS	Time Hopping Spread Spectrum

Abstract

The goal for the next generation mobile communication systems based on CDMA is to integrate a wide variety of communication services such as high speed data, video and multimedia traffic as well as voice signals. To effect these services, high rate data transmission and reception techniques are required.

As demands for communication services increase, there is considerable economic benefit in improving the spectral efficiency of the communication system. One of the proposed approaches for these high data rate and spectrally efficient transmission in CDMA systems is space-time coding. The use of space-time coding techniques in CDMA systems increases multiuser interference. The effect of multiple access interference can be mitigated by using multiuser receivers. Multiuser receivers in STC CDMA systems have been shown to provide significant improvements in spectral efficiency. However, despite the great promises, these receivers have not been widely deployed in commercial communication systems. This lack of commercial interest is partly because of issues related to receiver complexity.

Thus using analysis, reasonable approximations and simulation, this work attempts to realize a lower complexity multiuser receiver for STTC DS-CDMA systems. The complexity is reduced by dividing the users into groups having smaller trellises. The simulation results show that the proposed receiver performs close to the optimal receiver even at high cross correlations of the spreading sequences with a reduced complexity.

Keywords: CDMA, Spreading Sequences, Space-Time Coding, Fading Channels, Multiple Access Interference multiuser detection.

Chapter 1

Introduction

1.1 Motivation

The transmission of information has become a key feature of the modern way of life. The communication systems and information networks of the future change the way people work, shop and spend their leisure time. The demand for wireless communication services has increased rapidly and this trend is expected to continue.

Requirements on the capacity of communication systems are posed in terms of the number of users a system can serve simultaneously. In more appropriate words, as much information as possible should be transferred. This goal can be achieved by designing efficient source coding methods to compress the non-systematic redundancies in the information, by using smaller cells in cellular systems, by utilizing spatial and temporal signal processing techniques and by designing efficient multiple access systems and transceivers for them. For wireless cellular systems, restricted bandwidth allocations coupled with a potentially large pool of users limit the bandwidth available to each user. The success of wireless cellular and personal communication services thus depends on the development of new bandwidth efficient data transmission schemes.

Non-orthogonal multiple access can potentially offer better system capacity than orthogonal schemes [4, 38]. Space-time coding techniques can be used in non-orthogonal multiple access systems to further increase system capacity and to mitigate multi-path fading effects. The use Space-time coding techniques on the other hand enhances the undesired effect of multiuser interference because of the multiple antennas at the transmitter. The interference problem should be taken care at the receiver. Obtaining the maximum Shannon capacity requires joint decoding of the data of all users. The conventional single user receiver is not optimal in non-orthogonal multiple access systems, since it considers the multiple access interference as additive noise. An alternative to the conventional single user receiver is to apply a receiver designed to take the multiple access system interference into consideration. Such receivers are called multiuser receivers [46].

In space-time trellis coded direct sequence code division multiple access systems, the maximum likelihood multiuser sequence estimator which can be implemented with a viterbi type algorithm is optimal in the sense that it selects the most probable sequence of bits. Unfortunately, the optimal multiuser detector is prohibitively complex to implement for many practical applications. This receiver complexity serves as a motivation to devise a reduced complexity multiuser receiver for space-time coded direct sequence code division multiple access systems.

1.2 Review of Related Works

In 1986, Verdu suggested multiuser detection techniques as a potential method to reduce multiple access interference in multiple access systems. He proposed the idea that detection of code division multiple access signals should exploit the structure inherent in the multiple access interference. His works showed that multiuser detectors, which are able to reduce multiple access interference, outperform the conventional single user matched filter receiver.

John G.Proakis studied decorrelator type receivers [40]. This is a simple scheme, where the received signal can be fed into a bank of matched filters, each matched to the signature sequence of a different user. The decorrelator detector suppresses the interference by finding a linear transformation such that the transmitted symbols for all users are completely recovered in the absence of noise. When this detector eliminates the multiuser interference, it thereby increases the noise power [1, 30] resulting in poor performance in noise-limited environments. Further, decorrelator detectors require correlation matrix inversion which may be difficult to achieve in real time.

A multiuser detector that is closely related to the decorrelating detector is the linear minimum mean square error detector. This detector provides an alternative approach to select a transform to reduce the total mean square error at the receiver output, thus alleviating the problem of the noise enhancement. The linear detector chooses the weight matrix to minimize the mean square error between the transmitted user signals and the outputs of the linear transformation. The error rate performance of the minimum mean square error detector is similar to that for the decorrelator detector when the noise level is low. For low multiuser interference, the minimum mean square error detector results in a smaller noise enhancement compared with the decorrelator detector.

Mincheol Park studied multiuser receivers employing interference cancellation due to their relatively low complexity [13]. These receivers are based on successive interference cancellation and parallel interference cancellation techniques. In successive interference cancellation, if a decision has been made about an interferer's bit, then that interfering signal can be recreated at the receiver and subtracted from the received waveform. If the decision was correct, this would cancel the interfering signal otherwise, it will increase the contribution of the interferer. In parallel interference cancellation the n^{th} stage of the detector uses decisions of the $(n\text{-}1)^{th}$ stage to cancel the multiple access interference present in the received signal.

Shen and C.Reed in their thesis studied iterative receivers for turbo coded code division multiple access [19, 25]. The performance of these receivers is improved at the expense of a decoding delay that is incurred due to the iteration. There are, however applications wherein the decoding delays are very much undesired. For example, for voice transmission, the maximum tolerable delay is about 20ms and 10ms-80ms for real time data services [18]. These decoding delay restrictions result in limitations in performance. Non-iterative decoding schemes avoid this tradeoff and they provide an alternative to iterative receivers in situations where the decoding delay is to be limited.

1.3 Space-Time Multiuser Receivers

One of the basic problems of multiple access systems is multiuser interference. The problem becomes increasingly significant if the power level of the users is dissimilar which is called near-far problem. Through proper signal processing, the effect of interfering signals can be minimized [46].

In direct sequence code division multiple access systems, all the users transmit information simultaneously in the same frequency band. One approach to detect a user signal is to treat the channel as a single user channel and all other users as additive noise. Each user is assigned a pseudorandom signature sequence and the resulting multiuser interference is approximated as additive white Gaussian noise. This receiver is called the rake receiver and has been deployed in commercial code division multiple access systems [34, 39]. A single user receiver is not optimal in multiple access systems since it does not exploit any knowledge of the other user interference available at the base station. In contrast to single user receivers, multiuser receivers jointly decode all the received signals. Although the optimal multiuser receiver promises a much greater spectral efficiency than the single user receiver, it has significantly higher complexity [22]. For this reason, there have been numerous researches on suboptimal multiuser receivers that tradeoff performance for complexity.

Multiple antennas at the transmitter provide an additional spatial dimension for processing the signals received from different users. Space-time processing involves the joint processing of the received signals in the spatial and temporal dimensions. Space-time block codes using orthogonal designs were proposed for wireless communications over fading channels with multiple transmit antennas [45]. These codes require no knowledge of the channel state at the transmitter and linear processing at the receiver achieves maximum likelihood decoding. Space-time trellis codes are even much better than space-time block codes for they provide both coding gain and diversity gain [45]. These advantages of space-time trellis codes, however, are offset by complexities in code design and receiver implementation. In this thesis, a low complexity receiver structure for space-time trellis coded direct sequence code division multiple access systems has been studied. Performance and complexity tradeoffs have been analyzed. Simulations are used to study the performance of the proposed low complexity receiver. The performance of the proposed receiver has been simulated and compared with the optimal receiver, the decorrelator detector and the conventional single user detector for a few users.

1.4 Aim and Outline of the Thesis

In spite of the major research efforts invested in multiuser detection techniques, several practical as well as theoretical open problems still exit in the field of multiuser receivers. The aim of this thesis is to develop a practical multiuser detector for mobile communication systems with slowly varying Rayleigh fading channels and to analyze implementation complexities. The emphasis is restricted to the uplink of synchronous space-time trellis coded direct sequence code division multiple access systems, where users transmit simultaneously and are received by one centralized receiver.

This thesis is organized as follows: chapter 2 gives a brief overview of multiple access techniques. Chapter 3 describes wireless channels and diversity techniques which are employed to combat fading effects and increase capacity. Chapter 4 is concerned with the modeling of space-time trellis coded code division multiple access systems and the design of a reduced complexity multiuser receive for these coded systems. Chapter 5 presents the simulation results of this thesis work and finally chapter 6 gives conclusions and directions for future work.

Chapter 2

Multiple Access Techniques

2.1 Introduction

Multiple access refers to a technique used to share a common communication channel among multiple users. There are several ways in which multiple users can send information through the communication channel to the receiver. One simple method is to subdivide the available channel bandwidth into a number of non-overlapping sub-channels for each user. This method is generally called frequency division multiple access (FDMA).

Another method for creating multiple sub-channels for multiple access is to subdivide the duration T_f called frame duration into, say K non-overlapping sub-intervals each of which is duration T_f / K . Then a user who wishes to transmit information is assigned to a particular time slot within each frame. This multiple access technique is called time division multiple access (TDMA).

An alternative to FDMA and TDMA is to allow more than one user to share a channel or sub-channel by the use of direct sequence spread spectrum signals. This multiple access method is called code division multiple access (CMDA). More advanced multiple access techniques include space division multiple access (SDMA) and polarization division multiple access (PDMA) [4, 33].

In the following sections, FDMA, TDMA and CDMA will be reviewed with emphasis on CDMA and the spreading sequences which are user specific codes used in spread spectrum systems.

2.2 Frequency Division Multiple Access (FDMA) [4, 12, 38]

The oldest multiple access technique is frequency division multiple access in which each user signature waveform occupies its own frequency band and the receiver can separate the users' signals by simple bandpass filtering. In this technique, disjoint sub-bands of frequencies are allocated to the different users on a continuous time basis. In order to reduce interference between users allocated adjacent channel bands, guard bands which act as buffer zones are used. The guard bands are necessary because of the impossibility of achieving ideal filtering for separating the different users. FDMA is a simple scheme and applicable to both analog and digital modulation. It is not; however, very flexible for providing variable bit rates, which is an important requirement in current and future communication services.

2.3 Time Division Multiple Access (TDMA) [4, 12, 38]

The introduction of digital modulation enabled the appearance of time division multiple access (TDMA). In this technique, each user is allocated the full spectral occupancy of the channel but only for short periods of time called slot time. To reduce interference, guard times are inserted between the assigned time slots.

Interference in TDMA arises due to system imperfection, especially in synchronization schemes. TDMA is relatively simple to implement and it is very flexible for providing variable bit rates. Increasing the bit rate can be achieved by assigning to a user more transmission intervals.

However, the transmission of all the users must be exactly synchronized to each other. Due to simple implementation of more complicated modulation schemes in TDMA than in FDMA, the capacity of TDMA is significantly higher than that of FDMA systems.

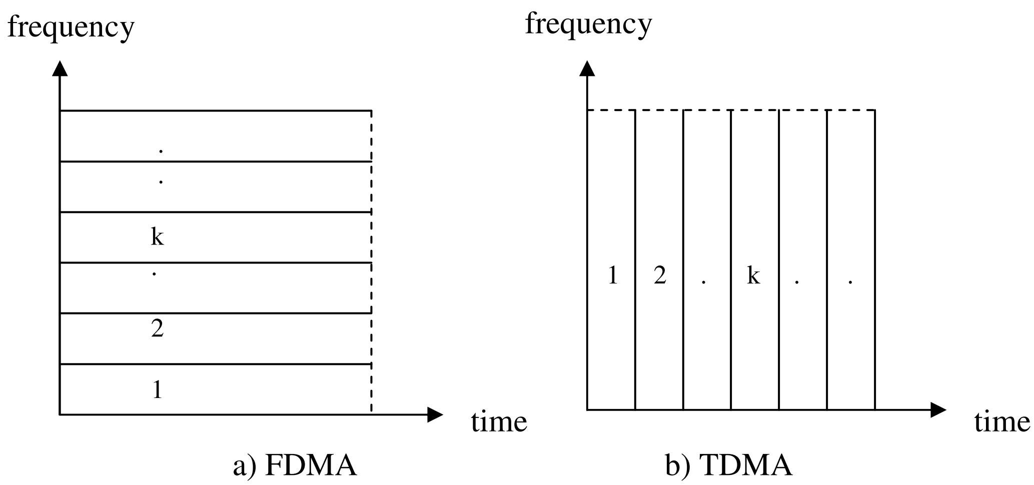


Figure 2.1 Resource allocation schemes in FDMA and TDMA

2.4 Code Division Multiple Access (CDMA) [5, 11, 32, 40]

In FDMA and TDMA, the frequency bands and time slots are allocated to users very rigidly. Because of this rigid allocation of frequency bands and time slots, FDMA and TDMA are not efficient multiple access techniques if the data from the users accessing the network is bursty in nature [40]. This inflexibility of FDMA and TDMA has resulted in the development of a new system based on uncoordinated spread spectrum concept.

Spread spectrum techniques for communication systems with anti-jamming and low probability of undesired interception capabilities lead to the idea of code division multiple access (CDMA) [30].

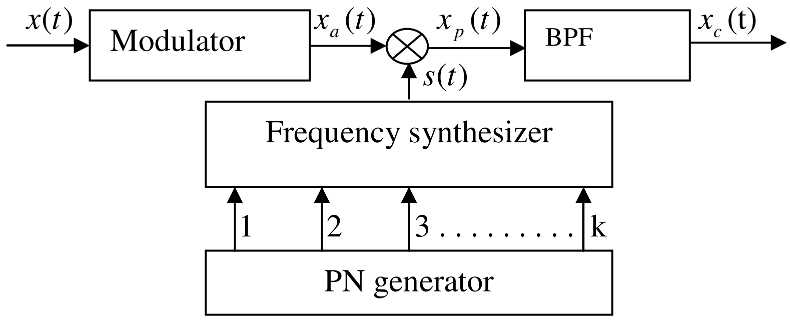
CDMA can be implemented in numerous ways including frequency hopping spread spectrum (FHSS), time hopping spread spectrum (THSS) or direct sequence spread spectrum (DSSS) techniques as well as multi-carrier (MC) techniques. Hybrid CDMA systems based on combining all or some of these techniques are possible [4].

In CDMA, the narrow band message is multiplied by a large bandwidth signal which is a pseudo noise (PN) sequence. All users in a CDMA system use the same frequency band and transmit simultaneously [32, 40]. The transmitted signal is recovered by correlating the received signal with the PN code used at the transmitter. The CDMA signature waveforms usually have significantly large bandwidth than FDMA and TDMA waveforms. An important attribute of CDMA signature waveforms is that they provide protection against externally generated interfering signals and fading, which are impairments of mobile radio channels [34].

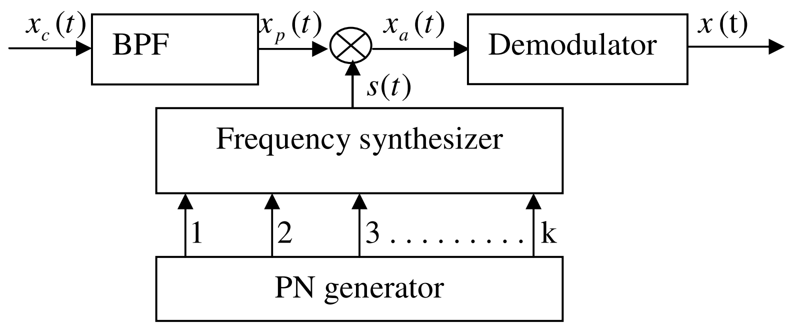
2.4.1 Frequency Hopping Spread Spectrum (FHSS)

In frequency hopping spread spectrum (FHSS) communication systems, the available channel bandwidth is subdivided into a large number of contiguous frequency bands [28, 33]. In any signaling interval the transmitted signal occupies one or more of the available frequency slots. The selection of the frequency slots in each signaling interval is made randomly according to the output from a Pseudo noise generator. The center frequency of the signal changes from one hop to the next in discrete increments in a pattern generated by the code sequence changing from one sub-band to the other.

In FHSS, the message is modulated using some carrier frequency f_c. The modulated message is then mixed with the output of a frequency synthesizer. The frequency synthesizer output is one of the 2^k values, where k equals the number of outputs from the pseudo noise (PN) generator. The bandpass filter (BPF) selects the sum term from the mixer for transmission on the channel.



a) Transmitter



b) Receiver

Figure 2.2 Frequency hopping spread spectrum

2.4.2 Direct Sequence Spread Spectrum (DSSS)

Direct sequence spread spectrum (DSSS) which is also called direct sequence code division multiple access (DS-CDMA) is the most common version of spread spectrum techniques in use today because of its simplicity and ease of practical implementation. In DSSS, the data signal is modulated by the PN sequence and the code bit rate is much greater than the information signal bit rate.

In DSSS, users are separated neither in the time domain nor in the frequency domain but all signature waveforms occupy the whole allocated frequency band for the transmission at all times. The data of the users can be separated at the receiver since the waveforms in DSSS are formed by spreading sequences which are unique to each user.

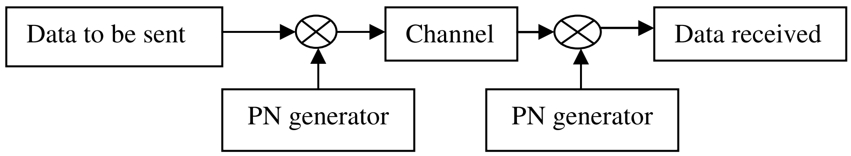


Figure 2.3 Block diagram of a DSSS system

Traditional FDMA and TDMA are designed to be orthogonal [30]. DSSS on the other hand can be designed to be either orthogonal or non-orthogonal. In the case where the spreading sequences are designed to be orthogonal, if the signals of all the users arrive at the receiver with the same time delay and the transmission medium does not cause time dispersions, the signature waveforms appear as orthogonal at the receiver.

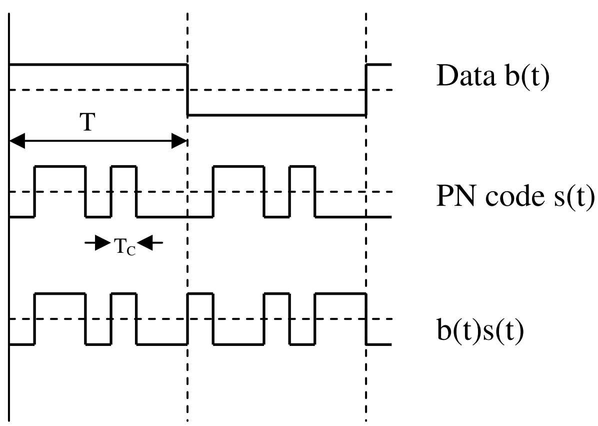


Figure 2.4 Direct spread signal

With unequal time offset, which is often the case in real applications, the signature waveforms lose orthogonality at the receiver [35, 37]. The spreading sequences may also be designed to be non-orthogonal and this non-orthogonal design is more flexible than orthogonal multiple access techniques, since there is no hard limit on the number of users as there is in orthogonal multiple access techniques due to the finite dimensionality of the signal space [4, 20]. The gain in SNR obtained by the use of spread spectrum is measured by the Processing gain (PG) in decibels [15].

$$PG = 10\log(W_c / W_b) \qquad (2.1)$$

Where, W_c is the spread spectrum bandwidth and W_b is the bandwidth of the data signal

2.4.3 Spreading Sequences

A spreading sequence or a pseudorandom (PN) sequence is a bit stream of zeros and ones occurring randomly with some unique properties. This code bits are called chips. The spread spectrum CDMA scheme is one in which the transmitted signal is spread over a wide frequency band, much wider than the minimum bandwidth required to transmit the information being sent.

In general, the spread spectrum signals are commonly used for [34, 40]

- Combating or suppressing the effects of interference due to jamming and interference arising from other users of the channel, and self-interference due to multi-path propagation.
- Hiding a signal by transmitting it at low power density, thus making it difficult for an unintended listener to detect it in the presence of background noise.
- Achieving message privacy in the presence of other listeners.
- Sharing the limited frequency resource to satisfy the demands for increasing capacity in the public network.

The desired properties of PN sequences depend on the application. The main emphasis in this study is on the control of the interference arising from other users of the channel.

In DS-CDMA, a user signal is multiplied by a spreading code sequence at the transmitter. This sequence must be known by the receiver to be able to realize dispreading. To be used in real systems, the sequence should be constructed from a finite number of randomly selected parameters. The receiver dispreading operation of DS-CDMA system is a correlation operation with the spreading code of the desired transmitter.

The specific amount of interference from a user employing a different spreading code is related to the cross correlation between the spreading codes [33]. Orthogonal codes have zero cross correlation levels, thus multiuser interference can be totally eliminated. However, it is usually not the case in the sense that perfect orthogonality is hard to achieve in a practical DS-CDMA system. Even if the sequences are designed to be orthogonal at the transmitter, it is difficult to maintain this orthogonality at the receiver because of the different time offsets [34]. Hence, low cross correlation spreading codes, which also exhibit other more desired features mentioned in section 2.4.4 are sought.

Random codes here refer to randomly selected sequences that are completely chosen at random while preserving the condition of uniqueness. An ideal spreading code would be an infinite sequence of equally likely random binary digits. Unfortunately, the use of an infinite random sequence implies infinite storage at both the transmitter and receiver [34]. Hence, periodic PN codes are always employed. The disadvantage of using finite length random codes is that they have poor cross correlation properties. Noise-like wideband spread spectrum signals are generated using pseudo noise or pseudorandom spreading code sequences, which are so called because they exhibit the properties of random numbers. However, if the code sequences were truly random, then nobody, including the intended receiver, could access the channel. Thus, PN sequences should be relatively easy to generate and should appear as random noise to everyone else, except to the transmitter and the intended receiver.

The maximum-length shift register sequences, or m-sequences for short, are the most widely known PN sequences [23, 28, 40]. An m-sequence has length $N=2^m - 1$ and it is generated by m-stage shift register with linear feedback. The sequence is periodic with period N. Each period of the sequence contains 2^{m-1} ones and $2^{m-1} - 1$ zeros. In direct sequence spread spectrum, the binary sequence with elements {0, 1} is mapped to a sequence with elements {-1, 1}.

The feedback shift register used to generate m-sequences consists of an ordinary shift register made up of two state flip flops and a logic circuit that is interconnected to form a multiple feedback [15].

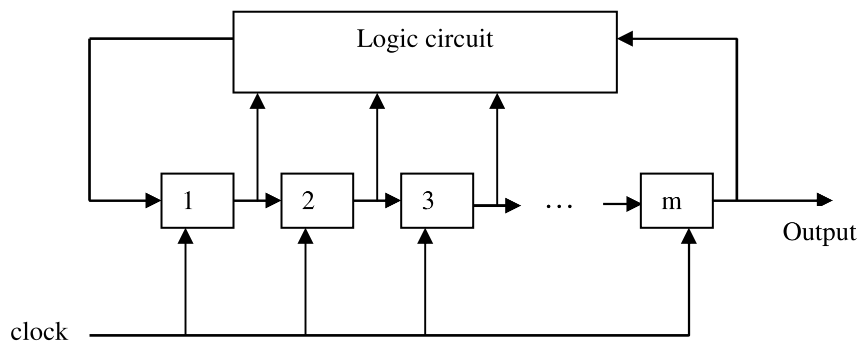


Figure 2.5 Feedback shift register

The shift register is regulated by a single timing clock. At each pulse of the clock, the state of the memory stages is shifted to the next one. With each clock pulse, the logic circuit computes a Boolean function of the states of the memory registers. The result is fed back as an input to the first memory register, thereby preventing the memory register from being empty.

If $s(n)$ for $n = 0, 1, 2, \cdots N-1$, is a PN sequence consisting of N chips that take values {-1, 1}, the signature waveform resulting from this sequence may be expressed as in [40]

$$s(t) = \sum_{n=0}^{N-1} s(n)\, p(t - nT_c), \quad 0 \le t \le T \tag{2.2}$$

Where, $p(t)$ is a pulse of duration T_c which is called the chip interval.

If $s(t)$ represents the waveform resulting from a PN sequence of length N, then the period of the waveform is $T = NT_c$. The autocorrelation function of the waveform $s(t)$ is defined as

$$\rho(\tau) = \tfrac{1}{T}\int_0^{\tau} s(t)s(t-\tau)\,dt \tag{2.3}$$

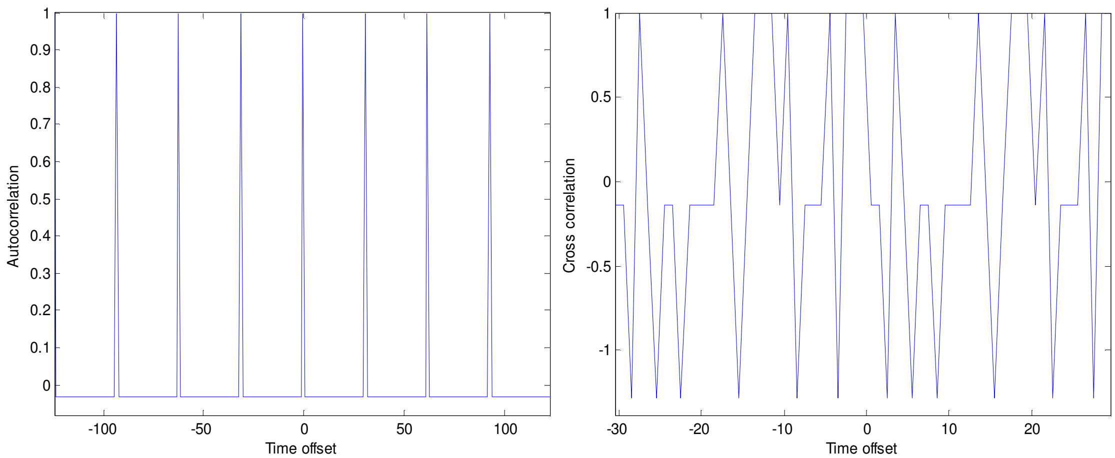
For any two waveforms $s_i(t)$ and $s_j(t)$ resulting from two distinct sequences, the cross correlation between them is defined as [33, 40]

$$\rho_{ij}(\tau) = \begin{cases} \tfrac{1}{T}\int_{\tau_c}^{T} s_i(t)s_j(t-\tau)\,dt\,, & i < j \\ \tfrac{1}{T}\int_0^{\tau} s_i(t+T+\tau)s_j(t)\,dt\,, & i > j \end{cases} \tag{2.4}$$

This definition is for a general asynchronous CDMA and for synchronous transmission, we need only $\rho_{ij}(0)$.

In DS-CDMA, the cross correlation property of the spreading sequences is as important as the autocorrelation properties. For m-sequences, the peak value of the cross correlation is a very large percentage of the autocorrelation function as shown in figure 2.6b. Such high values for the cross correlations are undesired in CDMA applications.

Spreading sequences that have better periodic cross correlation properties than the m-sequences are called gold sequences [23, 33]. Gold sequences are generated from a pair of preferred m-length sequences. From a pair of preferred m-length sequences say $S_1(n) = [s_1(0) \quad s_1(1) \quad s_1(2) \quad \cdots \quad s_1(N-1)]$ and $S_2(n) = [s_2(0) \quad s_2(1) \quad s_2(2) \quad \cdots \quad s_2(N-1)]$,



a) Autocorrelation b) Cross correlation

Figure 2.6 Autocorrelation and cross correlation of m-sequences

Gold sequences are constructed by taking the modulo-2 sum of $S_1(n)$ with the cyclically shifted version of $S_2(n)$. The generation of Gold sequence from two preferred m-length sequences described by the parity polynomials $h_1(x)$ and $h_2(x)$ given below is shown in figure 2.7.

$$h_1(x) = x^5 + x^3 + 1 \text{ and } h_2(x) = x^5 + x^4 + x^3 + x + 1$$

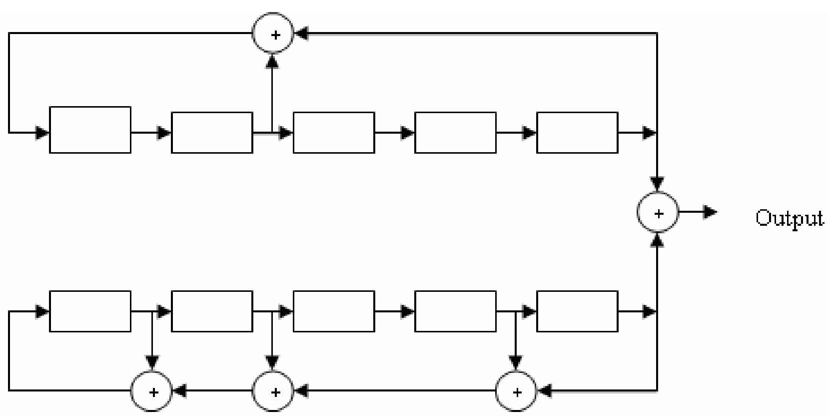
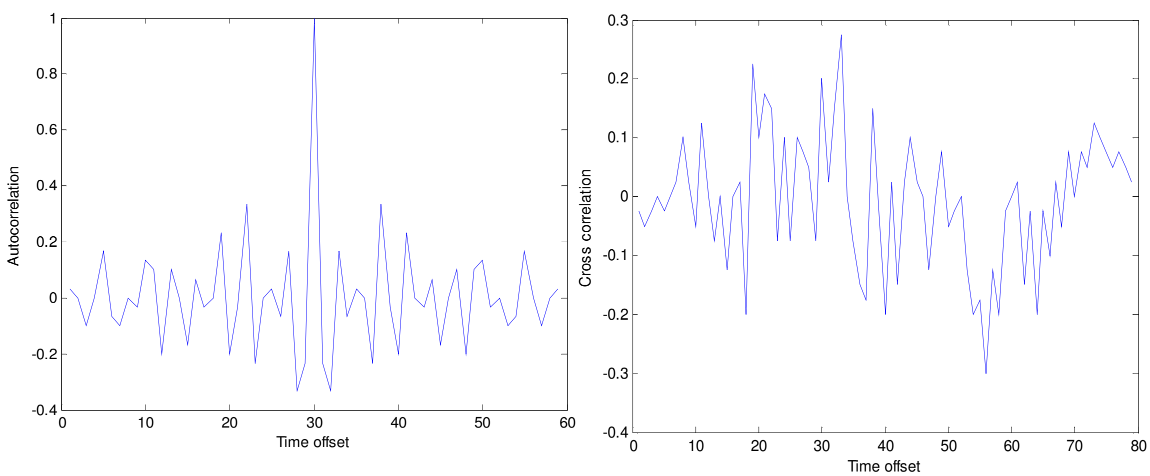


Figure 2.7 Generation of Gold sequences of length 31

Gold sequences have good cross correlation properties and they are used especially for CDMA systems to reduce the interference caused by multiple users. The autocorrelation and cross correlation between two Gold sequences of length 31 is shown in Figure 2.8.

Gold codes help to maintain separation among different users at the receiver. In the uplink communication of CDMA, Gold codes are used to spread the signals from different mobile terminals. Each mobile terminal has a distinct spreading code. The receiver should know the spreading code for each terminal in advance.



a) Autocorrelation b) Cross correlation

Figure 2.8 Autocorrelation and cross correlation of Gold sequences

2.4.4 Properties of Spreading Sequences [5, 15, 40]

The spreading sequences that are used in CDMA applications have the following properties

- The number of ones generated is one more than the number of zeros.

- Half of the run lengths are unity, one-quarter are of length two, one-eighth are of length three and so on. A run is a sequence of a single type of digit.
- The autocorrelation function of PN sequences is similar to the correlation properties of random noise in that there is a single autocorrelation peak.
- If the sequence is shifted by any non-zero number of elements, the resulting sequence will be another sequence
- Every possible state of the shift register exists at some time in a complete sequence.
- The modulo-2 sum of an m-sequence and any shift of the same sequence is another sequence.
- The statistical properties of ones and zeros are well defined and always the same.

These properties among others make the spreading sequences suitable for CDMA applications. For example, the first property above is essential for equal spreading of the energy over the whole frequency band.

Chapter 3

Wireless Channels and Diversity Techniques

3.1 Introduction

When an electrical signal is propagating through the open air, it will be corrupted by additive noise. Moreover, it will suffer from other channel impairments due to the physical environment characterizing the channel and interfering signals in multiple access systems. In such an environment, a transmitted signal takes different propagation paths, with different attenuations, phase shifts and distortion levels due to the reflecting objects encountered along the way. At the receiver, constructive or destructive interference may occur which results in signal amplitude fluctuation. This fluctuation of the signal amplitude is called multi-path fading. Furthermore, transmitter or receiver mobility induces a time varying nature to this phenomenon, which makes the communication channel more challenging [8].

In wireless communication channels, the multi-path fading effect is a major challenge in providing high data rate and spectrally efficient communication links. An effective technique to mitigate multi-path fading is transmitter power control. One major drawback of this method is that the transmitter must be aware of the channel conditions as observed at the receiver. In general, this information must be fed back to the transmitter which results in performance degradation both in terms of rate and complexity. The required transmitter power may not even be available due to power limitations.

Diversity techniques provide an effective alternative. Previously, diversity techniques such as temporal diversity, frequency diversity and antenna diversity have been used to deal with this problem. The performance of these diversity techniques can further be increased by offering the diversity through temporal and spatial dimensions and this concept has brought the idea of space-time coding. Space-time coding is considered to be a promising solution to achieve spectrally efficient communication links in mobile channels.

3.2 Characterization of Wireless Channels

The determination of optimal modulation and demodulation techniques depends on the accurate characterization of the communication channel. Impairments in the propagation channel have the effect of disturbing the information carried by the transmitted signal. The major channel disturbances in wireless communication links can be a combination of additive noise, multiple access interference and fading.

3.2.1 Noise

When an information carrying signal propagates through a wireless channel, it will invariably be corrupted by additive noise. This noise term may arise from thermal agitation of the atmosphere, electromagnetic interference, lightening, shot noise and so on. Because these sources are many and act independently of each other, the nature of the noise can be considered as gassian and this is true from the central limit theorem. If $x(t)$ is the transmitted signal in a noisy channel, the signal at the front end of the receiver denoted by $y(t)$ is given by

$$y(t) = x(t) + n(t) \tag{3.1}$$

Where, $n(t)$ is the baseband representation of the noise with a two sided power spectral density of $\frac{N_0}{2}$.

3.2.2 Multiple Access Interference

In cellular FDMA and TDMA, multiple access interference is not a major problem because the frequency bands and the timeslots do not overlap [17]. On the other hand, CDMA is a multiple access interference limited application [31]. Multiple access interference in CDMA is because of the deviation of the spreading sequences from perfect orthogonality. Multiuser receivers in CDMA systems use the structural property of this multiple access interference for a better decoding operation [33]. In cellular systems, two kinds of interference exist. The first one comes from other mobile stations within the cell (in the uplink case) also known as intra- cell interference. The second one comes from all the mobile stations in other cells (inter-cell interference). This total interference limits the capacity and performance of CDMA because as the number of users increases so does the interference, thereby limiting the number of users that are allowed in a cell to guarantee a certain transmission quality.

3.2.3 Channel Fading

While traveling over a wireless communication channel, signals are affected by several phenomena, one among these is fading which reveals itself as an attenuation of the average signal power and it is induced by terrain obstacles like hills and buildings encountered between transmitter and receiver [8]. Fading channels can generally be classified according to their time and frequency characteristics. The behavior of fading channels in time can be one of the following:

Fast fading: the time over which the channel characteristics can be considered constant known as the coherence time is considerably smaller than the time used to transmit the signal. Therefore, the transmitted data will be unequally altered by the channel and suffer from distortion.

Slow fading: the coherence time of the channel is larger than the time used to transmit the signal. Therefore, the whole signal will be equally affected by the channel and will not suffer from the distortion introduced as in the fast fading case.

Similarly, its frequency behavior can be classified into two main categories:

Frequency Selective fading: the bandwidth over which the channel characteristics can be considered constant known as the coherence bandwidth is considerably smaller than the bandwidth occupied by the transmitted signal. Therefore, the spectrum of the transmitted data we wish to transmit will be unequally altered by the channel causing the original signal to be spread in time which might introduce inter-symbol interference (ISI).

Frequency non-selective fading: the coherence bandwidth of the channel is larger than the bandwidth of the transmitted signal. Therefore, all signal frequency components are equally affected by the channel and no ISI is present. This channel model is also known as frequency flat fading.

The received signal at the output of a wireless channel is the superposition of multiple noisy signals arriving from different paths. These paths generally have different delays and attenuations and if the delays and attenuations are constant over time, the channel is a static multi-path channel. The impulse response of the channel in this case can be characterized by a time-invariant function of the delay. However, the characteristics of the different paths vary with time in a realistic channel. Thus, the impulse response has to be a function of time. This means, the impulse response h(τ, t) is time variant. If h(τ, t) has a zero mean, then the magnitude of h(τ, t) has a Rayleigh probability distribution given by

$$p(r) = \frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right) \tag{3.2}$$

Where, σ^2 is the total power of the multi-path signal. Otherwise, if h(τ, t) has a non-zero mean, which implies the presence of a significant line of sight component, then the magnitude of h(τ, t) has a Ricean probability distribution given by:

$$p(r) = \frac{r}{\sigma^2} \exp\left(-\frac{r^2 + s^2}{2\sigma^2}\right) I_0\left(\frac{rs}{\sigma^2}\right) \quad (3.3)$$

Where, s^2 is the power of the line of sight component, and I_0 denotes the zero order modified Bessel function of the first kind.

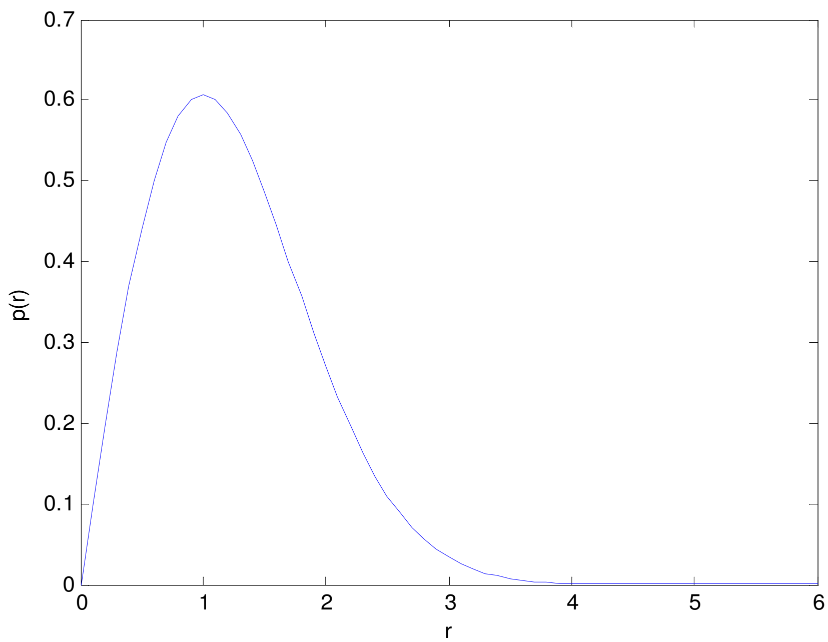


Figure 3.1 Rayleigh distribution

3.3 Diversity Techniques [3, 14]

Signal power in a wireless channel fluctuates with time, frequency and space. When the signal power drops dramatically, the channel is said to be in fade. This power penalty can in principle be completely overcome using diversity techniques. Diversity is used in wireless systems to combat fading and increase spectral efficiency. The basic principle behind diversity techniques is to provide the receiver with several replicas of the transmitted signal over independently fading links or diversity branches. As the number of diversity branches increases, the probability that at any instant of time one or more branches are not in fade increases.

Diversity is available in the form of time, frequency or antenna diversity. The use of time or frequency diversity often incurs a penalty in data rate due to the utilization of time or bandwidth to introduce redundancy. Antenna diversity does not incur a penalty in data rate while providing the array gain advantage.

3.3.1 Time Diversity

In time diversity, replicas of the transmitted signal are provided across time by means of channel coding and time interleaving strategies. Ideally, the separation between the replicas exceeds the coherence time in order for the replicas to be independent [6]. Therefore, the channel must provide sufficient variations in time for this form of diversity to be useful. One of the drawbacks of time diversity is that due to the redundancy introduced in the time domain, there is a loss in bandwidth efficiency [3, 6].

3.3.2 Frequency Diversity

In frequency diversity, replicas of the transmitted waveform are provided on different frequency carriers. Assuming that the coherence bandwidth of the channel is small compared to the bandwidth of the signal, the replicas will suffer independent fading providing frequency diversity. Like time diversity, frequency diversity induces a loss in bandwidth efficiency due to the redundancy introduced in the frequency domain [3, 6].

3.3.3 Antenna Diversity

This form of diversity is provided across different antennas at the receiver or at the transmitter. In antenna diversity, the separation between the antennas should be large enough to assure independent fading of the different received signals. The separation requirements between the antennas vary with antenna height, propagation environment and frequency. Typically, a separation of a few wavelengths is enough to obtain uncorrelated signals [3, 6].

Unlike time diversity and frequency diversity, antenna diversity doesn't induce a loss in bandwidth efficiency. This property is very attractive for high data rate wireless communication.

Antenna diversity schemes can be classified by looking at whether the diversity is applied at the transmitter (transmit diversity) or at the receiver (receive diversity). The idea behind transmit diversity is to introduce controlled redundancies at the transmitter which can be exploited at the receiver. Receive diversity schemes may consist of combining different diversity branches in order to maximize the quality of the received signal. The effectiveness of any diversity technique relies on the availability of independent faded replicas of the transmitted signal at the receiver. A key concept here is that of diversity order which is defined as the number of uncorrelated diversity branches available at the transmitter or at the receiver. The performance of the system depends on the ability of the receiver to properly combine these diversity branches in order to maximize the signal quality.

3.4 Space-Time Coding (STC)

Previously, multi-path fading in a wireless system was mostly dealt with by time diversity, frequency diversity and antenna diversity with receive antenna diversity being the most widely applied technique.

Space-time coding for multiple transmit and/or receive antenna systems has recently attracted considerable attention in the wireless communications area. It has been shown theoretically that for certain channels, the capacity of a multiple antenna system increases linearly with the minimum number of antennas [3]. This increase in capacity is very attractive because it comes at no extra bandwidth or power consumption.

Space-time coding is one of the most efficient diversity schemes which combines channel coding, modulation, transmit antennas diversity and optional receive antenna diversity to achieve high data rate transmission and to combat fading. Space-time coding is essentially a joint design of coding, modulation and antenna diversity. In space-time coding, there should be two or more number of antennas at the transmitter and/or at the receiver, one or more antennas can be exploited. There are two main types of space-time codes; these are space-time block codes and space-time trellis codes. Space-time block codes operate on a block of input symbols and produce a matrix output whose columns represent time and rows represent antennas. In contrast to single antenna block codes for AWGN channels, space-time block codes do not generally provide coding gain. Their main feature is the provision of full diversity with a very simple linear decoding scheme. On the other hand, space-time trellis codes map one input symbol at a time to N output symbols, where N is the number of transmit antennas. The encoder in space-time trellis coding is a set of convolutional encoders. Since a convolutional encoder has memory, the vector codewords are correlated in time [44].

3.4.1 Space-Time Block Coding (STBC)

Alamouti, who discovered a remarkable scheme for transmission using two antennas, initiated the field of space-time block codes. With the advent of the Alamouti's scheme for two transmit antenna systems, there arose a natural question whether it is possible to generalize this simple scheme to more general systems having multiple antennas. The search of Alamouti type schemes for an arbitrary number of transmit antennas has resulted in the development of a general theory of space-time block codes [1]. Since this thesis work is related to space-time trellis codes, only the simplest scheme of space-time block coding for two transmit antennas, which is the Alamouti scheme is reviewed and the general theory of space-time block coding for more than two transmit antennas is not considered.

Alamouti proposed a simple transmit diversity scheme for two transmitting antenna systems [1, 16]. At the receiver, only linear processing is needed to decode Alamouti's scheme. This scheme transmits consecutive pairs of complex input symbols $x(i)$ and $x(i+T)$. During the first symbol period, the symbols $x(i)$ and $x(i+T)$ are transmitted from the first and second antennas respectively. During the second symbol period, the complex conjugate symbols $-x^*(i+T)$ and $x^*(i)$ are transmitted from the first and second antennas respectively.

If we assume a quasi-static Rayleigh fading channel, the fading coefficients are constant for the duration of the two consecutive symbol times. Denoting the received signals corresponding to the symbol times i and $i+T$ by $r(i)$ and $r(i+T)$ respectively, and let h_1 be the path gain from the first antenna to the receiver and h_2 be the path gain from the second antenna to the receiver, then [21]

$$r(i) = h_1 x(i) + h_2 x(i+T) + n(i) \tag{3.4a}$$

$$r(i+T) = -h_1 x^*(i+T) + h_2 x^*(i) + n(i+T) \tag{3.4b}$$

Writing this in matrix form

$$y(i) = \begin{bmatrix} r(i) \\ r^*(i+T) \end{bmatrix} = \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix} \begin{bmatrix} x(i) \\ x(i+T) \end{bmatrix} + \begin{bmatrix} n(i) \\ n^*(i+T) \end{bmatrix} = HX(i) + \eta(i) \tag{3.5}$$

Where $H = \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix}$, $X = \begin{bmatrix} x(i) \\ x(i+T) \end{bmatrix}$, $\eta(i) = \begin{bmatrix} n(i) \\ n^*(i+T) \end{bmatrix}$

$n(i)$ and $n(i+T)$ are independent gaussian noise samples at instants i and $i+T$.

For a channel that fades according to the Rayleigh probability distribution, if the transmitting antennas are separated by a distance greater than half the wavelength of the carrier signal, the columns of the channel matrix H are independent as a result

$$z(i) = (H)^H y(i) = \left(|h_1|^2 + |h_2|^2\right) \begin{bmatrix} x(i) \\ x(i+T) \end{bmatrix} + (H)^H \eta(i) \tag{3.6}$$

Where, $(.)^H$ represents the Hermitian operator.

Due to the orthogonality of the columns of *H*, it follows that the components of the noise vector are still independent and gaussian. Thus, the joint maximum likelihood decoding of the symbols $x(i)$ and $x(i+T)$ can be decoupled into two simple decoding problems.

3.4.2 Space-Time Trellis Coding (STTC)

Space-time block codes do not provide coding gain unless they are concatenated with an external channel encoder, but the encoding and decoding operations for these codes can be implemented easily [45]. Space-time trellis codes perform more complex operations and they are better in performance than space-time block codes since they provide both coding gain and diversity gain.

Space-time trellis codes use several convolutional coding and multiple antennas to achieve correlation in temporal and spatial dimensions [7]. The receiver performs a kind of maximum likelihood sequence estimation that can be implemented with a viterbi type algorithm to decode the received signals.

Space-time trellis codes have shown considerable performance gains for wireless communications in fading channels at the expense of a rising decoding effort and they are defined by the number of transmit antennas, by the associated state diagram and by the modulation scheme employed. In space-time trellis coding, the channel encoder is actually a convolutional encoder in which the binary outputs are mapped into numeric alphabets that are elements of the set {0, 1, 2…, M-1} for *M*-PSK symbol mapping.

3.4.2.1 Delay Diversity Scheme

Space-time trellis coding is a generalization of delay diversity scheme [1, 6, 29]. For two transmit antenna delay diversity for example, each symbol is transmitted by both antennas with a delay of one symbol period in between them. It is called delay diversity because delayed replicas of the same information-bearing signals are transmitted through different antennas as shown in figure 3.2.

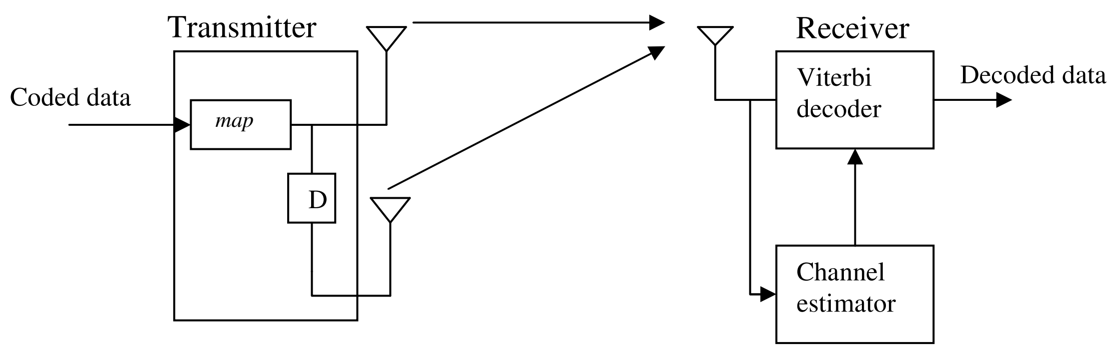
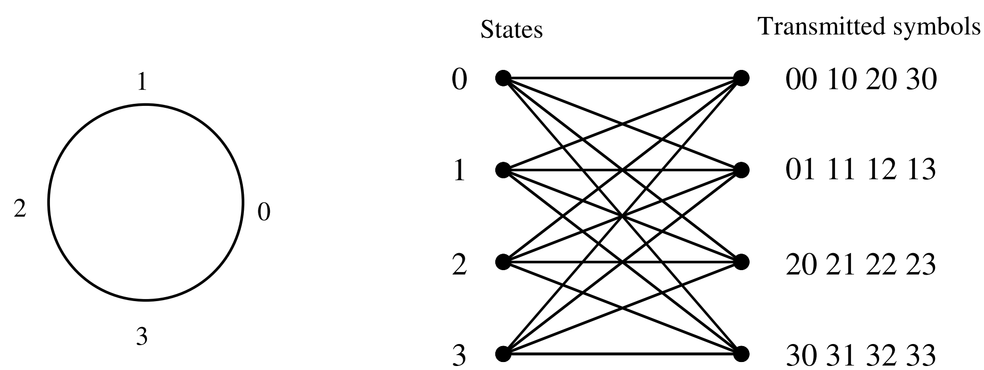


Figure 3.2 Delay diversity scheme

To explain the encoding processes of delay diversity, the simplest four-state quadrature Phase Shift keying (QPSK) symbol mapping, which has two transmit antennas is considered here. The signal constellation and a single stage of the trellis structure are shown in figure 3.3 for illustration.



a) QPSK constellation points b) Four-state QPSK space-time trellis

Figure 3.3 Trellis of four-state QPSK space-time encoder

The numbers on the left hand side of the trellis represent the states and the two symbols in each pair on the right hand side of the trellis are the encoded symbols to be transmitted through the two transmitting antennas.

At any time instant *t*, the four-state QPSK delay diversity encoder transmits symbols $x_1(t)$ and $x_2(t)$ over the two transmit antennas Tx_1 and Tx_2 respectively. The encoding operation for these codes can be explained with the help of the shift register shown in figure 3.4 where the adder performs a modulo-4 addition.

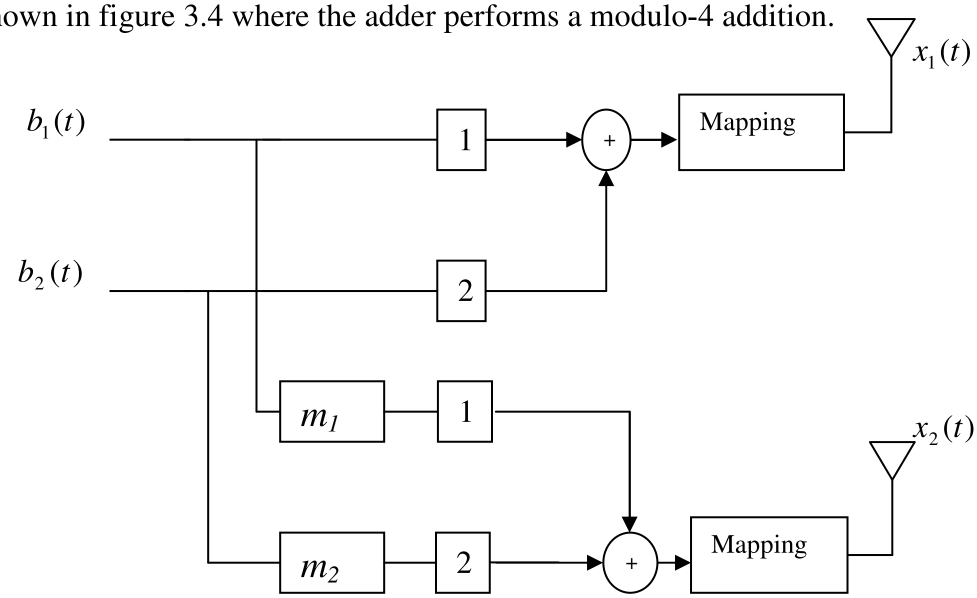


Figure 3.4 Four-state QPSK delay diversity encoder

In every timeslot, two bits $b_1(t)$ and $b_2(t)$ arrive at the encoder are multiplied by 1 and 2 respectively and added modulo-4 and the result will be mapped to a complex signal to be transmitted through the first antenna. At the same time these bites are saved in the registers m_1 and m_2 where the two bits $b_1(t-1)$ and $b_2(t-1)$ of the former timeslot were saved. The two bits arriving at the encoder at time *t* are converted into corresponding symbols which are elements of the set {0, 1, 2, . . ., M-1}, where $M = 4$ for QSPK symbol mapping. The outputs of the encoder at any instant *t* are the symbol pair $u_1(t)\ u_2(t)$ which can be obtained from $b_1(t)\ b_2(t)$ and $b_1(t-1)b_2(t-1)$ as follows.

$$u_1(t) = \left[[b_1(t)\ b_2(t)] \begin{bmatrix} 1 \\ 2 \end{bmatrix} \bmod 4 \right] \quad (3.7a)$$

$$u_2(t) = \left[[b_1(t-1)\ b_2(t-1)] \begin{bmatrix} 1 \\ 2 \end{bmatrix} \bmod 4 \right] \quad (3.7b)$$

The symbol pair $u_1(t)\ u_2(t)$ thus obtained will be one of the pairs on the right side of the trellis in figure 3.3.

By introducing a vector of binary elements $b(t) = [b_1(t)\ b_2(t)\ b_1(t-1)\ b_2(t-1)]$ and a generator matrix G, the outputs of the encoder can be obtained using a mapping function as

$$[u_1(t)\ u_2(t)] = b(t)G = [b_1(t)\ b_2(t)\ b_1(t-1)\ b_2(t-1)]G \quad \bmod\ 4 \quad (3.8)$$

The generator matrix G in this case is given by

$$G = \begin{bmatrix} 1 & 2 & 0 & 0 \\ 0 & 0 & 1 & 2 \end{bmatrix}^T$$, Where T represents transpose.

The output symbols will be mapped to QPSK signal constellation points as

$$x_1(t) = \exp\left(j\frac{2\pi}{4} u_1(t) \right) \quad (3.9a)$$

$$x_2(t) = \exp\left(j\frac{2\pi}{4} u_2(t) \right) \quad (3.9b)$$

To illustrate the way how this delay diversity encoder works, let's consider the encoding of the random input binary data 01111000. Starting in the zero state due to the input bits 01, the third branch in the trellis diagram of figure 3.3 is selected for leaving the initial state (00) and entering state 01. The output symbols $u_1(t)$ and $u_2(t)$ of the encoder are calculated as

$$[u_1(t)\ u_2(t)] = [0\ \ 1\ \ 0\ \ 0] \begin{bmatrix} 1 & 2 & 0 & 0 \\ 0 & 0 & 1 & 2 \end{bmatrix}^T \bmod\ 4 = [2\ \ 0] \quad (3.10)$$

The symbols 2 and 0 are mapped further to complex signals -1 and 1 and will be transmitted by the first and the second antenna respectively. The steps to encode the remaining binary sequence can be done in an analogous manner and it is summarized in table 3.1.

Table 3.1 Operation of the delay diversity encoder in Figure 3.4

Input sequence	Time instant	Input bits	Shift register	State	Output Symbols
00011110	0	--	00	0	--
000111	1	01	00	0	20
0001	2	11	01	2	32
00	3	10	11	3	13
	4	00	10	1	01
	5	--	00	0	--

The path along the trellis and the output symbols generated in each state transition is given in figure 3.5.

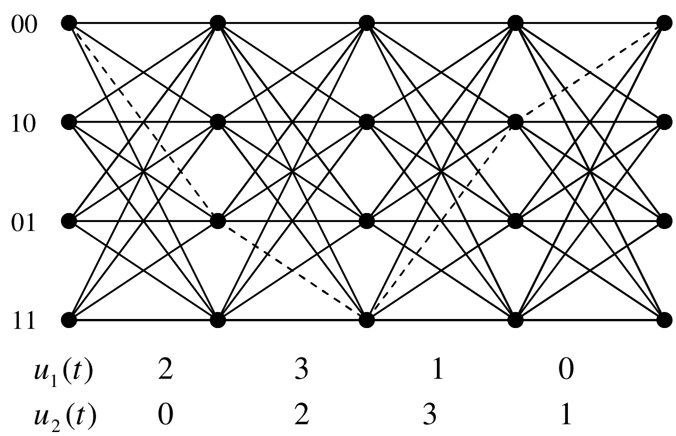


Figure 3.5 The path along the trellis and the corresponding outputs

Consequently the input data sequence is encoded into the output codeword matrix

$$U = \begin{bmatrix} 2 & 3 & 1 & 0 \\ 0 & 2 & 3 & 1 \end{bmatrix} \qquad (3.11a)$$

This codeword matrix U will be mapped further into a matrix X having entries of complex symbols.

$$X = \begin{bmatrix} -1 & -j & j & 1 \\ 1 & -1 & -j & j \end{bmatrix} \qquad (3.11b)$$

It can be observed that the above space-time encoder describes the delay diversity scheme for two transmit antennas.

The question of whether it is possible to generalize this delay diversity scheme to better channel codes than this repetition code has led to the so called space-time trellis coding.

3.4.2.1 Generalized Space-Time Trellis Coding [3, 12, 44]

For generalized space-time trellis codes having N transmit antennas, the encoder maps every input symbol to N coded symbols which are transmitted from all antennas and the mapping function is described by the trellis diagram. The encoder is composed of generator sequences that determine the symbols transmitted simultaneously.

To analyze the encoding process, consider the STTC encoder that uses an M-PSK symbol mapping with N transmit antennas. The input message stream denoted by $b = [b(0)\ \ b(1)\ \ \cdots\ \ b(t)\ \ \cdots\ \ ]$, where $b(t)$ is a group of $m = \log_2^M$ information bits at time *t*, that is $b(t) = [b_1(t)\ \ b_2(t)\ \ \cdots\ \ b_m(t)]$ is mapped into an M-PSK signal sequence x such that $x = [x(0)\ x(1) \cdots x(t) \cdots]$, where $x(t)$ is a space-time symbol at time *t* given by $x(t) = [x_1(t)\ \ x_2(t)\ \ \cdots\ \ x_N(t)]^T$.

The symbols $x_1(t), x_2(t), \cdots x_N(t)$ are transmitted simultaneously from N transmit antennas. As shown in figure 3.6, the m binary sequences $[b_1\ \ b_2\ \cdots\ b_m]$ are fed into the space-time trellis encoder, which consists of m feed forward shift registers. The kth input sequence $b_k = [b_k(0)\ \ b_k(1)\ \ \cdots\ \ b_k(t)\ \ \cdots\ \ ]$ for $k = 1, 2, \cdots m$ is passed to the kth shift register and multiplied by an encoder coefficient set. The multiplier outputs from all shift registers are added modulo-M and give the encoder output

$$x = [x_1 \quad x_2 \quad \cdots \quad x_N].$$

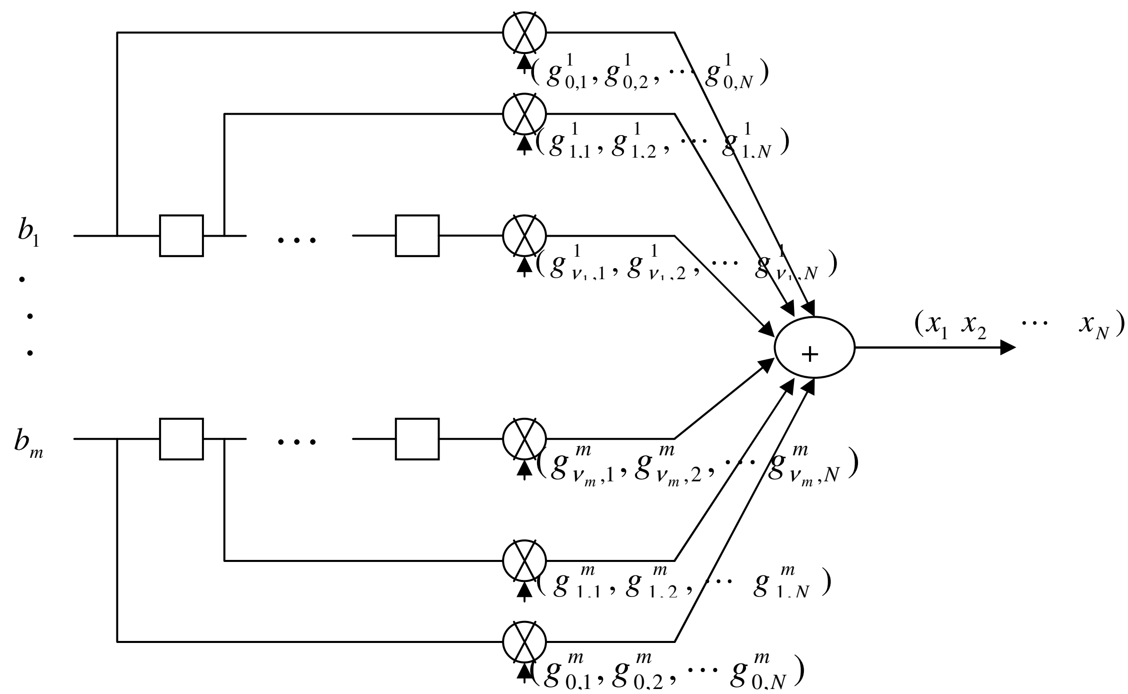


Figure 3.6 Generalized space-time trellis encoder

The m multiplier coefficient set sequences that describe the encoder are given by

$$g^1 = \left[(g^1_{0,1}, g^1_{0,2}, \cdots g^1_{0,N}), (g^1_{1,1}, g^1_{1,2}, \cdots g^1_{1,N}), \quad \cdots \quad (g^1_{\nu_1,1}, g^1_{\nu_1,2}, \cdots \; g^1_{\nu_1,N})\right]$$

$$g^2 = \left[(g^2_{0,1}, g^2_{0,2}, \cdots g^2_{0,N}), (g^2_{1,1}, g^2_{1,2}, \cdots g^2_{1,N}), \quad \cdots \quad (g^2_{\nu_2,1}, g^2_{\nu_2,2}, \cdots \; g^2_{\nu_2,N})\right]$$

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$$g^m = \left[(g^m_{0,1}, g^m_{0,2}, \cdots g^m_{0,N}), (g^m_{1,1}, g^m_{1,2}, \cdots g^m_{1,N}), \quad \cdots \quad (g^m_{\nu_m,1}, g^m_{\nu_m,2}, \cdots \; g^m_{\nu_m,N})\right]$$

Where, $g^k_{j,i}$, $k = 1,2, \cdots m$, $j = 1,2, \cdots \nu_k$, $i = 1,2, \quad \cdots \quad N$ is an element of the M-PSK constellation set and ν_k is the order of the memory of the k^th shift register.

The encoder output at time t for transmit antenna i denoted by $x_i(t)$ can be computed as

$$x_i(t) = \sum_{k=1}^{m} \sum_{j=0}^{v_k} g_{j,i}^{k} b_k(t-j) \mod M \ , \ i = 1,2, \ \cdots \ N \qquad (3.12)$$

These outputs are the elements of an M-PSK signal set. The space-time trellis coded M-PSK can achieve a bandwidth of m bits/s/Hz

3.4.3 Decoding of Space-Time Trellis Codes

The decoder for space-time trellis codes is based on channel estimation of the fading coefficients and maximum likelihood sequence estimation. To illustrate the decoding process of space-time trellis codes, the baseband representation of two transmit and one receive antenna is considered. At any transmission instant *t*, the symbols $x_1(t)$ and $x_2(t)$ are transmitted by the transmit antennas Tx_1 and Tx_2 respectively. The received signal is the superposition of the faded noisy signals from the two transmitting antennas and given by

$$y(t) = h_1 x_1(t) + h_2 x_2(t) + n(t) \qquad (3.13)$$

Where h_1 and h_2 are the fading coefficients of the channel and $n(t)$ is the noise.

Aided by the channel estimator, the maximum likelihood decoder based on the viterbi algorithm first finds the branch metric associated with every transition in the decoding trellis diagram. For every state transition, we have two estimated transmitted symbols $\hat{x}_1(t)$ and $\hat{x}_2(t)$ for which the branch metric is

$$BM = \left| y(t) - h_1 \hat{x}_1(t) - h_2 \hat{x}_2(t) \right|^2 = \left| y(t) - \sum_{j=1}^{2} h_j \hat{x}_j(t) \right|^2 \qquad (3.14)$$

The relation can be generalized to any number of transmit and receive antennas.

If the number of transmit and receive antennas is N and M respectively, the branch metric is given by [3]

$$BM = \sum_{i=1}^{M} \left| y_i(t) - \sum_{j=1}^{N} h_{ij} \hat{x}_j(t) \right|^2 \quad (3.15)$$

When all the transmitted symbols are received and the branch metric of each transition is calculated, the maximum likelihood sequence estimator invokes the Viterbi algorithm in order to find the maximum likelihood path associated with the lowest accumulated metric.

Chapter 4

STC DS-CDMA and Multiuser Receivers

4.1 Introduction

The realization of wireless communication systems that provide high data rate and high quality information has become a new challenge. CDMA is a recommended technology for the third generation wireless communication systems that support high speed data transmission [4, 24, 36]. CDMA has been designed to add features such as multimedia capabilities, high data rate and multi-rate services [35]. The data rates proposed are 144-384 kbps for high mobility applications and 2Mbps for indoor services [1, 42].

The space-time coding techniques that provide a significant increase of capacity and combat multi-path fading effects in single user channels have been adopted in cellular DS-CDMA communication systems with multiple transmit and receive antennas to support the demand on high data rate services. However, the use of space-time coding in DS-CDMA is limited by the multiple access interference which will be enhanced by the presence of multiple antennas at the transmitter [41]. To reduce the effect of multiple access interference, multiuser receivers can be deployed at the base station. Multiuser receivers are centralized receivers in multiple access systems in which the effect of multiple access interference is mitigated by exploiting the cross correlation properties of the spreading sequences of the multiple users [33]. By combining space-time coding and multiuser detection, the spectral efficiency of CDMA systems can be improved.

4.2 Single User versus Multiuser Detection

In addition to the noise term, interference is a major limiting factor in CDMA systems [13]. The interference from a single user might be small, but since there are multiple users in CDMA systems, the effect of interference usually exceeds the effect of noise unless it is mitigated through proper detection techniques.

CDMA systems support a multiplicity of users within the same bandwidth by assigning different typically unique codes to different users for their mutual communication in order to be able to distinguish their signals from one another. When the transmitted signals are subject to wireless propagation in CDMA environments, the signals of different users interfere with each other because of the deviation of the spreading sequences from perfect orthogonality [38]. Hence, CDMA systems are interference limited due to the interference generated by all active users transmitting information within the same bandwidth simultaneously.

The conventional single user CDMA detectors such as matched filter and rake receiver are optimized for detecting the signal from a single desired user and consider the multiple access interference that comes from other users as additional noise [34, 38]. Single user detectors are not optimal for CDMA environments as they are blind to the existence of other users in the system. A significantly better performance can be achieved by exploiting the additional knowledge of the users spreading codes [25]. Multiuser detection that jointly detects the signals from different users has been under intense investigation as a potential method to improve the performance of CDMA systems. In multiuser detection, instead of being separately estimated as in single user detection, the users are jointly detected for a better performance. Multiuser detection can be either optimal or suboptimal. The optimum multiuser receiver consists of a bank of matched filters for each user whose outputs produce a set of sufficient statistics followed by a Viterbi decoder that decodes the data sequences of all users based on the maximum likelihood decision algorithm. The optimum multiuser detector based on the maximum likelihood decision rule has a very large complexity that grows exponentially with the number of users. Hence, the optimum multiuser detector is prohibitively complex to implement in real applications. This computational complexity constraint led to numerous so called suboptimal multiuser detectors and interference cancellation schemes, which are new research areas these days.

4.3 Synchronous DS-CDMA in AWGN Channel

Consider K active users transmitting data to the base station over a period of time. The k^{th} user is assigned a signature waveform $s_k(t)$ which extends over the symbol period and takes the value zero outside the interval [0, T]. If the data symbol of user k is $b_k(t) = [b_k(0), b_k(1), \cdots, b_k(L-1)]$. For BPSK modulation, $b_k(i) \in \{-1, 1\}$, $i = 1, 2, 3, \cdots, L-1$. The transmitted baseband signal of this user at time t is given by [36, 40]

$$x_k(t) = \sum_{i=0}^{L-1} A_k b_k(i) s_k(t - iT) \tag{4.1}$$

Where, A_k is the amplitude of the transmitted signal. It is assumed that the signature waveforms are normalized that is $\int_0^T s_k^2(t)dt = 1$, k = 1, 2, . . . , K.

Each user sends a data block of length L. In the asynchronous case, there is a delay associated with each user. Assuming that the delay of user k is τ_k, the baseband signal at the front end of the receiver, which is the composite sum of the noisy signals transmitted by all the users can be represented as [40]

$$y(t) = \sum_{k=1}^{K} \sum_{i=0}^{L-1} A_k b_k(i) s_k(t - iT - \tau_k) + n(t) \tag{4.2}$$

Where, $n(t)$ is the additive white gaussian noise with a variance of $N_0/2$.

If the delays are equal and set to zero, that is $\tau_1 = \tau_2 = \tau_3 = \cdots = \tau_K = 0$, we get a model for synchronous CDMA system.

Through out this thesis work, synchronous DS-CDMA is assumed. In the synchronous model of CDMA, the symbol boundaries of each user are aligned, hence it is sufficient to consider one symbol interval and the received signal over one period becomes

$$y(t) = \sum_{k=1}^{K} A_k b_k(0) s_k(t) + n(t) \qquad 0 \le t \le T \qquad (4.3)$$

The receiver consists of a bank of filters matched to the signature waveforms assigned to the users and a multiuser detector. The output of the filter matched to the signature waveform of user k and sampled at T is given by

$$y_k = \int_0^T y(t) s_k(t) dt \qquad = A_k b_k + \underbrace{\sum_{i=1, i \neq k}^{K} A_k \rho_{ik} b_i}_{MAI} + n_k \qquad (4.4)$$

$n_k = \int_0^T n(t) s_k(t) dt$ is the noise at the output of the k^{th} user matched filter and

$\rho_{ik} = \int_0^T s_i(t) s_k(t) dt$ represents the cross correlation between the waveforms of users i and user k.

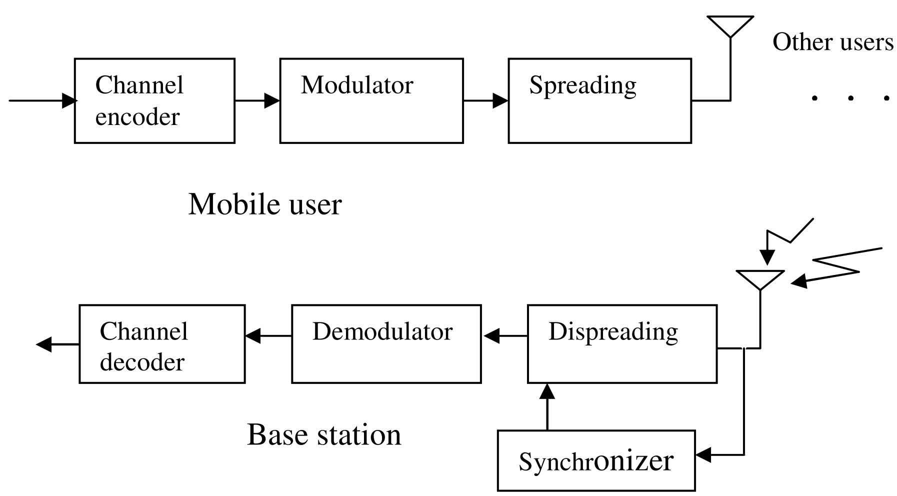

Figure 4.1 Block diagram of a CDMA system

The outputs of the bank of matched filters constitute all the sufficient statistics for the decoding problem and can be expressed in matrix form as [29]

$$y = \begin{bmatrix} y_1 & y_2 & \cdots & y_K \end{bmatrix}^T = RAb + n \tag{4.5}$$

Where R is a $K \times K$ correlation matrix whose entries are the values of the correlation between every pair of the signature waveforms and A is a diagonal matrix of signal amplitudes.

Since $\rho_{ik} = \rho_{ki}$, the matrix R is symmetric and can be broken into two matrices one representing the autocorrelation and the other the cross correlation which contains all the off-diagonal elements.

Denoting the off-diagonal matrix by Q, R can be expressed as

$$R = (I)_{KxK} + Q \tag{4.6}$$

The output of the matched filters can now be written as

$$y = (I + Q)Ab + n = Ab + QAb + n \tag{4.7}$$

The first term on the right hand side of equation (4.7) given by Ab is the decoupled data weighted by the signal amplitudes, the second term given by QAb is the multiple access interference and the third term denoted by n is the additive gausian noise [25]. The detector which gives the optimal performance in the synchronous multiuser environment chooses the most likely transmitted sequence b that maximizes the joint a-posteriori probability that b was sent given y(t) was received. Thus the optimal detector selects the most likely sequence of bits such that

$$\max_b \{P(b / y(t) , \forall_t)\}$$

Under the assumption that all the possible transmitted sequences are equally probable, this decoding scheme is called the maximum likelihood sequence estimator (MLSE).

This detector promises huge capacity and performance gains; however, it is computationally complex as it performs an exhaustive search over all 2^K possible data bit sequences.

4.4 Space-Time Coding in DS-CDMA

Most previous works in space-time coding techniques have been concerned with single user channels. The use of space-time coding in DS-CDMA is an effective way to increase the spectral efficiency of multiple access systems as well as to combat fading effects. However, the benefits of space-time coding techniques in DS-CDMA systems are offset by the multiple access interference which is enhanced further by the use of multiple transmit antennas that are used in space-time coding. In DS-CDMA, the effect of multiple access interference is mitigated by multiuser detection that exploits the structural property of the multiple access interference and attempts to reduce its effect. Space-time trellis coding combined with multiuser detection promises good capacity gain in DS-CDMA systems. The optimum decoding operation for space-time trellis coded DS-CDMA systems that can be implemented with a viterbi type algorithm is very complex. This computational complexity is a motivation for researchers to devise reduced complexity suboptimal receivers that can be used in practical systems.

4.4.1 Model of STTC DS-CDMA Systems

To develop a framework for understanding the communication system of interest, a model of the system is created. Discussion is restricted, without loss of generality, to the baseband model to avoid carrier frequency notation.

In cellular communication systems, the antenna diversity burden is recommended to be at the base station. This is because antenna diversity in the mobile receiver increases the cost and complexity of the terminal apparatus. For example, in GSM cellular systems, multiple receive antennas at the base station are used [3]. At present, network operators have implemented antenna diversity in the mobile handset and this is likely to continue such that in the upcoming third generation and fourth generation mobile networks, it will become more common [9].

Consider the uplink of a DS-CDMA system consisting of K active users, N transmit antennas each and one receive antenna at the base station. The binary information sequence $b_k(i)$ of user k for k=1, 2, . . ., K, is space-time coded by a space-time encoder to produce a codeword $D_k(i) = [d_{k,1}(i)\ d_{k,2}(i) \cdots d_{k,N}(i)]^T$. The encoded data is distributed over N transmit antennas by a serial to parallel converter. The encoded symbols in each parallel stream are spread by a unique signature waveform $s_k(t)$ and are transmitted simultaneously from the N transmit antennas. It is assumed that the same spreading waveform is used for the N antennas and the antenna elements experience independent fading. The baseband transmitted signal by the k^{th} user at time *t* can be written as [1, 10]

$$x_k(t) = \sum_{i=0}^{L-1} \sum_{j=1}^{N} \left(A_k / \sqrt{N}\right) d_{kj}(i)\ s_k(t - iT) \tag{4.8}$$

For QPSK symbol mapping, $d_{kj}(i) \in \{j, -j, \quad 1, \quad -1\}_{i=o}^{L-1}$ is the space-time encoder output of the k^{th} user on transmit antenna *j* at time *i* and L is the length of channel symbols per user in a data frame. The signature waveforms are normalized and are confined in the symbol period and zero outside, that is

$$\int s_k^2(t)dt = \begin{cases} 1, & for \quad 0 \le t \le T \\ 0, & otherwise \end{cases} \qquad \text{for} \quad k = 1, 2, 3, \cdots, K \tag{4.9}$$

Assuming that the fading is slow such that the channel characteristics are uniform over a transmitted data frame, the corresponding received baseband signal at the front end of the receiver is given by [1]

$$y(t) = \sum_{i=0}^{L-1} \sum_{k=1}^{K} \sum_{j=1}^{N} \left(A_k / \sqrt{N}\right) h_{kj} d_{kj}(i)\ s_k(t - iT) \quad + n(t) \tag{4.10}$$

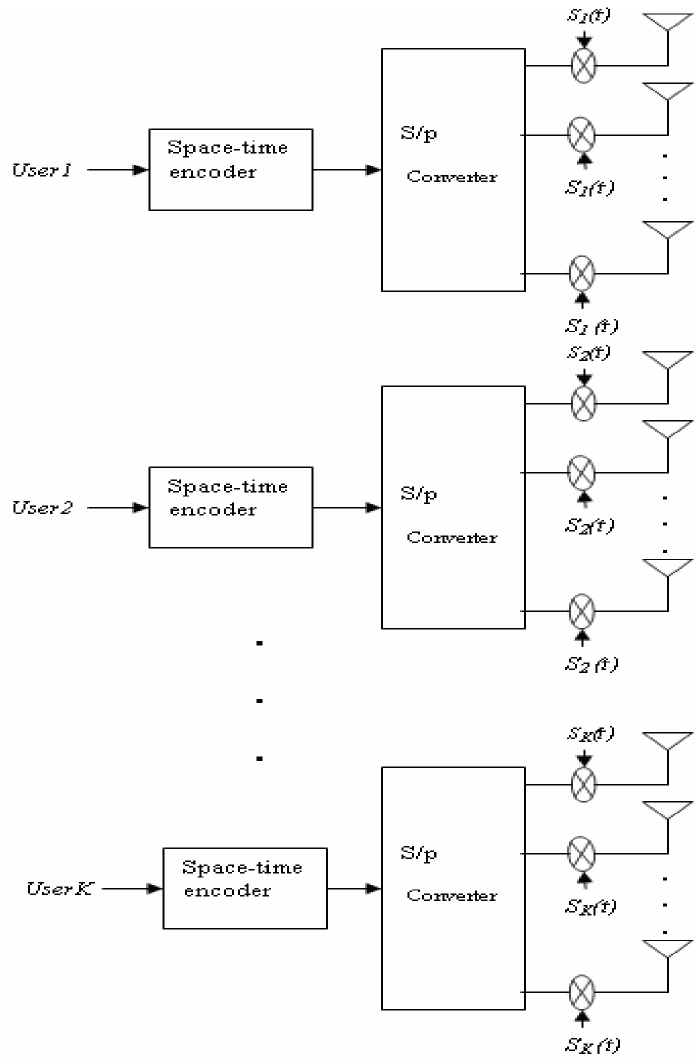

Figure 4. 2 Baseband Transmitter model of STTC DS-CDMA uplink

The complex fading channel coefficients h_{kj} are assumed to be zero mean gaussian random variables with independent real and imaginary parts with a variance of 0.5 per dimension. Equivalently, the channel gains have uniform phase and Rayliegh amplitude probability distribution.

4.4.2 Joint ML multiuser decoding of STTC DS-CDMA

Denoting the transmitted symbol of user k on antenna j at time i by $d_{k,j}(i)$, the vector $D_k(i) = [d_{k,1}(i)\ d_{k,2}(i) \cdots d_{k,N}(i)]^T$ represents the k^{th} user transmitted symbols on the N transmit antennas at time i. With $D_k = [D_k(0)\quad D_k(1)\quad \cdots\quad D_k(L-1)]$ representing the data frame of user k for $k = 1, 2, 3, \cdots, K$, define a joint codeword D of dimension $NK \times L$ of all uses as

$$D = \begin{bmatrix} D_1 \\ D_2 \\ D_3 \\ \vdots \\ D_k \end{bmatrix} = \begin{bmatrix} D_1(0) & D_1(1) & D_1(2) \cdots\cdots\cdots\cdots D_1(L-1) \\ D_2(0) & D_2(1) & D_2(2) \cdots\cdots\cdots\cdots D_2(L-1) \\ D_3(0) & D_3(1) & D_3(2) \cdots\cdots\cdots\cdots D_3(L-1) \\ \vdots \quad \vdots & \vdots & \vdots \qquad \ddots \qquad \vdots \\ D_K(0) & D_K(1) & D_K(2) \cdots\cdots\cdots D_K(L-1) \end{bmatrix} \tag{4.11}$$

The joint codeword D of all the users is called the super codeword. The space-time encoded output from all the users and all antennas at time i is a $KN \times 1$ column vector denoted by

$$D(i) = \begin{bmatrix} D_1(i) \\ D_2(i) \\ \vdots \\ D_K(i) \end{bmatrix} \tag{4.12}$$

The channel fading coefficients of the k^{th} user can be collected into a vector as $H_k = [h_{k,1}\quad h_{k,2}\quad \cdots\quad h_{k,N}]$ and we can combine these vectors of channel coefficients of all users into a matrix of dimension $K \times KN$ as

$$H = \begin{bmatrix} H_1 & 0 & 0 & \cdots & 0 \\ 0 & H_2 & 0 & \cdots & 0 \\ 0 & 0 & H_3 & \cdots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & 0 & 0 & H_K \end{bmatrix} \tag{4.13}$$

With this notation, the output $y(i) = [y_1(i)\quad y_2(i)\quad \cdots\quad y_K(i)]^T$ of a bank of K matched filters at time i is given by

$$y(i) = RAHD(i) + \eta(i) \qquad (4.14)$$

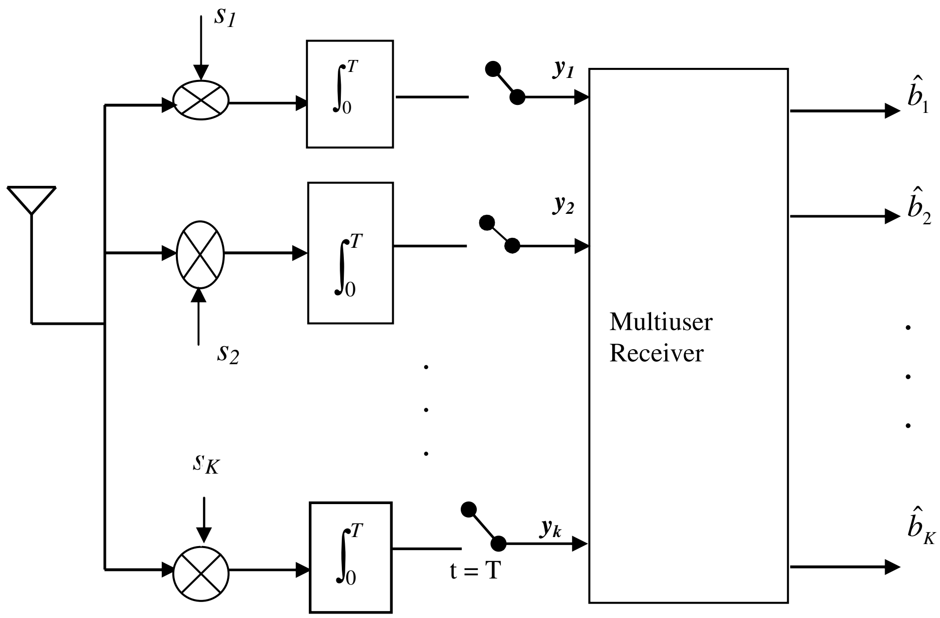

Figure 4.3 Multiuser receiver

Where A is a diagonal matrix of signal amplitudes given by

$$A = diag \begin{bmatrix} \frac{A_1}{\sqrt{N}} & \frac{A_2}{\sqrt{N}} & \cdots & \frac{A_K}{\sqrt{N}} \end{bmatrix} \qquad (4.15)$$

R is the correlation matrix of the users' signature waveforms, $\eta(i)$ is the zero mean colored gaussian noise vector with $E[\eta(i)(\eta(i))^H] = N_O R$ and $(.)^H$ is a Hermitian operator.

Denoting the vector of length L of the k^{th} matched filter outputs corresponding to the complete received codeword as $y_k = [y_k(0) \;\; y_k(1) \;\cdots\; y_K(L-1)]$ and the matrix of dimension $K \times L$ corresponding to the outputs of all the matched filters by

$$y = \begin{bmatrix} y_1 \\ y_2 \\ y_3 \\ \vdots \\ y_K \end{bmatrix} = \begin{bmatrix} y_1(0) & y_1(1) & y_1(2)\cdots\cdots\cdots y_1(L-1) \\ y_2(0) & y_2(1) & y_2(2)\cdots\cdots\cdots y_2(L-1) \\ y_3(0) & y_3(1) & y_3(2)\cdots\cdots\cdots y_3(L-1) \\ \vdots & \vdots & \vdots \quad \vdots \quad \ddots \quad \vdots \\ y_K(0) & y_K(1) & y_K(2)\cdots\cdots\cdots y_K(L-1) \end{bmatrix} \qquad (4.16)$$

Then we can write

$$y = RAHD + \eta \qquad (4.17)$$

η is the zero mean colored gaussian noise matrix of dimension $K \times L$ with $E[\eta\eta^H] = N_O R$

The joint maximum likelihood multiuser decision rule for the space-time coded DS-CDMA system is given by [1, 38]

$$\hat{D} = \arg\{ \underbrace{\max}_{D} P(y / D)\} . \qquad (4.18a)$$

Or equivalently,

$$\hat{D} = \arg\left[\underbrace{\max}_{D} \left(2\,\mathrm{Re}\left[D^H H^H A^H y\right] - D^H H^H A^H RHD\right)\right] \qquad (4.18b)$$

$(.)^H$ is the complex conjugate transpose of the matrix (.).

The maximum likelihood decoder must estimate the transmitted symbols from all users and all antennas. For a DS-CDMA system, if *K* is the number of active users each employing N transmit antennas and M-PSK space-time coding, the number of possible combination of codewords transmitted from *KN* antennas is M^{KN}.

The maximization problem in the joint maximum likelihood multiuser decision rule above is carried out over all the valid super codewords. This joint maximum likelihood detector and decoder performs a very exhaustive search over a super trellis made up by combining all the users' space-time code trellises. The super trellis for two users each employing four-state QPSK space-time coding is illustrated in figure 4.4 when the combined state changes from $s_{1,i}\, s_{2,i}$ to $s_{1,i+1}\, s_{2,i+1}$, where $s_{1,i}$ and $s_{2,i}$ are the states of the first and the second user respectively at an instant *i*. This huge complexity burden makes the joint maximum likelihood multiuser detector impractical for real systems.

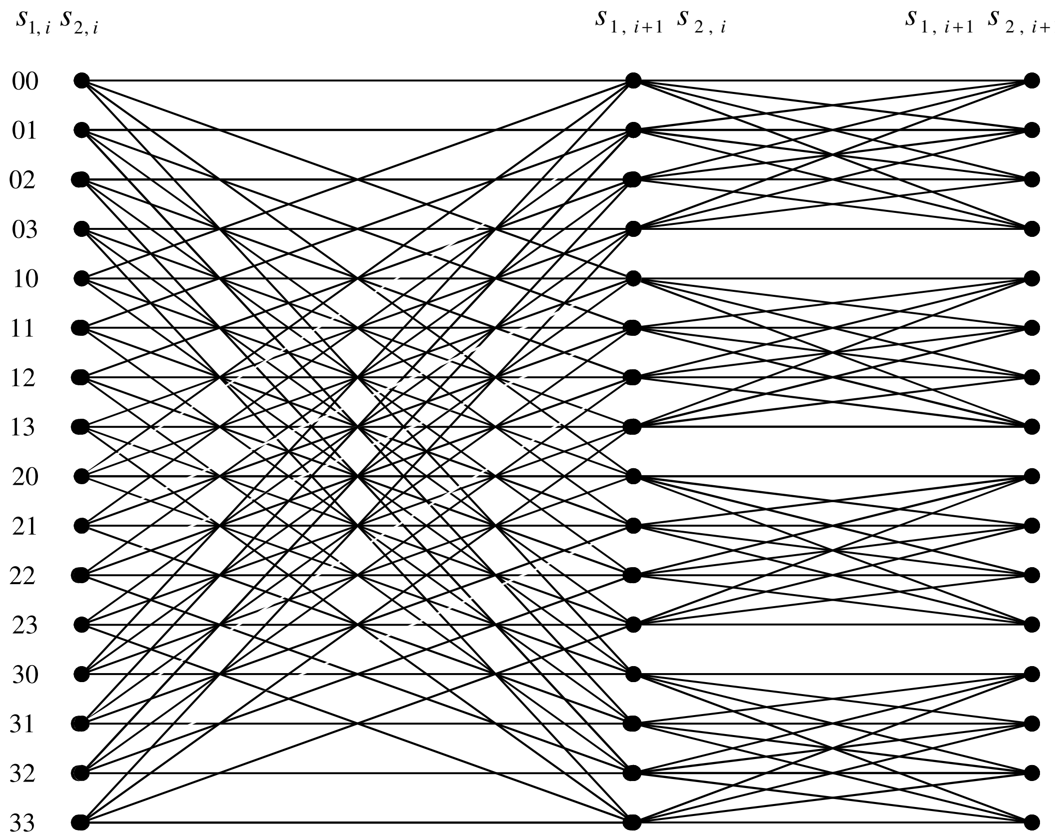

Figure 4.4 Super trellis made by combining the trellises of two users

4.5 Low Complexity Multiuser Receiver for STTC DS-CDMA

This low complexity detector first performs a linear transformation on the outputs of the bank of matched filters. The linear operation can be done by a simple whitening filter. Since a different signature waveform is assigned to each user, the correlation matrix R is symmetric and positive definite. This guarantees the fact that R can be decomposed into a product of upper and lower triangular matrices. The channel whitening process can be carried out by performing the Cholesky factorization of R as in [43]

$$R = F^T F \tag{4.19}$$

Where, F is a lower triangular matrix and it is called the cholesky factor of R.

If $y(i) = RAHD(i) + \eta(i)$, which is the output of the matched filters at time i is fed to $(F^T)^{-1} = FR^{-1}$, the output is

$$Z(i) = (F^T)^{-1} y(i) = FAHD(i) + n(i) \tag{4.20}$$

$$E[n(i)] = 0_{Kx1} \quad \text{and} \quad E[n(i)n(i)^H] = N_O I_{Kxk}$$

The matrix F above is a lower triangular matrix of the following form

$$F = \begin{bmatrix} f_{1,1} & 0 & 0 & \cdots & 0 \\ f_{2,1} & f_{2,2} & 0 & \cdots & 0 \\ f_{3,1} & f_{3,2} & f_{3,3} & \cdots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ f_{K,1} & f_{K,2} & f_{k,3} & \cdots & f_{K,K} \end{bmatrix} \tag{4.21}$$

If $X(i) = AHD(i)$, then the output of the whitening filter at time i can be written as

$$Z(i) = FX(i) + n(i) \tag{4.22a}$$

or more explicitly

$$\begin{bmatrix} z_1(i) \\ z_2(i) \\ z_3(i) \\ \vdots \\ z_K(i) \end{bmatrix} = \begin{bmatrix} f_{1,1} & 0 & 0 & \cdots & 0 \\ f_{2,1} & f_{2,2} & 0 & \cdots & 0 \\ f_{3,1} & f_{3,2} & f_{3,3} & \cdots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ f_{K,1} & f_{K,2} & f_{k,3} & \cdots & f_{K,K} \end{bmatrix} \begin{bmatrix} x_1(i) \\ x_2(i) \\ x_3(i) \\ \vdots \\ x_K(i) \end{bmatrix} + \begin{bmatrix} n_1(i) \\ n_2(i) \\ n_3(i) \\ \vdots \\ n_K(i) \end{bmatrix} \tag{4.22b}$$

The channel whitening filter is important for two reasons. First, unlike the decorrelator detector, it reduces the noise variance and second it acts on the input data $y(i)$ such that the output $Z(i)$ of this whitening filter is partially decorrelated in the sense that the output $z_1(i)$ for user 1 (k =1) is interference free, for user 2 (k=2) interference comes only from user 1, for user 3 (k=3) interference comes from user 1 and user 2 and so on. In general, for the k^{th} user, interference comes from users with user numbers 1,2,3 . . .k-1 and this is due to the lower triangular nature of the matrix F.

Equation (4.22b) in matrix form is a set of K linear equations for which, in the case of optimal decoding, the decoder, in each state transition and at each state, attempts to estimate not an exact solution but a best solution from a given domain that contains all the constellation points. The solution is best in the sense that $\left|Z(i) - F\hat{X}(i)\right|^2$ is minimized, where $\hat{X}(i)$ is the estimated transmitted sequence at time i weighted by the corresponding fading coefficients and signal amplitudes.

For any user $k, 1 \le k \le K$, the output of the whitening filter can be written as

$$z_k(i) = f_{k,1}x_1(i) + f_{k,2}x_2(i) + \cdots + f_{k,k-1}x_{k-1}(i) + f_{k,k}x_k(i) + n_k(i) \quad (4.23a)$$

$$z_k(i) = f_{k,k}x_k(i) + \underbrace{\sum_{j=1}^{k-1} f_{k,j}x_j(i)}_{MAI} + n_k(i) \quad (4.23b)$$

The whitening filter is followed by a set of decoders. In the simplest scheme, there are K decoders each one decoding a single user by way of the viterbi algorithm.

In each state transition, the k^{th} decoder for k=1,2, . . .K, selects $\hat{x}_k(i)$ that minimizes the branch metric given by

$$BM = \left|z_k(i) - f_{k,k}\hat{x}_k(i) - \hat{I}_k(i)\right|^2 \quad (4.24)$$

Where, $\hat{I}_k(i)$ is the estimated multiuser interference given by

$$\hat{I}_k(i) = \sum_{j=1}^{k-1} f_{k,j}\hat{x}_j(i) \quad (4.25)$$

and

$$\hat{x}_k(i) = \left(\hat{A}_k / \sqrt{N}\right)\hat{H}_k\hat{D}_k(i) \quad (4.26)$$

Estimating $\hat{x}_k(i)$ involves the estimation of signal amplitudes, channel coefficients and transmitted sequences. In this work, the signal amplitudes and channel coefficients are provided to the receiver as side information (perfect signal amplitude and channel estimation is assumed) and only the transmitted sequences are unknown to the receiver. With this assumption, estimating $\hat{x}_k(i)$ is equivalent to estimating the transmitted sequences of user k.

In this simplest scheme, first, decoder 1 uses $z_1(i)$ to compute the branch metric and estimates $\hat{x}_1(i)$ independently as $z_1(i)$ is interference free. In the second stage of the decoding operation, the estimated value $\hat{x}_1(i)$ is used to regenerate and subtract the interference power that the first user has on the second and decoder 2 computes the branch metric and estimates $\hat{x}_2(i)$. In general, the k^{th} stage metric computation uses the estimated values $\hat{x}_1(i),\ \hat{x}_2(i),\ \cdots\ \hat{x}_{k-1}(i)$ to regenerate the interference power. Thus, the metric computation of the k^{th} decoder makes use of the estimated values of the k-1 users. One drawback of this scheme is the error propagation. For any user k, $1 \le k \le K-1$, if there is an error in estimating $\hat{x}_k(i)$, this error negatively affects not only the next user but also all users $k+1,\ k+2, \cdots K$. To reduce this negative effect of error propagation, the users can be decoded in groups. So in the generalized scheme of the proposed receiver, the output signals of the whitening filter are divided into groups and the interference from each group is estimated and subtracted from the groups next to it. First, estimates of the transmitted sequence for the users in the first group are obtained by the maximum likelihood detection rule. The multiple access interference of this group of users is then regenerated and subtracted from the next group of users. After the interference from the users in the first group is subtracted, estimates of the transmitted sequence for the second group of users will be obtained once again and this will continue until the last group of users is reached.

Assuming that there are G group decoders and M_g is the number of users decoded by the g^{th} group decoder for $g = 1,2,3, \cdots G$, then all the G group decoders decode the K users that is

$$\sum_{g=1}^{G} M_g = K \tag{4.27}$$

With out any loss of generality, lets assume that each group decoder decodes M users so that

$M_1 = M_2 = M_3 = \cdots = M_G = M$. Then, $\sum_{g=1}^{G} M_g = \sum_{g=1}^{G} M = GM = K$

For mathematical analysis of the proposed receiver in its generalized form, the following new notations are introduced

$$F_g = \begin{bmatrix} f_{(g-1)M+1,1} & f_{(g-1)M+1,2} & \cdots & f_{(g-1)M+1,(g-1)M+1} & 0 & 0 & \cdots & 0 \\ f_{(g-1)M+2,1} & f_{(g-1)M+2,2} & \cdots & f_{(g-1)M+2,(g-1)M+1} & f_{(g-1)M+2,(g-1)M+2} & 0 & \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & & \ddots & \vdots \\ f_{gM,1} & f_{gM,2} & \cdots & & & & & f_{gM,gM} \end{bmatrix}$$

For $j = 1, 2, \cdots, g$

$$F_{gj} = \begin{bmatrix} f_{(g-1)M+1,(j-1)M+1} & f_{(g-1)M+1,(j-1)M+2} & \cdots & f_{(g-1)M+1,\,jM} \\ f_{(g-1)M+2,(j-1)M+1} & f_{(g-1)M+2,(j-1)M+2} & \cdots & f_{(g-1)M+2,\,jM} \\ \vdots & \vdots & \vdots & \vdots \\ f_{gM,(j-1)M+1} & f_{gM,(j-1)M+1} & \cdots & f_{gM,\,jM} \end{bmatrix}$$

$$Z_g(i) = \begin{bmatrix} z_{g1}(i) \\ z_{g2}(i) \\ \vdots \\ z_{gM}(i) \end{bmatrix},\; X_g(i) = \begin{bmatrix} x_{g1}(i) \\ x_{g2}(i) \\ \vdots \\ x_{gM}(i) \end{bmatrix} X^g(i) = \begin{bmatrix} X_1(i) \\ X_2(i) \\ \vdots \\ X_g(i) \end{bmatrix}, N_g(i) = \begin{bmatrix} n_{g1}(i) \\ n_{g2}(i) \\ \vdots \\ n_{gM}(i) \end{bmatrix}, N^g(i) = \begin{bmatrix} N_1(i) \\ N_2(i) \\ \vdots \\ N_g(i) \end{bmatrix}$$

Where, for $g = 1, 2, 3, \cdots G$ and $m = 1, 2, 3, \cdots M$.

F_g is the g^th block of the matrix F, F_{gj}'s are sub matrices of the matrix F_g, $Z_g(i)$ is the input corresponding to the g^th group decoder, $X_g(i)$ is the possible weighted transmitted sequences corresponding to the g[th] group decoder and $N_g(i)$ is the corresponding noise of the g[th] group decoder. For example, if there are six users and three group decoders each one decoding two users, then the three decoders decode all of the six users. In this case the above matrices will be

$$F = \begin{bmatrix} f_{1,1} & 0 & 0 & 0 & 0 & 0 \\ f_{2,1} & f_{2,2} & 0 & 0 & 0 & 0 \\ f_{3,1} & f_{3,2} & f_{3,3} & 0 & 0 & 0 \\ f_{4,1} & f_{4,2} & f_{4,3} & f_{4,4} & 0 & 0 \\ f_{5,1} & f_{5,2} & f_{5,3} & f_{5,4} & f_{5,5} & 0 \\ f_{6.1} & f_{6,2} & f_{6,3} & f_{6,4} & f_{6,5} & f_{6,6} \end{bmatrix}$$

$$F_1 = \begin{bmatrix} f_{1,1} & 0 \\ f_{2,1} & f_{2,2} \end{bmatrix}, \quad F_2 = \begin{bmatrix} f_{3,1} & f_{3,2} & f_{3,3} & 0 \\ f_{4,1} & f_{4,2} & f_{4,3} & f_{4,4} \end{bmatrix}, \quad F_3 = \begin{bmatrix} f_{5,1} & f_{5,2} & f_{5,3} & f_{5,4} & f_{5,5} & 0 \\ f_{6,1} & f_{6,2} & f_{6,3} & f_{6,4} & f_{6,5} & f_{6,6} \end{bmatrix}$$

$$F_{1,1} = \begin{bmatrix} f_{1,1} & 0 \\ f_{2,1} & f_{2,2} \end{bmatrix}, \quad F_{2,1} = \begin{bmatrix} f_{3,1} & f_{3,2} \\ f_{4,1} & f_{4,2} \end{bmatrix}, \quad F_{2,2} = \begin{bmatrix} f_{3,3} & 0 \\ f_{4,3} & f_{4,4} \end{bmatrix}$$

$$F_{3,1} = \begin{bmatrix} f_{5,1} & f_{5,2} \\ f_{6,1} & f_{6,2} \end{bmatrix}, \quad F_{3,2} = \begin{bmatrix} f_{5,3} & f_{5,4} \\ f_{6,3} & f_{6,4} \end{bmatrix}, \quad F_{3,3} = \begin{bmatrix} f_{5,5} & 0 \\ f_{6,5} & f_{6,6} \end{bmatrix}$$

In $z_{gm}(i)$, index g is group decoder number and index m is user number in group decoder g. With these notations, the matrix equation in equation (4.22b) can be decomposed into G separate matrix equations corresponding to the G parallel group decoders as

$$Z_g(i) = F_g X^g(i) + N^g(i), \quad g = 1,2,3, \cdots G \tag{4.28}$$

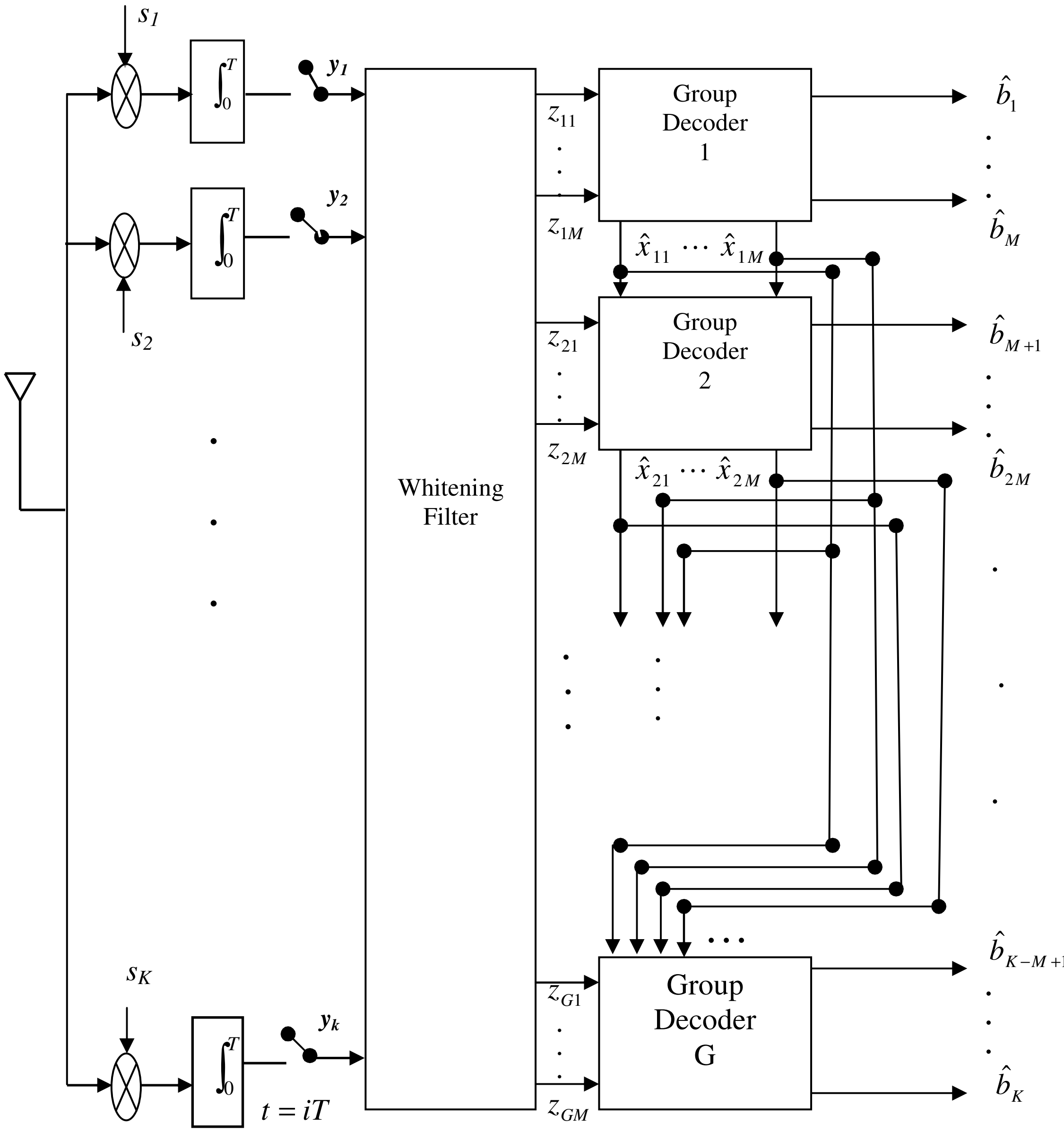

Figure 4.5 Low complexity multiuser receiver

Two extreme cases:

If M=1, then G = K, in this case, the proposed receiver is reduced to its simplest form and there are K decoders each decoding a single user and there are K linear equation corresponding to each decoder.

If G =1, then M = K, in this case, there is only one decoder that decodes all the K users optimally by the viterbi algorithm.

In the generalized scheme of the proposed receiver, the metric computation of the group decoders is such that each group decoder selects $\hat{X}_g(i)$ that minimizes the branch metric given by

$$BM = \left\| Z_g(i) - F_{g,g}\hat{X}_g(i) - \hat{I}_g(i) \right\|^2 \tag{4.29}$$

Where, $\hat{I}_g(i) = \sum_{j=1}^{g-1} F_{g,j}\hat{X}_j(i)$ is the estimated multiuser interference.

At the first stage of the decoding operation, the first group decoder uses $Z_1(i)$ to compute the branch metric and estimates $\hat{X}_1(i)$, at the second stage of the decoding operation, the interference power that comes from the users in the first group is regenerated and subtracted from $Z_2(i)$ and the second group decoder computes the branch metric and estimates $\hat{X}_2(i)$ and so on. When all the transmitted sequences are received, each group decoder invokes the viterbi algorithm and selects the sequence with the lowest accumulated metric.

When each group decoder computes the branch metric, the interference power that comes from other group decoders is subtracted to cancel out its effect. So the group decoders are optimal decoders for the M users in the group with interference cancellation schemes.

4.6 Complexity Analysis

The complexity comparison is based on the requirements of memory units that are used to store the surviving path along with its metric and the total number of operations per symbol. Assuming that each user encoder has S states, the total state of the optimum decoder with super trellis is given by

$$S_{opt} = S^K \tag{4.30}$$

The number of total states required by the proposed receiver with M_g users per group decoder is

$$S_{prop} = \sum_{g=1}^{G} S^{M_g} \tag{4.31}$$

The memory requirement for the receiver depends on the number of states and number of users. The number of memory units for the optimum decoder with super trellis and the proposed receiver are given respectively as

$$M_{opt} = S_{opt}(K+1) \tag{4.32}$$

$$M_{prop} = \sum_{g=1}^{G} S^{M_g}(M_g + 1) \tag{4.33}$$

Assuming that the comparison operation is as complex as addition, the number of operations refers to the total number of comparison, addition and multiplication operations. In the viterbi decoder with N transmit antennas, if B is the number of branches leaving or entering a state, then there are $B-1$ comparison operations, BN addition operations and $B(N+1)$ multiplication operations for calculating and selecting the branch metric per state. The total number of operations per state is the sum of comparison, addition and multiplication operations and it is given by

$$\Gamma = 2B(N+1) - 1 \tag{4.34}$$

Thus, for each decoded symbol of all users, the number of operations required by the optimal decoder and the proposed receiver are given respectively by

$$O_{opt} = M_{opt}\Gamma \tag{4.35}$$

$$O_{prop} = M_{prop}\Gamma \tag{4.36}$$

The memory requirements and number of operations of the optimal decoder and the proposed low complexity receiver per symbol of each user are summarized in table 4.1

Table 4.1 Summery of computational complexities

Decoder type	Number of total states	Memory size	Comparison operations per state	Addition operations Per state	Multiplication operations per state
MLSE	S^K	$S^K(K+1)$	$B-1$	BN	$B(N+1)$
Proposed receiver	$\sum_{g=1}^{G} S^{M_g}$	$\sum_{g=1}^{G} S^{M_g}(M_g+1)$	$B-1$	BN	$B(N+1)$

Chapter 5

Simulation and Results

5.1 Simulation Modeling

In the simulation, the uplink of a space-time coded DS-CDMA is considered. It is assumed that each user employs two transmit antennas and there is one receive antenna at the base station. The user number K is varied. Bipolar Gold sequences of length $L=31$ are generated for spreading user data. For DS-CDMA, the length of PN sequences is the same as the gain, thus the gain in dB is $10\times\log_{10}^{L}=14.9$. Random binary numbers corresponding to each user will be generated using random number generator. These binary numbers will be grouped into frames called symbols. The space-time encoder of each user takes the frame as input and generates codeword pairs for each symbol input simultaneously for all the transmit antennas. Since there is interest in high rate data transmission, four-state QPSK symbol mapping is preferred over BPSK. The outputs of the space-time encoder are complex signals which will be spread by signature sequences that are unique for each user. After spreading, the signals will be transmitted through a fading channel.

It is assumed that the channel is quasi-static that fades according to the Rayleigh probability distribution. The channel gains are modeled as complex Gaussian random variables with zero mean and variance of 0.5 per dimension. At the front end of the receiver, noise is generated and added to the composite signal. The signals are modeled in baseband, thus modulation and demodulation operations are not carried out. In each simulation, binary data of length 260 are generated for each user and at a given signal to noise ratio, the simulation is repeated 50 times and the results are averaged

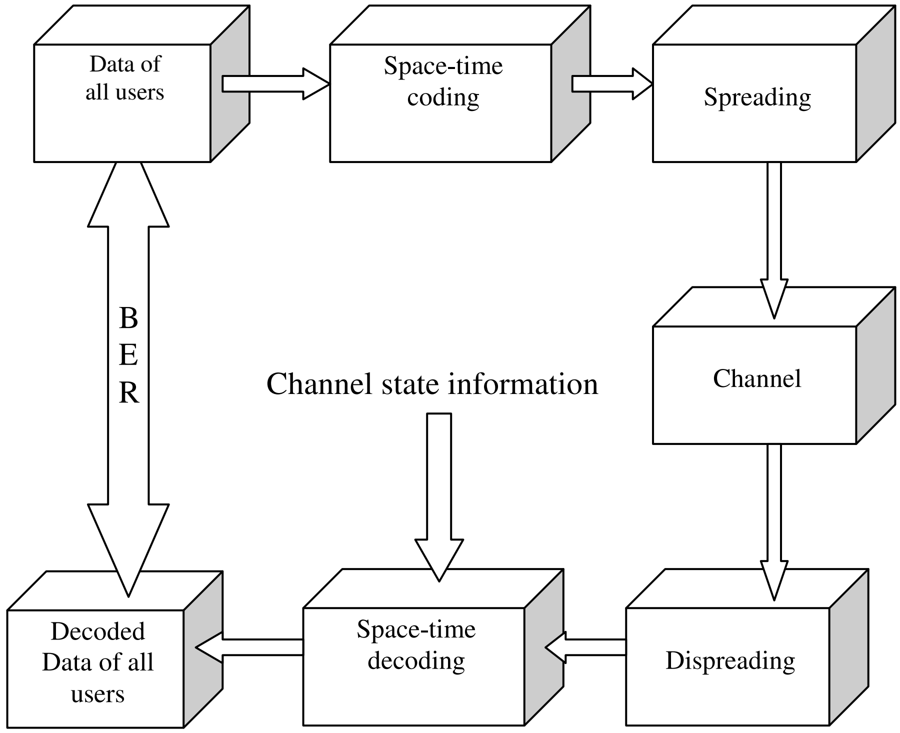

Figure 5.1 Simulation system model

The channel gains and signal amplitudes are assumed to be known at the receiver. It is also assumed that the users are moving so slowly such that the effect of Doppler shift is negligibly small. At the receiver, the signals from different users of the system are assumed to have the same amplitudes or the same power levels.

5.2 Performance Analysis

In this section, the performance of the optimal decoder and the proposed receiver will be studied using simulation carried out in matlab. Comparative performances of the proposed receiver, the decorrelator detector and the conventional single user receiver are also simulated.

5.2.1 Comparative Performance at Low Cross Correlation

In this section, the performances of the optimal decoder and the proposed receiver have been studied at low cross correlation of the spreading sequences. Just for comparison, a theoretical channel supporting two users $(K = 2)$ in which the multiuser interference power is controlled by a single cross correlation parameter has been considered. This means, for the proposed receiver, there are two decoders each one decoding only one user.

The simulation result in figure 5.2 shows the performance curves of the optimal decoder and the proposed receiver at a cross correlation value of 0.16 as the SNR varies from 0 to 30 dB. The single user bound is used as a reference. As can be seen in figure 5.2, the proposed receiver performs very close to the optimal decoder with a reduced complexity.

This performance result is as expected, because at low cross correlation of the spreading sequences, the influence of multiple access interference is minimal.

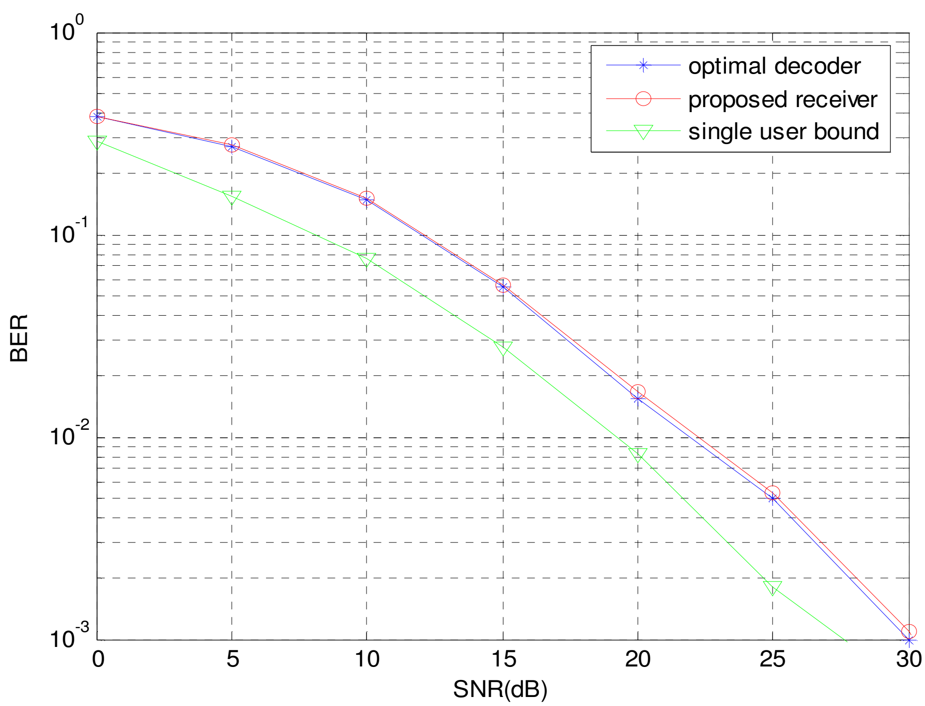

Figure 5.2 Performance at low cross correlation $\rho = 0.16$

5.2.2 Comparative Performance at High Cross Correlation

In this section, the simulation in section 5.2.1 is repeated at a cross correlation of value of 0.36. The corresponding complexities are also analyzed. In figure 5.3, we see that when the cross correlations of the spreading sequences is high, the performances of both the optimal and the proposed receivers are degraded. This degradation in performance is because of the influence of the multiuser interference which becomes dominant at high cross correlation of the spreading sequences.

The performance of the proposed low complexity receiver is still closer to the optimal one, but comparatively gets poorer at high cross correlation of the spreading sequences and this is the price to be paid in an attempt to reduce the complexity.

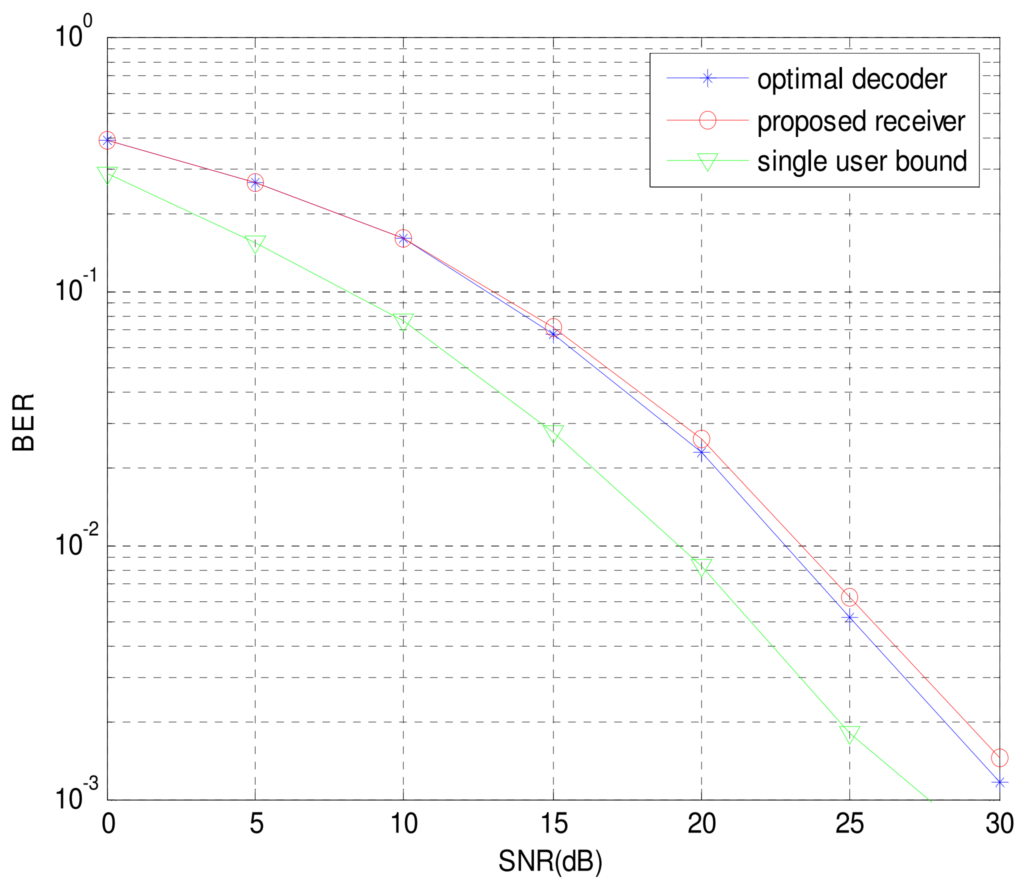

Figure 5.3 Performance at a cross correlation value of $\rho = 0.36$

Since each user encoder is a four-state space-time encoder, from equation (4.32), the optimal decoder requires a total memory of $M_{opt} = 48$ and from equation (4.33), the proposed receiver requires a total memory of M_{prop}=16. Then the optimal decoder requires an order of memory three times greater than the proposed receiver. Since each user has two antennas and there are four branches that leave a state, from equation (4.34) $\Gamma = 23$ .The number of operations of the optimal receiver per symbol of all users from equation (4.35) is therefore $O_{opt} = 1104$ and from equation (4.36), the number of operation of the proposed receiver is $O_{prop} = 184$, which is one sixth as much as the operations of the optimal decoder.

5.2.3 Performance with Varying Number of Users per Group Decoder

In this section, the performance of the proposed receiver with varying number of users per group decoder has been simulated.

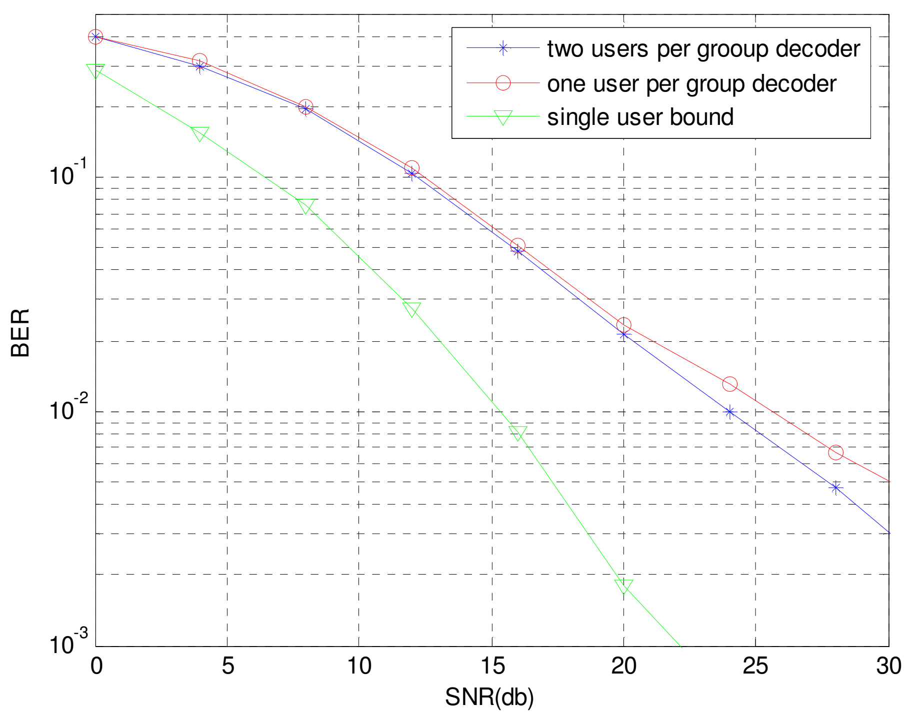

Figure 5.4 Performance with varying number of users per group decoder

For the simulation, a total of four users are considered ($K = 4$) and the performance of the proposed receiver with two users per group decoder is compared with its performance with one user per group decoder (simplest form). For the simulation result in figure 5.3, the correlation matrix is given below.

$$R = \begin{bmatrix} 1.0000 & 0.0968 & 0.3548 & 0.0323 \\ 0.0968 & 1.0000 & 0.0323 & 0.3548 \\ 0.3548 & 0.0323 & 1.0000 & 0.0968 \\ 0.0323 & 0.3548 & 0.0968 & 1.0000 \end{bmatrix}$$

As can be seen in figure 5.4, as the number of users in the group decoder increases, the performance of the proposed receiver gets better but this is with an increased complexity which in this case is twice. On the other hand, as the number of users per group decoder is decreased, complexity will be lowered at the expense of performance degradation.

5.2.4 Comparative Performance with the Decorrelator Detector

In this section, the performance of the proposed receiver has been compared with the performance of the decorrelator detector. It is assumed that there are two active users

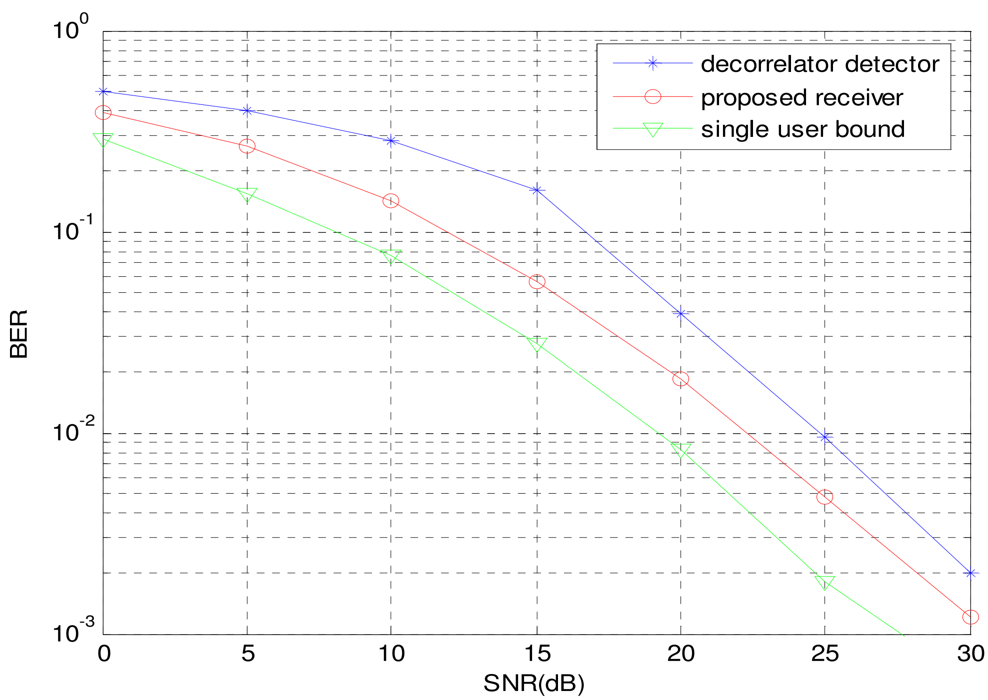

Figure 5.5 Performance at a cross correlation value of $\rho = 0.16$

and the proposed receiver has two decoders each on decoding a single user. The simulation is carried out at a cross correlation value of 0.16. From figure 5.5 we can see that the performance of the decorrelator detector is poor specially at low signal to noise ratio and this is because when the decorrelator detector carries out the decorrelation operation, it thereby increases the noise variance.

5.2.5 Comparative Performance with the Conventional Receiver

In this section, the performances of the proposed receiver and the conventional single user matched filter receiver are simulated and compared at low and high cross correlation values of the spreading sequences. The simulation is carried out at cross correlation values of 0.16 and 0.36. Figure 5.6 shows the performance curves of the conventional receiver and the proposed receiver at a cross correlation value of 0.16. The single user bound is used as a reference as usual. Even at low cross correlation values, the performance of the conventional single user receiver is poor since it considers the multiple access interference as noise..

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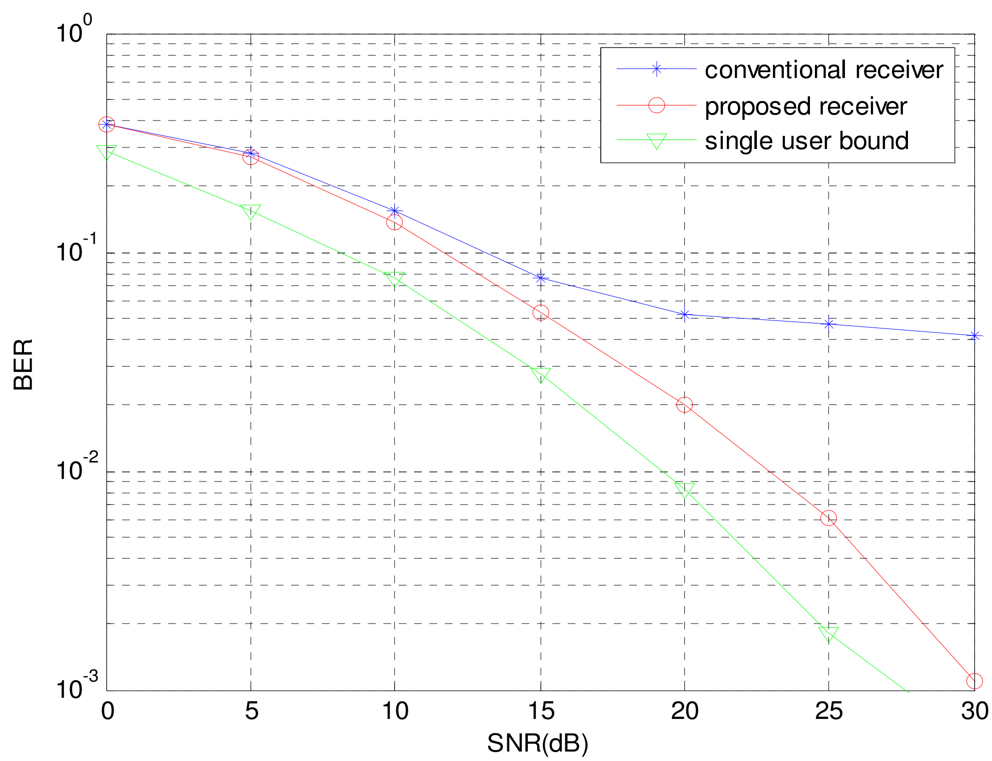

Figure 5.6 Performance at a cross correlation value of $\rho = 0.16$

Figure 5.7 shows the performance curves at a cross correlation value of 0.36. From figure 5.7, we see that the conventional single user receiver is totally useless at high cross correlation values of the spreading sequences

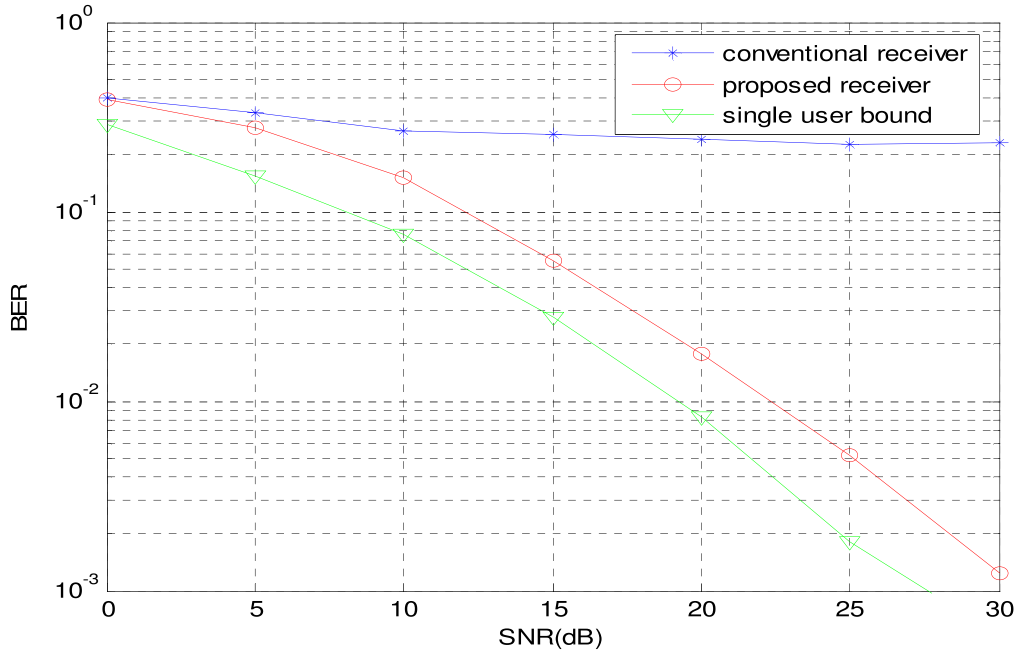

Figure 5.7 Performance at a cross correlation value of $\rho = 0.36$

5.2.6 Comparative Performance as a Function of Cross Correlation

In this section, the performances of the proposed receiver and the optimal decoder are provided as a function of the cross correlation values at an SNR value of 20dB. A theoretical channel consisting of two users is considered and the interference power is controlled by a single cross correlation parameter. As we can see in figure 5.8, at low cross correlation of the spreading sequences, the proposed receiver performs very close to the optimal decoder and at high cross correlation values, the performance of the proposed receiver becomes comparatively poorer because at high cross correlation values of the spreading sequences, there is a great deal of multiuser interference that in turn causes the error propagation.

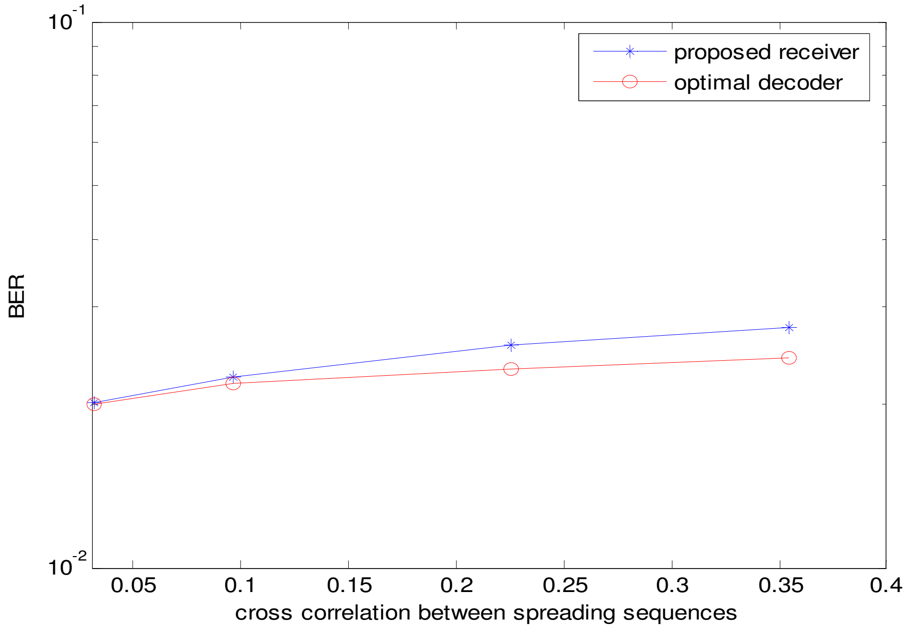

Figure 5.8 Performance as a function of cross correlation at SNR=20dB

5.3 Results of Related woks

Mark C.Reed, using iterative interference cancellation and antenna arrays at the receiver, has shown that it is possible to approach to a theoretical limit by multiple iterations. But iterative receivers are not convenient for real time applications like voice because of the time delays incurred due to the iterations. Fig 5.9 shows his simulation results.

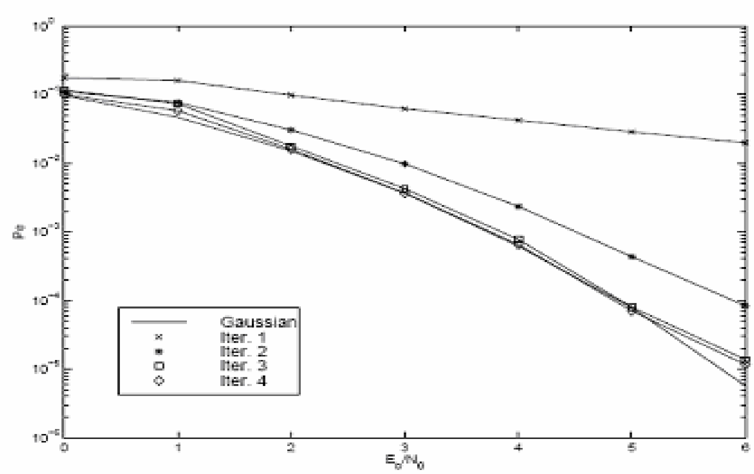

Figure 5.9 Multi-path combining receiver performance with multi- path DS-CMDA and antenna arrays

Mincheol Park, using parallel interference cancellation and with BPSK modulation was be able to lower the error rate down to 10^{-3} in a flat Rayleigh fading channel at a signal to noise ratio of about 25dB. But he didn't address the question as to how high data rate communication links can be achieved in CDMA systems. His simulation results are shown in Fig 5.10.

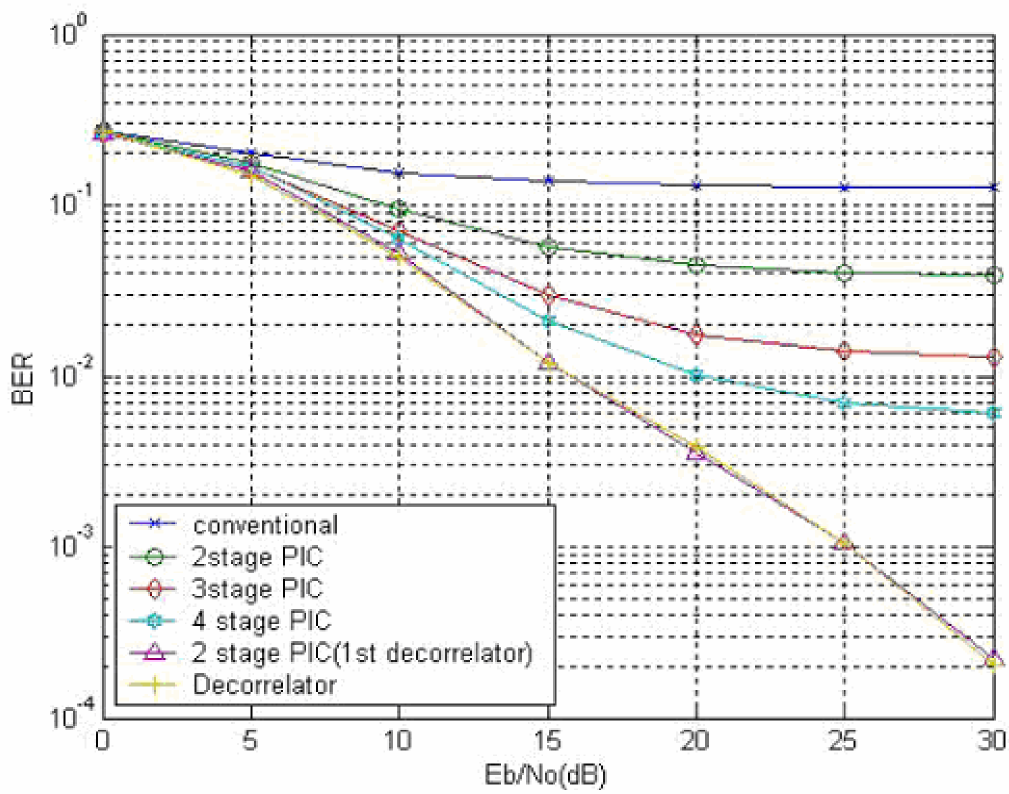

Figure 5.10 Performance with perfect channel estimation under flat fading channels

Chapter 6

Conclusions and Directions for Future Work

In this chapter, conclusions based on the theoretical ground and the simulation results in chapter 5 will be presented. Directions for future work are also suggested.

6.1 Conclusions

The proposed low complexity multiuser receiver performs near to the optimal decoder even at high cross correlation values of the spreading sequences with a reduced complexity. specially when the cross correlation between the spreading sequences is low, which is often the case in CDMA applications, the proposed low complexity multiuser receiver performs very close to the optimal decoder such that the performance curves seem to overlap.

For the proposed receiver, when the number of users to be decoded by a group decoder is decreased, complexity will be reduced but performance is degraded. On the other hand, when the number of users to be decoded by a group decoder is increased, performance will be improved with an increased complexity. Hence, the number of users to be decoded by a group decoder is a design parameter that needs to be adjusted for a best tradeoff.

The reduction in the complexity of this low complexity multiuser receiver is not without any price. The performance of the proposed multiuser receiver is always inferior to the performance of the optimal multiuser receiver and this degradation in performance is the price to be paid in an attempt to reduce complexity. This price is not too expensive to afford as compared to the reduction in complexity that is being made.

Thus, the proposed multiuser receiver is a very good alternative for space-time coded DS-CDMA with adjustable tradeoffs between performance and complexity.

6.2 Future Work Directions

This work assumes synchronous DS-CDMA, but the uplink of mobile DS-CDMA is generally asynchronous and the performance of the proposed multiuser receiver for asynchronous DS-CDMA can be one of the research areas in the future.

In this thesis work, perfect power control is also assumed and the power levels of the received signals of all users at the receiver are equal. But in reality, the power levels at the receiver are different which results in near-far problem and the performance of the proposed multiuser receiver in the face of near-far effect can be studied further.

It is also assumed that the receiver has perfect knowledge of the channel coefficients. But in real applications, the channel coefficients should be estimated which introduces a further reduction in performance. So the performance of the proposed receiver with channel estimation can be studied for a more realistic performance evaluation.

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