High Capacity CDMA and Collaborative Techniques

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Declarations

I, Indu Lal Shakya hereby certify that the materials presented in this thesis are my own work, except where indicated. Any information or materials that are not my own creations are indicated explicitly as references including their origins, publication dates.

I also declare that this thesis has not been submitted, either in the same or different form to this or any other university for a degree.

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Summary

The thesis investigates new approaches to increase the user capacity and improve the error performance of Code Division Multiple Access (CDMA) by employing adaptive interference cancellation and collaborative spreading and space diversity techniques. Collaborative Coding Multiple Access (CCMA) is also investigated as a separate technique and combined with CDMA. The advantages and shortcomings of CDMA and CCMA are analysed and new techniques for both the uplink and downlink are proposed and evaluated.

Multiple access interference (MAI) problem in the uplink of CDMA is investigated first. The practical issues of multiuser detection (MUD) techniques are reviewed and a novel blind adaptive approach to interference cancellation (IC) is proposed. It exploits the constant modulus (CM) property of digital signals to blindly suppress interference during the despreading process and obtain amplitude estimation with minimum mean squared error for use in cancellation stages. Two new blind adaptive receiver designs employing successive and parallel interference cancellation architectures using the CM algorithm (CMA) referred to as 'CMA-SIC' and 'BA-PIC', respectively, are presented. These techniques have shown to offer near single user performance for large number of users. It is shown to increase the user capacity by approximately two fold compared with conventional IC receivers. The spectral efficiency analysis of the techniques based on output signal-to interference-and-noise ratio (SINR) also shows significant gain in data rate. Furthermore, an effective and low complexity blind adaptive subcarrier combining (BASC) technique using a simple gradient descent based algorithm is proposed for Multicarrier-CDMA. It suppresses MAI without any knowledge of channel amplitudes and allows large number of users compared with equal gain and maximum ratio combining techniques normally used in practice.

New user collaborative schemes are proposed and analysed theoretically and by simulations in different channel conditions to achieve spatial diversity for uplink of CCMA and CDMA. First, a simple transmitter diversity and its equivalent user collaborative diversity techniques for CCMA are designed and analysed. Next, a new user collaborative scheme with successive interference cancellation for uplink of CDMA referred to as collaborative SIC (C-SIC) is investigated to reduce MAI and achieve improved diversity. To further improve the performance of C-SIC under high system loading conditions, Collaborative Blind Adaptive SIC (C-BASIC) scheme is proposed. It is shown to minimize the residual MAI, leading to improved user capacity and a more robust system. It is known that collaborative diversity schemes incur loss in throughput due to the need of orthogonal time/frequency slots for relaying source's data. To address this problem, finally a novel near-unity-rate scheme also referred to as bandwidth efficient collaborative diversity (BECD) is

proposed and evaluated for CDMA. Under this scheme, pairs of users share a single spreading sequence to exchange and forward their data employing a simple superposition or space-time encoding methods. At the receiver collaborative joint detection is performed to separate each paired users' data. It is shown that the scheme can achieve full diversity gain at no extra bandwidth as inter-user channel SNR becomes high.

A novel approach of 'User Collaboration' is introduced to increase the user capacity of CDMA for both the downlink and uplink. First, collaborative group spreading technique for the downlink of overloaded CDMA system is introduced. It allows the sharing of the same single spreading sequence for more than one user belonging to the same group. This technique is referred to as Collaborative Spreading CDMA downlink (CS-CDMA-DL). In this technique T-user collaborative coding is used for each group to form a composite codeword signal of the users and then a single orthogonal sequence is used for the group. At each user's receiver, decoding of composite codeword is carried out to extract the user's own information while maintaining a high SINR performance. To improve the bit error performance of CS-CDMA-DL in Rayleigh fading conditions, Collaborative Space-time Spreading (C-STS) technique is proposed by combining the collaborative coding multiple access and space-time coding principles. A new scheme for uplink of CDMA using the 'User Collaboration' approach, referred to as CS-CDMA-UL is presented next. When users' channels are independent (uncorrelated), significantly higher user capacity can be achieved by grouping multiple users to share the same spreading sequence and performing MUD on per group basis followed by a low complexity ML decoding at the receiver. This approach has shown to support much higher number of users than the available sequences while also maintaining the low receiver complexity. For improved performance under highly correlated channel conditions, T-user collaborative coding is also investigated within the CS-CDMA-UL system .

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List of Publications

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List of Abbreviations

Abbreviations	Definition
3G	Third Generation
4G	Fourth Generation
AWGN	Additive White Gaussian Noise
BA	Blind Adaptive
BA-PIC	Blind Adaptive Parallel Interference Cancellation
BASC	Blind Adaptive Subcarrier Combining
BA-SIC	Blind Adaptive Successive Interference Cancellation
BC	Broadcast Channel
BECD	Bandwidth Efficient Collaborative Diversity
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
C-BASIC	Collaborative Blind Adaptive Successive Interference Cancellation
CCMA	Collaborative Coding Multiple Access
CDMA	Code Division Multiple Access
CM	Constant Modulus
CMA	Constant Modulus Algorithm
CMA-SIC	Constant Modulus Algorithm Successive Interference Cancellation
CS-CDMA-DL	Collaborative Spreading CDMA Downlink
CS-CDMA-UL	Collaborative Spreading CDMA Uplink
CSI	Channel State Information
C-SIC	Collaborative Successive Interference Cancellation
C-STS	Collaborative Space-Time Spreading
CV-CCMA	Complex Valued Collaborative Coding Multiple Access
dB	Decibel
DCOR	Decorrelator
DS CDMA	Direct Sequence Code Division Multiple Access
E{}	Expectation
E_b/N_0	Energy per Bit over Noise Ratio
EGC	Equal Gain Combining
FDMA	Frequency Division Multiple Access
FLOP	Floating Point Operation
GA	Gaussian Approximation
GSM	Global System For Mobile Communications
IC	Interference Cancellation
$\Im$	Imaginary

Abbreviations	Definition
IQ	In-phase and Quadrature
ISI	Inter Symbol Interference
LLN	Law of Large Numbers
LMS	Least Mean Squares
MA	Multiple Access
MAC	Multiple Access Channel
MAI	Multiple Access Interference
MAP	Maximum Aposteriori
MC-CDMA	Multicarrier Code Division Multiple Access
MF	Matched Filter
MIMO	Multiple Input Multiple Output
ML	Maximum Likelihood
MMSE	Minimum Mean Square Error
MMSEC	Minimum Mean Square Error Combining
M-PSK	M-ary Phase Shift Keying
M-QAM	M-ary Quadrature Amplitude Modulation
MRC	Maximum Ratio Combining
MSE	Mean Squared Error
MUD	Multiuser Detection
OCDMA	Orthogonal Code Division Multiple Access
OCDMA/OCDMA	Orthogonal CDMA/ Orthogonal CDMA
OFDM	Orthogonal Frequency Division Multiplexing
OFDMA	Orthogonal Frequency Division Multiple Access
OWMA	Orthogonal Waveform Multiple Access
PIC	Parallel Interference Cancellation
PN	Pseudo-noise
QOS	Quasi Orthogonal Sequences
QPSK	Quadrature Phase Shift Keying
$\Re$	Real
RLS	Recursive Least Squares
SDMA	Space Division Multiple Access
SIC	Successive Interference Cancellation
SINR	Signal to Noise and Interference Ratio
SISO	Single Input Single Output
SNR	Signal to Noise Ratio
STBC	Space-Time Block Code
STS	Space-Time Spreading
STTC	Space-Time Trellis Code
TDMA	Time Division Multiple Access
$\text{var}\{\}$	Variance
VBLAST	Vertical Bell Layered Space-Time
WBE	Welch Bound Equality
WCDMA	Wideband Code Division Multiple Access
WH	Walsh Hadamard
WIMAX	Worldwide Interoperability for Microwave Access
ZF	Zero Forcing

List of Tables

List of Figures

Contents

Chapter 1

Introduction

Mankind, since its early development, has always aspired to find an efficient way to communicate with each other using some medium. And, with the advent of new ideas, more and more efficient techniques have either replaced the old techniques or completely changed the ways we use for communications. Among many ways, communication over the wireless medium has become an essential technology with profound impact on modern life. The revenue generated from wireless data communications products and services has risen to phenomenal rate due primarily to the growth of Internet in recent years. The success of wireless communications technology seen in last two decades is made possible by tremendous amount of efforts put forth by many researchers. Careful study of this subject area has led to many ground breaking inventions such as capacity approaching Turbo Codes, MIMO-VBLAST and more [1].

The wireless medium in contrast to its wired counterpart, has some interesting properties. These are: 1) The information communicated over the medium is propagated in all directions such that anyone tuned to transmission frequency can recover the information (i.e. there is no centralized control over the medium). 2) Depending upon its location, each user's receiver obtains the information attenuated uniquely by its own channel. More specifically, the attenuation (fading) effects caused by multipath propagation over wireless channels brings many practical challenges for reliable communication over such channels. Despite the challenges, a new wave of ideas have recently emerged that even make use of the aforementioned properties of wireless channels to significantly increase the capacity of communications [2].

The multiuser wireless communications is a very interesting and challenging area of digital communications and is the focus of study in this thesis. There are two main approaches for the communications: Infrastructure based and Ad-hoc . The former usually employs cellular architecture [3] to enable transmission from low power mobile handsets and some multiple access schemes to offer communication links to the simultaneous users. An example of a typical cellular wireless telephone system is shown in Figure 1.1. Mobile users carrying handsets use uplink (also called

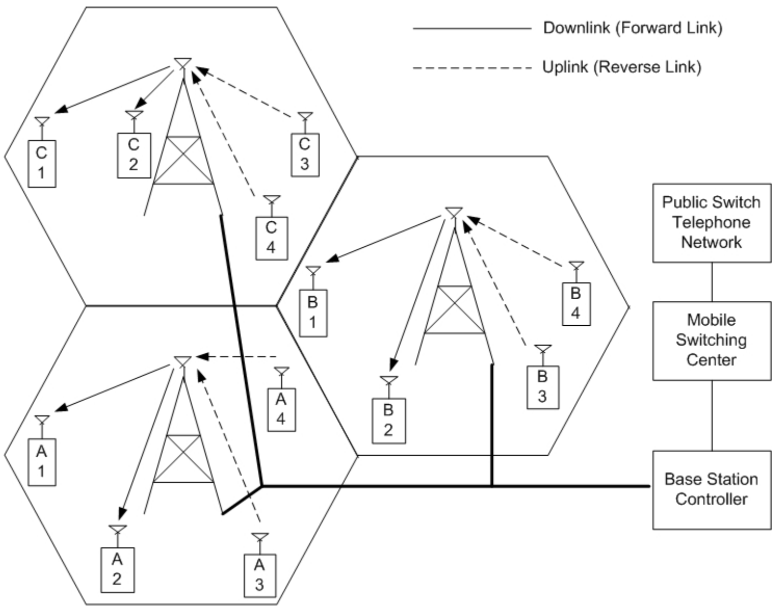


Figure 1.1: An example of a typical cellular wireless telephone system

reverse link) channels to transfer information (call) via a base station to other users; the base stations on the other hand, use downlink (also called forward link) channels to deliver the information to the users from other users. The information flow to/from different base stations are controlled by Base Station Controllers (BSC) and then processed at the Mobile Switching Center (MSC). The calls made to users outside the cellular networks are sent to a Public Switched Telephone Network (PSTN). While the figure is used to explain the operation of a typical voice call made through a cellular wireless systems, more advanced techniques for data transmission over current and future cellular wireless systems also employ similar system architecture. The simple model of Figure 1.1 will be used throughout the thesis for the study and analysis of different techniques to be proposed.

1.1 Background and Motivations

The third generation (3G) cellular wireless communication is now a well established technology for providing voice, multimedia and data communications worldwide [4]. The core technique that has derived success of 3G is CDMA or Code Division Multiple Access. The CDMA is a multiple access scheme that allows simultaneous transmission of data over the same frequency band by many users with use of so called unique pseudorandom spreading codes [5]. In contrast to other narrowband multiple access schemes such as TDMA/FDMA (described in Chapter 2), CDMA is inherently more resilient to narrowband interference, noise, and multipath fading. The universal frequency reuse, exploitation of voice activity of users' calls are the benefits of CDMA, making it the clear winner in terms of network capacity. It is long known that the theoretical capacity of

CDMA with multiuser detection is much higher than that is achieved currently using conventional correlator or RAKE receivers [6].

There are however many important challenges that need to be addressed to achieve a robust system with high user capacity, bandwidth efficiency and reliable communications. From economical viewpoint, to maximize the revenue from the installed systems, cellular operators are looking for more efficient techniques for multiple access so that the number of supportable subscribers within given geographical location is increased. Parallel to the demand of increased number of supportable subscribers, higher spectral efficiency of allocated links is becoming another prime necessity. This is due to the need to support realtime multimedia communication, which requires much higher throughput than traditional voice traffics. Since the radio spectrum is a very limited resource, it is of utmost interest to look for techniques that achieve higher capacity (spectral efficiency) from the existing resources.

New techniques for increasing the capacity of CDMA such as multiuser detection and interference cancellation, though offer much improved capacity compared with conventional detection techniques, there are however many practical issues and shortcomings of these techniques. Therefore, there is a need for a careful study of the key problems limiting their performance and find simple and practical propositions. Techniques to achieve the high capacity multiple access can also be found by following a view on the problems from a different angle. Examples of such ideas are user cooperation [7, 8], collaborative coding and decoding [9, 10, 11, 12, 13, 14, 15] and combination of these ideas with current multiuser detection techniques.

1.2 Research Aims and Objectives

Inspired by the potential high capacity achieved from multiuser detection techniques as noted earlier, this thesis aims to provide effective solutions to the problems of improving the capacity of multiple access by investigating the possibilities of incorporating the techniques into novel system design frameworks. The main aims and objectives of this study are outlined as follows:

1. Provide literature review on recent development on multiuser detection techniques and identify the main problems and practical issues for achieving high user capacity in CDMA systems. Propose new and more efficient approaches and evaluate their performances.

2. Propose new ideas such as user collaboration and cooperative diversity to support higher number of users and improve the error performance in practical wireless environments. Provide theoretical and simulation performance analysis to support the ideas.

To achieve these aims, following research objectives have also been identified:

1. Investigate blind adaptive techniques for CDMA multiuser detection and interference cancellation to overcome the practical problems of channel estimation while also providing robust operation in non ideal user power and channel estimation conditions. Extend the blind adaptive approach for subcarrier combining method to suppress interference and better exploit the frequency diversity in MC-CDMA.

2. Investigate the idea of collaborative space diversity for CCMA and CDMA to improve the performance in fading channel conditions. Combine the collaborative scheme with low complexity interference cancellation techniques to maximize the achievable diversity. Make use of the idea of grouping several users to share the same spreading sequence to address the bandwidth efficiency loss for collaboration in CDMA.

3. Investigate higher user capacity schemes for CDMA using the idea of collaborative coding and spreading. Exploit the independent fading channels of users along with group collaborative spreading to achieve user multiplexing and improve the bandwidth efficiency. Provide a unified system design for the collaborative spreading CDMA to offer robust detection performance under different channel correlation conditions.

1.3 Contributions and Outline of the Thesis

The materials in the rest of thesis are organized as follows. **Chapter two** is the technical background and literature review. The Chapter reviews the basic principles and problems of many important techniques used for multiuser systems. Based on the review on each problem, conclusions are made and how it is addressed is also briefly discussed. The next three chapters are the major areas of contributions consisting of work to achieve the objectives outlined in the previous section. Summary of each contribution and organization of the remaining part of the thesis are enumerated below:

1. **Chapter Three** consists of three contributions addressing the MAI in uplink of CDMA. Summary of each is given below:

   In Section 3.2, an effective design of successive interference cancelation (SIC) multiuser detector for Direct Sequence CDMA using Constant Modulus Algorithm (CMA) also referred to as 'CMA-SIC' [16] is proposed. SIC receivers are simple in structure and robust in combating MAI and near far effect. However, two of the main practical limitations of SIC are the error propagation and imperfect channel estimation. To overcome these problems, a simple adaptive CMA algorithm is employed within SIC despreader to provide reliable estimate of the desired signal amplitude and for mitigating the effect of fading channel. Modeling and simulation of the proposed detector shows impressive single user bound reaching performance in fading channel with AWGN and robustness in different system loading and near far conditions.

In Section 3.3, an effective design of multistage PIC receiver using blind adaptive despreader and pre-respreader interference estimator also referred to as 'BA-PIC' [17] for uplink of CDMA is proposed and analysed. A novel algorithm is designed that exploits constant modulus (CM) property of the users' transmitted signals and inherent channel condition to perform adaptive despreading based on minimum error variance criteria. This is carried out by blind adaptive weighting of each chip signal for accurate tracking of the desired user's signal power and hence for more improved data detection at the output of each stage of PIC. Furthermore, the despreader weights are used within the adaptive pre-respreader interference estimation and cancellation to obtain online scaling factors during every symbol period, without any knowledge of users' channels or the use of training sequences. It is found that this way of estimation is optimal in minimum mean squared error (MMSE) sense, and hence, significant reduction in interference and noise variance is observed in detection and estimation of the desired users' signals compared with conventional PIC. Bit error probability of the proposed PIC is obtained using Gaussian Approximation method. Extensive simulation results are shown which demonstrate impressive performance advantage in fading environments, high system loading, and severe nearfar conditions.

Section 3.4 presents a new subcarrier combining technique for MC-CDMA receiver in mobile Rayleigh fading channel. It exploits the structure formed by repeating spreading sequences of users on different subcarriers to simultaneously suppress multiple access interference (MAI) and provide implicit channel tracking without any knowledge of the channel amplitudes or training sequences. This is achieved by adaptively weighting each subcarrier in each symbol period by employing a simple gradient descent algorithm to meet the constant modulus (CM) criterion with judicious selection of step-size. Improved BER and user capacity performance are shown with similar complexity in order of $\mathcal{O}(N)$ compared with conventional maximum ratio combining and equal gain combining techniques even under high channel Doppler rates. Extensive performance evaluation under different system loading, type of spreading sequences and step size are carried out.

2. In **Chapter Four**, four new contributions are presented to tackle the problems of providing space diversity in the uplink of non-orthogonal multiple access schemes e.g CCMA and CDMA.

   In Section 4.2, a simple transmitter diversity scheme is presented for uplink CCMA with users employing more than one antenna and a single antenna base-station receiver. The users transmit their codewords over two consecutive periods from separate antennas forming multi-user fading composite signal at the receiver. Joint detection and decoding is used to recover the transmitted data of the users using collaborative codeword combinations. The analysis and simulation results show that the scheme offers same diversity order performance as that of the CCMA with receive antenna diversity.

Also a novel user cooperation diversity scheme [18] is introduced for uplink CCMA system in Section 4.3. In this scheme, the cooperating users transmit their own collaborative codewords and that of their partners'. The base-station receives two copies of the composite codeword signals via two different channels. Joint detection and decoding is employed at the receiver to recover the individual users' codewords and their data based on the minimum distance criteria of the collaborative composite codeword combinations. The bit error performance is analysed in flat Rayleigh fading channels with AWGN, and shown to provide significant gain compared to CCMA with no cooperation, near to that of CCMA using dual receive diversity.

Achieving full diversity gain from users' cooperation in a non-orthogonal multiuser channels, such as uplink of CDMA, is a challenging problem due to MAI. A joint scheme employing user collaboration diversity and SIC referred to here as Collaborative SIC (C-SIC) [19, 20] is proposed and analysed in Section 4.4 to address this problem. It has been shown that the performance of cooperation scheme can be improved considerably by C-SIC that removes the effects of MAI from each cooperating user's signal. The residual MAI due to imperfect cancellation of SIC, however, limits the achievable performance of the cooperation scheme. Collaborative Blind Adaptive SIC (C-BASIC) [21] using the CMA-SIC technique [16] is then investigated for achieving improved diversity gain from the user collaboration. The CMA-SIC method minimizes the variance of decision variable signal at each stage with adaptive despreading and more accurate MAI estimation for the cancellation. The results show that the C-BASIC scheme provides considerable improvement in capacity of uplink CDMA compared to that with C-SIC as well as collaborative matched filter (C-MF) receivers.

Finally, in Section 4.5 a near-unity-rate cooperative scheme also referred to as 'Bandwidth Efficient Collaborative Diversity' (BECD) is presented for uplink of CDMA. Under this scheme, a small number of users are grouped and assigned the same spreading sequence. The technique utilizes the fact that, the wireless propagation environment provides each user with distinct channel gain, which can be utilized to demodulate and decode the received composite signal to recover data from multiple independent users sharing the same fraction of bandwidth. Hence, the bandwidth efficiency loss incurred due to decode and forward processes performed at the cooperating user nodes is offset by the saving in the required number of sequences. Two possible methods for BECD scheme with full duplex operation capable user nodes are proposed. The first is based on simple forwarding of users' data from cooperating users and joint detection of superposition of co-spread users at the base. In the second, the cooperating users forward space-time encoded signals and allows for simple and lower complexity ML detection of the co-spread users' data. Performance of the techniques are evaluated under different ratios of inter-user channel SNR gain. It is shown that diversity performance within $1dB$ of the Alamouti scheme can be achieved when the ratios inter-user channel gains are high. Furthermore, a more practical BECD scheme with half duplex nodes with a relay assigned for each user is investigated. Similar gain is verified and comparisons are made.

3. **Chapter Five** consists of three important contributions to address the user capacity limitations of CDMA. A new idea of 'User Collaboration' is introduced for downlink and uplink of CDMA. Issues arising in practice such as propagation environments and channel estimation are also taken in to considerations.

   In Section 5.3, a new scheme called Collaborative Spreading for the downlink of CDMA (CS-CDMA-DL) is proposed to allow the sharing of the same spreading sequence by more than one user. In particular it addresses the problem of user overloading and maintain the use of the same set of available orthogonal sequences and simple receiver structure. In this scheme, the total K users are divided into G-group each of T-user which are collaboratively coded to form uniquely decodable composite codewords. These codewords are spread using a single sequence to perform the CDMA function between the groups. At the receiver, a low complexity ML joint detection and decoding is carried out over a small set of allowed composite codewords to recover the desired user's data. Theoretical and simulation performance analysis of the BER and user capacity are presented in different channel conditions. It is shown that the proposed collaborative spreading is a simple and very effective means for extending the user capacity at the cost of small increase in signal-to-noise ratio (SNR) compared with non-overloaded CDMA. It can achieve SINR and higher overloading ratio compared with Orthogonal CDMA/Orthogonal CDMA schemes and other group orthogonal CDMA schemes.

   To improve the error performance of the CS-CDMA-DL in fading channel conditions a new transmit diversity scheme referred to as Collaborative Space-Time Spreading (C-STS) is introduced in Section 5.4. It combines the ideas of collaborative spreading and space-time coding techniques to improve the performance for the downlink of overloaded CDMA. The scheme operates by performing first collaborative coding and spreading of each group of T-user data and then performing space-time transmission on multiple antennas. The scheme has shown to significantly enhance the BER performance of users compared with the previously proposed single antenna CS-CDMA-DL scheme. In other words, the C-STS is shown to offer full second order diversity for two transmit antennas. At the user overloading ratio of 3, it is shown to require only approximately 2 dB extra SNR compared with fully orthogonal Alamouti scheme and 3 dB compared to that with dual receive antenna diversity.

In Section 5.5, a new collaborative spreading for uplink of CDMA referred to as CS-CDMA-UL is presented to increase the number of users. The total K users in the system is first divided into G-group, each with T-user. The users within each group reuse the same spreading sequence to transmit their data by exploiting their unique channels. To successfully recover the data under such an overloaded condition, we propose to uniquely exploit the capabilities of linear MUD to perform suppression of MAI on the group basis for all K users using only G dimensional linear processing. Using the channel estimates, the users' data are then jointly detected using a low complexity ML detection method. Although co-channel interference (CCI) within each group necessitate a little added SNR compared with single user detection, it is easily compensated by the gain in terms of number of users i.e. reduction in MAI. It is shown that the total number of users supported with this scheme can be much higher than the available sequences while still outperforming conventional CDMA with high number of users. Collaborative coding is further incorporated to achieve robustness under correlated channel conditions. Using Gaussian Approximation, the probability of bit error analyses are presented, followed by the simulations results to affirm the gain. Furthermore, it is shown that the high capacity performance is retained under moderate channel estimation error conditions.

4. **Chapter Six** provides the conclusions of this research and recommendations for further investigations.

Chapter 2

Multiuser Wireless Communications

2.1 Introduction

This Chapter provides technical background and literature review of the thesis. It is organized as follows. The need and purpose of a multiple access scheme is described in Section 2.2. Section 2.3 is focused on providing overview of different aspects of capacity of multiuser channels. Two frequently used capacity terms are described, namely, the information theoretic capacity (sum spectral efficiency) in subsection 2.3.1, which together with the definition of user capacity in subsection 2.3.2 leads to a brief discussion of different multiple access schemes in Section 2.4. Two approaches to multiple access, namely, Orthogonal Waveform MA and Non-orthogonal Waveform MA are reviewed in subsection 2.4.1 and 2.4.2, respectively. Furthermore, the problems of an overloaded system and particularly using CDMA as the access method is discussed in subsection 2.4.3, which are motivated by the need to increase the user capacity in limited system bandwidth. It will be discussed that novel CDMA schemes proposed in this thesis can offer much higher capacity and SINR compared with different CDMA schemes and will be described in full details in Chapter 5.

Section 2.5 provides a review of multiuser detection techniques for the uplink of CDMA. It describes the principle and problems of conventional receiver and then discusses on general operation principles of different multiuser detection techniques that offer improvement over the conventional one. Some important practical issues faced by these MUD techniques are then discussed in subsection 2.5.2. The interference cancellation subclass of detection method for CDMA is given more focus and a survey of different approaches to overcome the problem of imperfect interference estimation and cancellation is provided. In this course, brief discussion on the main ideas behind the new contributions of this thesis to address the issues are also provided.

The remaining part of this Chapter is dedicated to new concepts such as user cooperation diversity along with the problem of detection in presence of co-channel as well as MAI users. It starts with discussion of basic concepts and technical details of fading channels in Section 2.6, followed by a review of different diversity techniques to combat the detrimental effects of fading in Section 2.7. In Section 2.8, the principles and practical issues of bandwidth efficiency loss and MAI in user cooperation diversity for uplink of CDMA is described. The idea of combining user cooperation diversity and multiuser detection that addresses the MAI, is discussed in subection 2.8.2. Finally, the summary of main ideas of this Chapter is provided in Section 2.9

2.2 The Need for Multiple Access Schemes

Future wireless systems will require for both higher spectral efficiency and user capacity. Furthermore, the mobile nodes require less complex detection techniques to maximize the battery life. To meet all these requirements, an efficient way of sharing the system resources or a multiple access scheme is required. The main purpose of a practical multiple access scheme is to allow maximum number users to simultaneous transmit their independent information to a common sink while requiring minimum complexity in demodulating the received signal [5]. The information carried over the transmission channels can be voice, digital data or multimedia messages. It is very important to make a clear distinction between the spectral efficiency of multiuser communication systems and another major objective of practical multiple access schemes that aim for higher user capacity. While high spectral efficiency is the ultimate figure of performance of any multiuser systems, achieving this does not necessarily mean that the objective of accommodating higher number of users is also met. Both are equally challenging to achieve in practice and motivates us to look for different new ideas and approaches for efficient multiple access. In this context, this thesis also deals with the problem of overloaded MAC and in particular a CDMA system. Compared with some other approaches that are highly complex but less efficient, some novel concepts on multiple access for overloaded channels are proposed and detailed in Chapter 5.5. To better understand these new ideas, basic concept of user capacity of different multiple access schemes are reviewed in the sequel.

2.3 Capacity of Multiple Access Channels

The capacity of multiuser wireless communication systems and different multiple access technologies are the topics of widespread interests and may have many practical implications for efficient system designs [22, 23]. Information Theoretic (IT) view of multiple access techniques and their capacity (number of bits per unit bandwidth) are discussed in early contributions by Gallager , El Gamal, Cover and Muelen, such as in [11], [24] and [13]. Unlike the IT approach, the user capacity centric approach [25], [26] are focused on maximizing the number of simultaneous users supported; which is more appealing in terms of generating more revenues i.e. users from limited system resources. Since the spectral efficiency is the ultimate performance metric of any wireless systems, the information theoretic perspective of MAC and the achievable rate region of different MA schemes are shown here in a simplified manner. The orthogonal waveform MA (OWMA) schemes, although are optimum in terms of minimizing outage, are suboptimal in terms of achievable rates [9]. It is known that superposition coding transmission by several cooperating users and successive interference cancellation at the receiver achieves rates much higher than orthogonal schemes [27]. A brief review of capacity of multiuser systems are carried out next.

2.3.1 Information Theoretic Capacity

The performance of multiuser systems are often described in terms of capacity region . This is defined as the sum of each user's maximum information rate that can be reliably transmitted over a unit bandwidth. For a K-user Gaussian multiple access channels (GMAC), the sum capacity C_{GMAC}, can be given by [6]

$$C_{GMAC} = \sum_{k=1}^{K} R_k; \Longrightarrow C_{GMAC} \leq \frac{1}{2}\log_2\left[1 + \frac{\sum_{k=1}^{K} P_k}{\sigma^2}\right] \tag{2.1}$$

where, R_k is rate of k^{th} user, $\frac{P_k}{\sigma^2}$ is the signal to noise ratio with P denoting signal power and σ^2 is the variance of noise over the system bandwidth. As noticed in (2.1), the sum rate of the MAC is less than the sum of individual user's rates. For example, the achievable rates for a 2-user GMAC is shown in Figure 2.1. Where, P_1 and P_2 are the power of user 1 and 2, respectively. The area within the pentagon defines the rate region of the Gaussian MAC. As can be seen, the individual rate of users R_1 and R_2 are effected by their power P_1 and P_2 along with their AWGN terms. Therefore the sum rate of the GMAC is always lower than that of single user system under the same power and bandwidth.

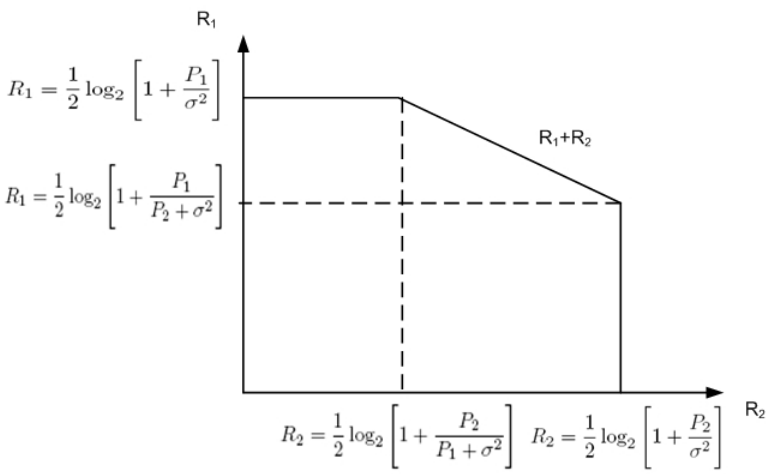


Figure 2.1: Achievable rates for a two user system in GMAC

In CDMA communications , each user uses the entire spectrum of the system by spreading its low rate data using a distinct spreading sequence **s** spanned over N chips. The well known signal model of synchronous CDMA $\mathbf{r} = \mathbf{SAb+n}$, as in [28], [29] is used here. where, **r**, **S**, **A**, **b** and **n**, are the received signal vector of dimension $1 \times N$, matrix of users' spreading sequences of size $K \times N$, diagonal matrix of users' amplitude, users' data vector, and noise vector, respectively. The sum capacity C_{spread} for such a system is given by [6]

$$C_{spread} = \frac{1}{2N} \log_2 \left[det\Big(\mathbf{I} + \frac{N}{\sigma^2}\mathbf{SPS}^T\Big) \right] \tag{2.2}$$

where, det denotes the determinant of a matrix, **I** is $K \times K$ identity matrix, $\mathbf{S}^T$ denotes transpose of matrix **S**, and $\mathbf{P}=\mathbf{AA}^T$ is diagonal matrix of users powers.

The capacity formulae (2.1) and (2.2) of Gaussian MAC and CDMA multiuser system set an ultimate limit on how much information can be transmitted over the channels. The users in the uplink CDMA system usually employ random sequences to transmit their information to the basestation receiver. The asymptotic spectral efficiency of such systems for different type of multiuser receivers at the basestation is investigated by Verdu and Shamai [6]. In Figure 2.2, the performance of different receivers obtained from [6] are shown. It can be seen that the spectral efficiency of the system with optimum receiver increases with number of users or system loading K/N. The conventional matched filter receiver shows much lower spectral efficiency and saturates after certain number of users. The performance of decorrelators and MMSE [29] are shown to be higher than MF at low system loading $K/N \leq 1$.

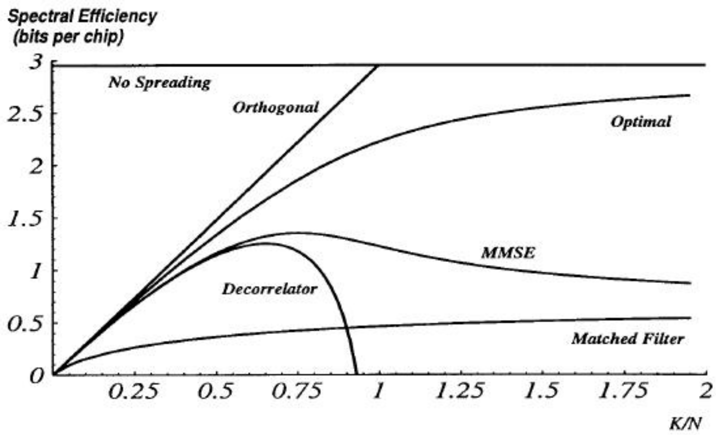


Figure 2.2: Large system spectral efficiency of randomly spread CDMA for different receivers at $E_b/N_0 = 10dB$ [6]

The information theory predicts that fading multiuser systems can offer higher sum capacity than that in AWGN channels, if the system resources (spreading sequences, channel power distribution, transmitter power control) are utilized efficiently [23]. The author investigated the performance of conventional decision feedback also known as SIC receivers in different fading channels with and without power ordering. The spectral efficiency analysis similar to [23] will be used to assess the performance of improved IC receivers [16],[17] proposed in Chapter 3.

It has been seen earlier that the multiuser systems incur loss in the spectral efficiency compared with equivalent single user Gaussian channels. This is because, when multiple users simultaneously access the common receiver, each user's signal, that is regarded as the desired signal, also contribute to the noise and interference components for other users. The optimum receiver jointly detects the signals of all users and as $K/N \longrightarrow \infty$, converges asymptotically to the capacity of single user channels [6]. The practical multiuser systems have to accommodate several tens to hundreds of users within a given system bandwidth while ensuring minimum interference for detection of users' data. In such cases, the employment of optimum receiver and SIC though they are known to achieve higher capacity, become impractical to implement due to their higher computational complexity. Therefore, the choice of a simple multiple access scheme that guarantees interference free access of users' information has become one of the main focus of practical multiuser wireless systems.

2.3.2 User Capacity

The user capacity of multiple access scheme is defined by the ratio of total number of users K to the number of degree of freedom N, i.e. $C_{user} = K/N$. The degree of freedom can be frequency bands, time slots or user separating codes. The channel is said to be under-loaded if $C_{user} \leq 1$ and overloaded if $C_{user} > 1$. When no multiple access scheme is employed and there are K users in the system, the user capacity is $C_{user} = K$. The conventional multiple access schemes, e.g. FDMA and TDMA, support maximum N users. Therefore, their user capacity is simply $C_{user} = 1$. Unlike these schemes, CDMA on the other hand, does not have hard limit on maximum allowable number of users. The system performance, which is dependent on SINR , dictates the maximum number of users K. CDMA also has many attractive advantages in practical cellular wireless systems, such as, universal frequency reuse, high resilience to jamming and exploitation of multipath fading for performance improvement. To motivate the discussion of novel techniques that increase the user capacity, brief review of different MA schemes are provided next.

2.4 Review of Multiple Access Schemes

Since there are large numbers of users accessing the MAC with their narrowband data, to minimize the cost of equipments, simple and safe way to avoid any interference among the users is often desirable. Wireless multiple access channels are subject to different distortions and non-idealities, such as: channel induced fading, nearfar problems due to unequal receiver power of users' signals and also interference due to delay spread of channels or non ideal cross correlation of individual users transmitted signals. The multiple access scheme that allocates orthogonal (non-overlapping) channel for each individual user for the purpose just reported are also broadly categorized as orthogonal waveform multiple access (OWMA) schemes. Conversely, a scheme which allocates each user non-orthogonal (overlapping) channel are termed Non-orthogonal waveform multiple access (NWMA).

2.4.1 Orthogonal Waveform Multiple Access Schemes

The multiple access using orthogonal channel for each user is very simple and effective technique and it has been the primary choice for traditional voice and low rate data traffics. Since the central receiver demodulates each user's signal separately without any interference from other users' signals, the error performance of such schemes resembles that of single user systems [5]. The orthogonal channels are obtained by dividing the total available bandwidth into finite number of non-overlapping frequency bands, time slots, or orthogonal codes. It can be noticed that the user capacity of OWMA is unity i.e., the maximum number of simultaneous users K supported is N. The most common OWMA multiple access schemes and their main characteristics are briefly reviewed next.

2.4.1.1 Frequency Division Multiple Access (FDMA)

FDMA scheme [5] employs N non overlapping frequency bands for accommodating data access for up to $K \leq N$ users as shown in Figure 2.3. To avoid possible interference because of carrier frequency offset due to imperfection in electronics components, the scheme also employs additional guard bands between adjacent frequency bands for the users. Since each user is allocated a single frequency band, the central receiver requires N separate demodulators, which may be very inefficient in terms of hardware requirements. In this discourse we are interested in the user capacity of FDMA schemes, as it can be noted that the FDMA scheme has the user capacity of $C_{user} = 1$.

2.4.1.2 Orthogonal Frequency Division Multiple Access (OFDMA)

The high speed data communications over realistic channels may incur severe inter-symbol interference (ISI) due to longer delay spread of channels that spans over several symbol periods. The ISI leads to irreducible error floor and equalization of channel becomes inevitable in such case. Since most equalization techniques require rather complex algorithm, a novel approach to FDMA, which is called Orthogonal FDMA [30], is now becoming more popular for broadband wireless communications. In OFDMA, each user is assigned one or more flat fading subcarriers (subbands) that preserve orthogonality even when the users' signals are transmitted through frequency selective fading channels. This considerably saves the computational complexity required for multi tap equalization of the wideband signals and user signal demodulation is performed by very efficient discrete Fourier transform at the receiver [30]. The use of OFDMA with channel coding is a promising technique for broadband wireless communications and has been adopted in many standards such as digital video broadcasting, cable networks and fourth generation wireless networks such as WIMAX [2]. The user capacity of this scheme is often $C_{user} \leq 1$, due to bandwidth loss due to insertion of cyclic prefix and guard interval etc.

2.4.1.3 Time Division Multiple Access (TDMA)

TDMA is popular multiple access scheme adopted in second generation cellular wireless network, such as GSM. The scheme accommodates multiple users by assigning non-overlapping time slots as shown in Figure 2.3. The data frame of multiple users are demultiplexed and demodulated separately at the central receiver. The scheme is more efficient than FDMA in the sense that it supports asymmetric user data traffic in uplink and downlink by varying the period of time slot assigned for the users. TDMA is also considered for third generation wireless cellular networks for providing full duplex capabilities for mobile users. Since TDMA uses orthogonal time slots for each user, its user capacity $C_{user} = 1$.

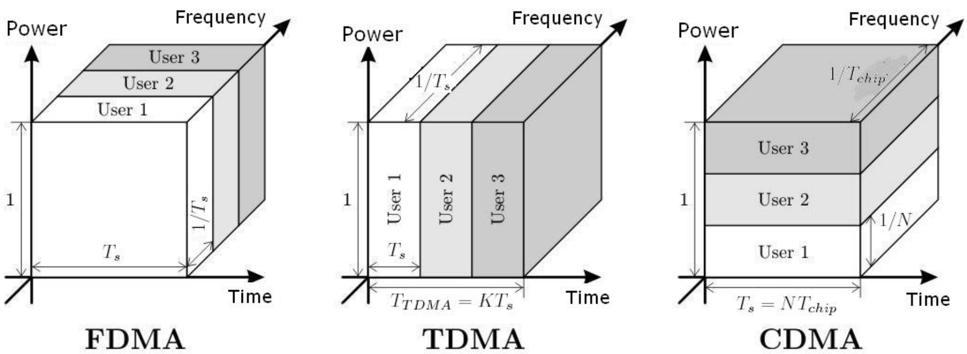


Figure 2.3: Orthogonal multiple access schemes

2.4.1.4 Code Division Multiple Access (CDMA)

The multiple access method of CDMA can be achieved by assigning distinct code to each user, where the user modulates its data using a distinct spreading code spanned over the entire frequency bandwidth as shown in Figure 2.3. The scheme differs from previous ones in that all users share the whole frequency band rather than a fraction for each user. A CDMA scheme, where the users employ orthogonal spreading sequences are termed as Orthogonal CDMA (OCDMA) and it shares the same properties of other schemes such as FDMA and TDMA in that the scheme allows interference free signal demodulation when the transmission channels are non-dispersive [2]. In OCDMA, the user specific sequences can be obtained using the rows of Hadamard Matrix [5]. Since the orthogonality can only be ensured when the signals of all users are perfectly synchronous, the OCDMA is usually employed in the downlink (forward link) of CDMA cellular wireless systems [25]. The user capacity of OCDMA is as noted is $C_{user} = 1$. The interference free performance in underloaded conditions of the OCDMA also serves as the basis for more spectrally efficient schemes which allows system to be overloaded i.e. $C_{user} > 1$. The so-called OCDMA/OCDMA schemes [25, 26], achieve the user capacity $C_{user} > 1$ by employing two sets of orthogonal sequences followed by iterative multistage detection.

2.4.1.5 Space Division Multiple Access (SDMA)

The aforementioned orthogonal multiple access techniques assume that signals from all users are received from the same spatial direction. Hence regardless of where the signals are transmitted from, the receiver can separate users' data based on the unique structure of their transmitted waveforms. It is now well recognized that the capacity of multiple access can be significantly increased if the additional spatial dimension (which is available for free) is also used see [31, 32, 33, 34, 35]. These different techniques have somewhat different purposes to exploit the spatial dimension like increasing diversity, spectral efficiency or user capacity. The schemes in [31, 33] are directly relevant to uplink multiple access channels and the method for achieving this is often termed as Space Division Multiple Access. The idea behind the SDMA is to exploit the space to increase the capacity with the use of a receiver with uniform linear antenna arrays consisting of several antenna elements. The elements are placed in a given space such as to ensure good array response towards desired direction. The composite response of the whole array is controlled by weighting each element by some weights. This approach also called beamforming is used to steer the higher gain beams towards desired users and low gain beams or nulls towards interfering users, see Figure 2.4. With this arrangement of K antenna arrays and signal processing tools, the SDMA may support up to $K-1$ number of users without subdivision in time, frequency or orthogonal codes [31, 33]. As can be noted here, the user capacity of the SDMA is usually $C_{user} = K - 1 >> 1$.

2.4.2 Non-orthogonal Waveform Multiple Access Schemes

The uplink of cellular wireless communications often characterizes with asynchronous data transmissions, nearfar user signal conditions. These conditions imply that the use of orthogonal sequences no more ensures that signal demodulation is interference free. Since synchronization of

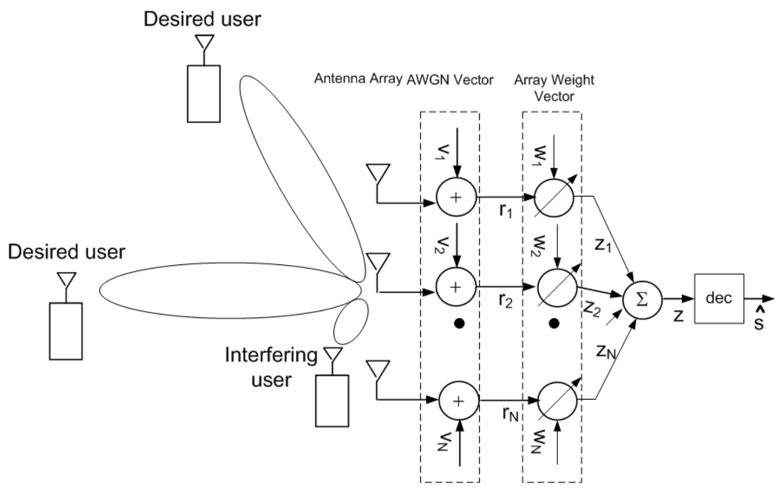


Figure 2.4: Space division multiple access system using antenna arrays and adaptive beamforming

users becomes more challenging in uplink, schemes such as CDMA with non-orthogonal spreading sequences offers attractive advantage for such environments.

2.4.2.1 Non-Orthogonal CDMA

The Non-orthogonal CDMA schemes benefit from interference averaging property from the wide-band nature of users' signals [2]. The nonzero cross correlation of users' signals introduce MAI in the system. It is a common practice to model the MAI as Gaussian noise for simplifying the performance analysis [36]. The user capacity of this schemes is limited by the total amount of MAI. If simple correlator receivers that ignore MAI are employed, the user capacity can be very low $C_{user} << 1$. Despite low user capacity in single cell settings, the universal frequency reuse factor of CDMA, offers much improved overall capacity in cellular settings [37]. The performance of Non-orthogonal CDMA schemes depend upon types of spreading sequences used [5], [38]. The simplest form of sequences are random sequences. Practical systems employ pseudo-noise (PN) sequences that exhibit properties much like random sequences and also offer near ideal autocorrelation property that can be exploited of synchronization purposes. On the other hand, sequences with good cross correlation properties are desirable for increasing the user capacity of CDMA. Gold and Kasami sequences are the examples of such sequences[38].

In CDMA, to achieve higher value of $C_{user} > 1$, signature sequences for users has to be carefully designed to create least interference to other users in the system such as Welch Bound Equality (WBE) Sequences [39]. Due to higher complexity and other practical limitations of WBE sequences as noted in [40], suboptimal sequences are used in practice. Under these circumstances interference may not be completely avoided, therefore multiuser interference canceling receivers is usually required to improve the detection performance. The problem of overloading is tackled from a different angle with Collaborative Coding Multiple Access, which is described next.

	C_1+C_2	(C_1)	
		(0 0)	(1 1)
	(0 0)	0 0	1 1
(C_2)	(0 1)	0 1	1 2
	(1 0)	1 0	2 1

Figure 2.5: A two user uniquely decodable collaborative codes with $R_{sum} = 1.292$ bits per channel use [12]

2.4.2.2 Collaborative Coding Multiple Access (CCMA)

Different from aforementioned schemes, CCMA [14],[15] achieves multiple access function without subdivision in time frequency or orthogonal codes. This scheme exploits the unique decodability property of composite received signal formed by superposition of collaborative coded transmitted signals (codewords) assigned to users. The sum rate of this scheme is found to be much higher than unity. The data symbols of users are encoded using a set of codewords from a codebook. For example, codewords of length $n = 2$ [12] for encoding binary data signals of two user CCMA is shown in Figure 2.5. Here user 1 is assigned two codewords $C_1 = \{00, 11\}, N_1 = 2$ and user 2 is assigned three codewords $C_2 = \{00, 01, 10\}, N_2 = 3$ capable of encoding binary or ternary data symbols. It can be seen that the sum rate of this scheme is $R_{sum} = \sum_{l=1}^{T} \log_2\{N_l\}/n = 1.29$ bits per channel use; therefore for the same signal bandwidth and transmit power, the CCMA offers much higher sum rate than the OWMA schemes. The user capacity of this scheme is defined by number of users T whose composite codeword signals are uniquely decodable. Therefore for CCMA, $C_{user} = T$. The higher user capacity of CCMA has motivated the design of several new overloaded multiple access schemes for CDMA, that will be described in fuller details in Chapter 5. In the next subsection, some preliminaries and recent work on different schemes to support overloaded MAC without the use of antenna arrays is presented.

2.4.3 Overloaded Multiple Access Channels

The causes for the user capacity limitation of OWMA and Non-orthogonal CDMA have been identified in the previous subsection. All these schemes are focused towards ensuring maximum orthogonality between users signals during demodulation process. If the information theoretic limits of MAC are closely looked at, it can be realized that the orthogonal schemes are clearly suboptimal compared with more efficient schemes that pools the available resources among the users. It is shown in [2] that superposition coding of users signals followed by successive interference cancellation at the central receiver achieves all the rate corners of the capacity regions of the Gaussian multiple access channels.

The recent work on this interesting topic of communication through an overloaded multiple access channels are now briefly reviewed. While the term MAC has been used, the downlink multiuser transmission channels or equivalently Broadcast Channels (BC) [30] are also the part of our study in this thesis. It will be interesting to mention about a very important idea that the MAC and BC are actually related intimately and information theory has been used to support the duality concept of MAC and BC and their capacity regions [22]. Since the objective of this thesis is to propose and analyse practical schemes for supporting an overloaded MAC and BC with low error rate performance, no detailed study is made on information theoretic aspects of the schemes. The new schemes are influenced by the principles of collaborative coding and decoding [12, 41, 42],[14] and in particular applied for downlink and uplink of an overloaded CDMA.

2.4.3.1 Review of Channel Overloading Schemes in CDMA

The most comprehensive research on overloaded CDMA is conducted by Vanhaverbeke in his doctoral dissertation [43]. Where it is shown that user capacity of CDMA can be increased substantially by effective system design both at transmitters and receivers. More specifically, he investigates the use of scheme using two sets of orthogonal sequences, also called OCDMA/OCDMA . Popular examples of such schemes include the well-known Quasi orthogonal CDMA [44], Random OCDMA/OCDMA [26], Improved OCDMA/OCDMA [40]. The common idea behind all these schemes is that, the first set of complete N orthogonal sequences are assigned to first N out of K users. The remaining $U = K - N$ users, where $U << N$, are assigned the same set of orthogonal sequences with set specific mask and overlaid on top of the N sequences.

The premise behind their detection schemes is that the received signals of the users within set with complete sequences (set1) are easier to demodulate since interference received from U (set2) users are low enough to make reliable estimates of set1 users [43]. To detect users' signals within the set2, the estimates of set1 users' data can be used. An iterative multistage receiver with decision feedback detection is employed to successively refine the estimates of users signals for both sets. The scheme has been shown to perform very well, giving near single user performance with just three iterations [26]. Similarly, a hybrid multiple access scheme consisting of OCDMA and TDMA has also been proposed in [25] to increase the user capacity of CDMA. As it can be seen that the schemes described are interference limited and therefore they rely on complex interference canceling receivers to overcome the effects of MAI arising due to the loss of orthogonality in overloaded channels.

The OCDMA/OCDMA [26] and hybrid OCDMA/TDMA [25] schemes are based on the idea of exploiting the state of reduced interference to group of N users in the larger set $\{1, 2, .., N\} < K$ of the system. Since the signals of U users within smaller set $\{N+1, N+2, .., K\}$ are also spread over N degree of freedom, the effective interference of these users on the first set of users will only be $\sigma^2 = U/N$. Therefore, as long as $U << N$, the users within the first set will have good SINR and therefore demodulation and estimation of their signals will be very reliable. The obtained estimates can now be incorporated within the next stage of detection to demodulate and estimate users within the second set. One very important point to note is that this process of iterative multistage detection technique comes with higher computational complexity and thus can not be employed in the downlink for mobile users' receivers. Therefore, it is concluded that while the performance of such schemes are attractive, they can only be applied at significant increase in the complexity. Now the same problem is viewed from a different angle, and the notion of ambiguity and unique decodability is used to address the problem of overloaded channels and to avoid the excessive interference under such channel conditions.

The interference in the above techniques arises from the fact that when the system is overloaded, the orthogonality is not maintained. In such condition the introduction of other users' randomly distributed signals on the desired signals may become so severe that the detector may end up with wrong decision of user signal. Lets's consider a simple example of two users transmitting simultaneously their BPSK modulated signals over the common channel to achieve sum rate $R_{sum} = 2$. In absence of noise the output of the combined signals can take three values only. Since there are four possibilities of users' transmitted signals and the two of them give exactly the same output i.e. addition of 0 and 1 and 1 and 0 both give output of 1. Therefore there is a detection ambiguity when users transmit these bits. Other two possibilities give values of 2 and 0, respectively and can be decoded correctly. The conventional approach to overloaded CDMA tries to avoid the situation of detection ambiguity by employing interference estimate generation and cancellation. This ensures that the remaining signal will most probably contain either 0 or 1 signal values with high probability so that decision error is minimized. A very simple way to resolve the ambiguity problem is to employ collaborative coding or uniquely decodable codes [14], [45] and described earlier in subsection in 2.4.2.2. Also, the multiple access scheme based on such T-user collaborative coding, CCMA has shown to achieve sum rate much higher compared to that of OWMA with $C_{sum} = 1$[14], [46]. This idea of resolving the ambiguity can be employed within synchronous CDMA to increase the user capacity. As can be noted, the number of users can be T times the available sequences.

The techniques of multiuser detection and interference cancellation though more complex than conventional matched filter receivers, have received much attention in multiuser CDMA research community due to their potential to achieve high capacity [47]. This thesis makes a number of new contributions in this subject area. For example, new techniques are proposed that overcome many practical issues limiting the performance of the interference cancellation subclass of multiuser receivers. The detailed descriptions of techniques and their performances and complexities are presented in Chapter 3. However, to provide a an overview of basic principle of different multiuser receiver techniques and a literature review of this subject area, in the next subsection, simplified descriptions of a broad class of multiuser detection techniques and their limitations is provided. Main motivations and ideas behind the new techniques proposed in this thesis are also highlighted.

2.5 Review of Multiuser Detection Schemes for CDMA

2.5.1 Principles and Techniques

CDMA is a multiple access scheme that benefits from randomness of users' spread spectrum signals. This is well suited for an uplink of cellular system where users transmit their signals asynchronously. The general idea of CDMA detection is to use good autocorrelation properties of PN sequence of desired user to acquire initial synchronization and exploit the interference averaging property to demodulate its signal [2]. Throughout the thesis, a synchronous CDMA system model is assumed (see Figure 2.6 for example). The received signal can also be shown in matrix form by

$$\mathbf{r} = \mathbf{SAb+n} \tag{2.3}$$

where, $\mathbf{r} = [r_1, r_2, ..r_N]^T$, is the received signal vector, $\mathbf{S}$ is matrix of users' spreading sequences of size $K \times N$, $\mathbf{A}$ is $K \times K$ diagonal matrix of users' amplitudes, $\mathbf{b} = [b_1, b_2, .., b_K]^T$ is a vector of users' transmitted data and $\mathbf{n}$ is $1 \times N$ noise vector. The objective of CDMA detection is to obtain the best estimate of the transmitted data vector $\mathbf{b}$ from the given received signal vector $\mathbf{r}$.

2.5.1.1 Conventional CDMA Receiver (Matched Filter)

The simplest way of obtaining the data estimate is to correlate the received signal with locally generated users' spreading sequences and sum the output over a symbol period. A block diagram of such receivers is shown in Figure 2.6. This process also called despreading, inverts the spreading operations performed at the transmitters and provides soft estimates of users' data vector $\hat{\mathbf{b}} = [\hat{b}_1, \hat{b}_2, .., \hat{b}_K]^T$. For BPSK modulated CDMA user, this can be written as:

$$\hat{\mathbf{b}} = sgn\left[\Re\left\{\mathbf{RAb} + \mathbf{n}\right\}\right] \tag{2.4}$$

where, sgn is the sign operator, $\Re\{.\}$ is real part of a complex number, $\mathbf{R} = \mathbf{SS}^T \neq \mathbf{I}$ is the cross correlation matrix. As can be noted here, the despreading process obtains data estimates while

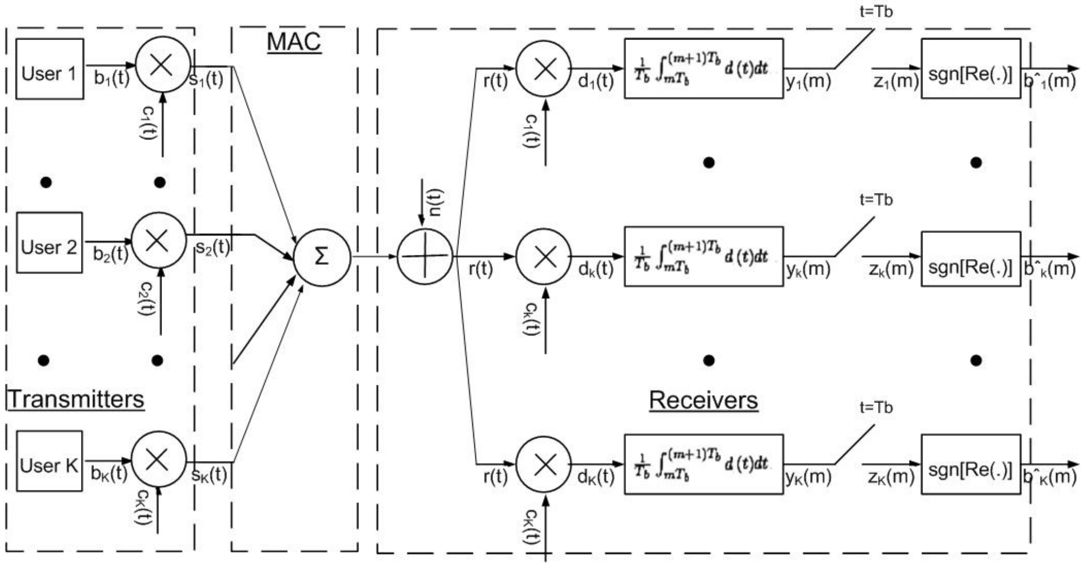


Figure 2.6: Uplink CDMA with conventional MF receivers

the amount of cross correlation among different users' signals or MAI still remains. The power of MAI becomes stronger with increase in number of users in the system. The amount of MAI is also dependent on type of sequences used and also the timing offset between users received signals [5]. For example, with spreading factor of N, the variance of MAI, σ^2, from each user for synchronous and asynchronous CDMA using random sequences with assumption of Gaussian distribution are well approximated to be $\sigma^2_{sync} = 1/N$ and $\sigma^2_{async} = 2/3N$ [36], [5]. For a K user system, the amount of MAI and receiver noise calculated as $\frac{K-1}{N} + N_0$, collectively defines attainable SINR. As it can be observed, the error performance of matched filter receivers degrade dramatically with increase in number of users. Therefore, to increase the spectral efficiency of the system, the study of more advanced multiuser detection techniques becomes inevitable.

Since the matched filter receivers ignore the presence of other users signals, the use of more advanced receivers that suppress MAI before detection known as multiuser detectors (MUD) have been proposed [28]. The MUD is probably one of the most intensely investigated topics in wireless communications in the past two decades . The essence of these techniques is highlighted in the sequel, followed by a closer look at the interference cancellation (IC) subclass of MUD. The purpose is to serve as an introduction of this technique to stimulate the discussion of our novel IC receivers that will be described in the Chapter three.

2.5.1.2 Optimum Multiuser Detection

The optimum MUD receiver proposed by Verdu [48] maximizes the likelihood of received vector $\mathbf{r}$ to a bit vector $\mathbf{b}_{opt}$ by searching over all possible vectors that satisfy following criterion:

$$\mathbf{b}_{opt} = \arg\max_{\mathbf{b}\in\mathbf{B}} 2\mathbf{b}^T\mathbf{A}\mathbf{r} - \mathbf{b}^T\mathbf{A}\mathbf{R}\mathbf{A}\mathbf{b} \tag{2.5}$$

where, $\mathbf{B}$ is a set of all possible bit vector of 2^K possibilities. As it can be seen that this method though optimum, requires computational complexity that of order exponential to number of users K. Although capacity obtained from the optimum MUD is substantial, its prohibitive computational cost has been the major issue for decades long intensive research on suboptimal, but low complexity MUD schemes and also for the techniques proposed in this thesis. Next, linear filtering based MUD schemes are reviewed briefly.

2.5.1.3 Linear Filtering Based Multiuser Detection

The linear filtering based MUD [29, 47] are designed on the principle of orthogonal projection of interfering signals from the received signal so that more reliable data demodulation can be made. These techniques attempt to replace the cross correlation matrix to an identity matrix, i.e. $\mathbf{R} \rightarrow \mathbf{I}_{K\times K}$. This results in interference free output signal, but with possibly some added noise due to the multiplication by $\mathbf{R}^{-1}$ both to signal and noise part as in (2.4). This is because the diagonal elements of $\mathbf{R}^{-1}$ have magnitudes higher than unity. The noise enhancement problem of decorrelation MUD [29] becomes particularly exacerbated as the system load increases. A more improved linear MUD is the MMSE detector that attempts to minimize the MSE between the data estimate and the received signal while suppressing the noise. This process requires the inversion of cross correlation matrix and also requires knowledge of receiver SNR, which can be shown as $[\mathbf{R} + \frac{N_0}{A^2}]^{-1}$.

The attractive properties of linear MUD are: at low system loading their error performance very near to that achieved by optimal MUD. Also, the schemes do not have problem of error propagation prevalent in interference cancellation type of receivers. The channel estimates obtained from these MUD are much more superior and, hence, they can achieve very low error performance. While these attractive properties make them very suitable for high performance multimedia communications [28], the biggest problem of these techniques is the complexity. The inversion of $K \times K$ matrix is quite expensive if not completely affordable for existing computing products. Therefore, more attention is now focused towards much lower complexity interference cancellation MUD receivers and are likely to be implemented at the earliest.

2.5.1.4 Interference Cancellation Based Multiuser Detection

The receiver technique based on interference cancellation is fundamentally different from the linear MUD techniques that have just been reviewed. These receivers make use of the fact that, given a rough estimates of users' data signals are available, these information along with amplitude estimates can be fed back and canceled from the received signal to further improve the detection performance of other users' data signals. An example of such scheme combined with linear MUD is investigated in [49]. There are two distinct approaches for interference cancellation MUDs. These are SIC and PIC. A very useful and comprehensive survey on recent development of IC receiver techniques is given by Andrews in [50].

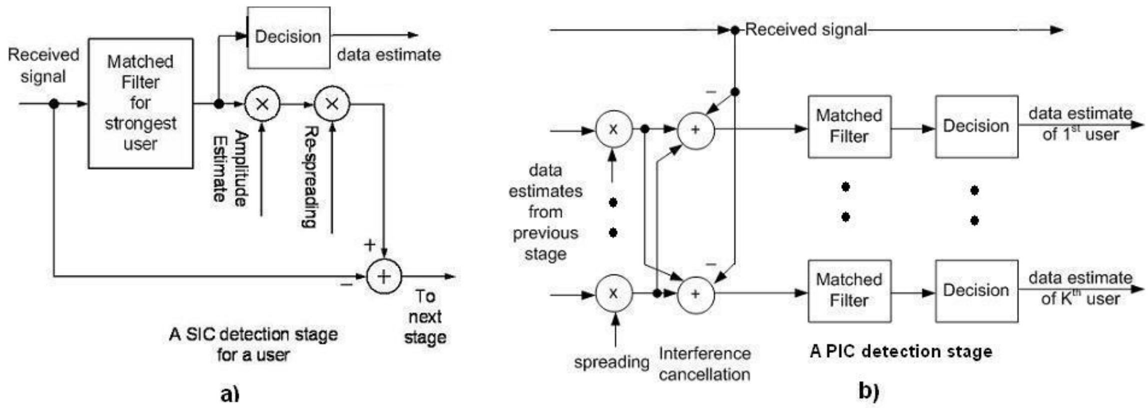


Figure 2.7: Successive and parallel interference cancelation receiver architectures

Before proceeding with the issues faced by these two receivers, the general principles of their operation are described next. As can be seen from Figure 2.7 (a), the SIC detects one user at a time, each time the estimated strongest user selected at each detection stage. This requires ordering of users' received powers, and are usually obtained from the bank of output of matched filter. The notion of SIC is that, the strongest is detected with much higher reliability and its cancellation from the received signal will remove significant amount of MAI from the system. This improves detection performance of later users to be detected. It has to be noted here that effective SIC operation requires each users' received signal to be of different power. This property of SIC is well suited in fading and near far user channel environments. An effective power control scheme is very important for a SIC [50] for maintaining equal BER performance under unequal user power conditions.

While SIC operates by detecting one user at a time, PIC, on the other hand, detects all users simultaneously, see Figure 2.7 (b). The initial stage of the PIC usually consists of a bank of matched filters matched to users' spreading sequences to provide a coarse estimate of users' data signals. These signals, along with the amplitude estimates of users, are used to generate MAI estimates for interference cancellation . The MAI signals are subtracted from the received signal to form cleaner input signal for improved detection of desired user's signals performance. The PIC receivers usually employ more than two stages of interference cancellation for satisfactory detection performance [51],[52]. It is widely believed that PIC performs far worse than SIC in fading and nearfar channels [50],[53], however, it will be shown in this thesis that an effective PIC can perform much better than SIC. Also, it will be shown that such PIC receivers are extremely nearfar resistant, which is contrary to common beliefs held by previous research [47],[50].

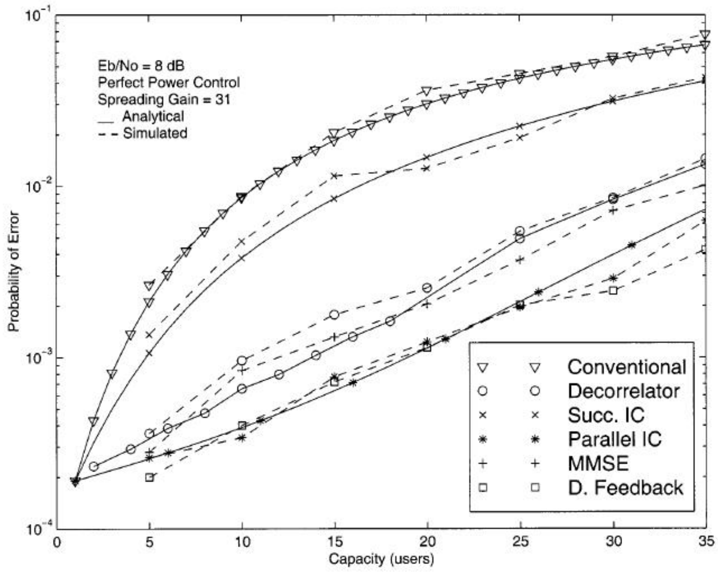


Figure 2.8: BER performance of different multiuser techniques using random sequences [47]

An excellent research work is conducted by Buehrer et. al. [47] to compare the performance of different MUD receivers under the common system settings. For example, Figure 2.8 is a very illustrative figure showing their relative performance in AWGN channel environment. It can be appreciated from the figure that MUD receivers offer substantial increase in the user capacity of CDMA.

2.5.1.5 Adaptive Filtering Based Detection

Another stream of researchers have focused their efforts on adaptive type of receivers based on MMSE or minimum output energy (MOE) criteria e.g. [54, 55, 56, 57, 58, 59]. The attractive features of these receivers are they do not require much information about other users' signals for performing interference suppression. These techniques suppress the interference using reference signals i.e. training data or fixed target value. The receivers are implemented as tapped delay lines and with adaptive weights for each tap that are updated to meet the predefined criteria. Among these schemes blind adaptive multiuser detection [57] proposed by Honig et. al. employs

MOE criterion to suppress MAI without any training sequence or knowledge of interfering users' signals. The scheme is shown to perform very well under static channels and with severe near far conditions. However it is also known to be very sensitive to mismatch in estimating the desired user's signature sequence. If the received sequence is different from the transmitted one, the receiver often suffers from cancellation of the desired signal and hence the performance degradation. Also the convergence to undesired local minima are well known problems of adaptive type of receivers.

We have now reviewed some interesting properties of different CDMA receivers and their individuals merits and shortcomings. We noticed that the multiuser receivers have potential to significantly improve the user capacity of CDMA. The improvement in performance of multiuser receivers do however come at significant cost in terms of computational complexity. Also favorable assumptions such as synchronous systems, ideal channel estimation do not accurately represent the scenarios that are encountered in practice. In the next subsections, some of the practical issues of MUD receivers are enumerated and contributions in this thesis to address them are described.

2.5.2 Practical Issues of Multiuser Detection

2.5.2.1 Computational Complexity

Much of the research on multiuser detection is motivated by the complexity issues of optimum receivers proposed by Verdu [48]. The linear MUD (decorrelator, MMSE) approach to CDMA offered performance comparable to optimum receivers (exponential in complexity) with linear increase in the complexity. The saving in complexity though sounds exciting, linear multiuser receivers are still far too complex than simple matched filter receivers. This is due to the requirement of the inversion of cross correlation matrix for the detection of each user at every symbol period which usually requires K^2 operations. The use of interference cancellation receivers does not require the matrix inversion operations and, hence, are attractive in terms of computational complexity. Below, a review of most critical problems faced by IC receivers and different approaches to address these issues as proposed by various researchers is provided. Finally, another view on the problems and novel approaches proposed by the author are also described.

2.5.2.2 Estimation and Cancellation of MAI

The IC receivers though are less complex than linear and optimum receivers, their performances are plagued by two practical phenomenons. They are the nearfar user conditions and imperfect channel parameter estimation; which often lead to the problem of serious error propagation and in some cases the performance can be even worse than MF receivers. The central cause for these problems is the very unreliable power estimation of conventional matched filter frontend. It is now well known that the MF introduce bias into the output decision statistic signals which increases linearly with increase in number of users [47]. Therefore, much of the research on IC receivers have assumed either equal power AWGN environments for PIC [51], [52] or perfect knowledge of received power of users signals in the case of SIC [47].

There exists different approaches in the literature [52], [53], [50] to tackle the problem of unreliable MAI estimate generation due to the use of conventional matched filter within IC receivers. A popular approach is the partial cancellation by Divsalar et.al. in [52] that shows much improved performance compared with conventional PIC receivers. The intuition behind partial cancellation is backed up by Buehrer et. al. in [60] where they also provide some theoretical analysis of the bias problem in high system loading. Since the partial cancellation approach is based on some trial and error search of cancellation factors, an adaptive weighting of interference estimates for cancellation using a least mean square (LMS) algorithm for multistage PIC is proposed by Xue et. al. in [61]. The scheme has shown to improve the performance of PIC considerably. Recently, another approach of PIC in Rayleigh fading channels for SINR maximization by exploiting the cross correlation of users' sequences is proposed by Tikiya et. al. in [62]. The scheme is shown to be much improved performance compared with conventional PIC at much higher computational complexity.

Similarly, the partial and limited number of cancellation approaches have been applied to SIC. Another approach of using multistage detection with soft decision at early stage is proposed by Hui and Letaief in [63]. This approach is later refined by Zha [64] which uses combination of hard and soft decision at each stage based on the magnitude of output signal and amplitude averaging [53] to improve the performance of a multistage SIC. Following another course of research on problem of SIC in practical environments using improved power control has been investigated thoroughly by Andrews et. al. in his doctoral dissertation [65]. His approach to solve the problem of imperfect estimation and cancellation is motivated by the fact that if one can learn the static of channel estimation error of the user's signal at the receiver, that information can be incorporated within the power control algorithm to overcome the capacity penalty due the imperfect cancellation. The optimum power vector is found to be achievable in different channel setting that equalizes the SINR for all users using an iterative algorithm [65]. A comprehensive survey of techniques addressing the issues of imperfect MAI estimation and cancellation can be exhaustive and, hence, only the most significant techniques have been highlighted in this discussion. Based on the brief survey on issues of aforementioned techniques, next a brief discussion of novel blind adaptive approach to address the problem of imperfect MAI estimation and cancellation is carried out.

Summing up from above, we conclude that IC receivers can be effective propositions to improve the user capacity of CDMA due to their lower complexity and potentials to overcome nearfar channel conditions. Many variants of IC receivers in the literature are mainly investigated assuming a more favorable assumptions such as perfect power control, ideal channel estimation, etc. Practical wireless systems are characterized by rapid changes in users' channel amplitudes and phases, also with non-ideal distribution of their power due to their varied locations. The main problem for IC receivers is therefore accuracy in the estimation and cancellation of MAI. To improve the detection and MAI estimation, a novel blind adaptive approach [16, 17] is proposed in this thesis and applied for both SIC and PIC receivers. The technique uniquely exploits the constant modulus property [66, 58] of users' transmitted data to acquire their channel estimates during the despreading process and makes use of the despreader weights to perform the MAI cancellation. The Chapter three describes these new techniques in greater details and their performance analyses.

In this section, it has been appreciated that capacity and performance of non-orthogonal CDMA that is limited by MAI than that of any other effects and MUD can be used to achieve a part of the capacity. The information theoretic capacity of the CDMA, however, lies far above what has been achieved in practice. The traditional approach to multiple access which treats each user as interferer, often require much higher complexity receivers. Another approach of collaborative coding multiple access with error control capabilities is introduced in [14], where it is shown that with simple decoding technique, the scheme achieves much higher sum rate than traditional FDMA, TDMA and CDMA schemes. Also, it is presented in [45] that collaborative coding can be embedded within downlink of CDMA to support an overloaded system with much improved performance compared with other overloaded schemes [44], [26], using very simple and low complexity receivers. We have also noted that, the MUD is an essential technique to improve the performance of CDMA. To achieve high capacity in fading wireless environments, some form of diversity is also needed. Therefore to design an effective system, one has to consider all the aspects users' channels and enabling techniques such as MUD and diversity. In the sequel, a brief discussion is carried out on what can be done to further improve the performance of existing multiuser systems in practical fading and interference limited wireless environments.

2.5.3 Discussion on New Approaches to Improve the Performance of Multiple Access in Fading Channels

The uplink of CDMA comes with many challenges making it much more interesting than its downlink counterpart. Therefore the remaining part of the Chapter is focused on novel approaches for increasing the capacity in uplink of CDMA. While there is only one source (base station) transmitting data to users in the downlink, there are K *independent* sources (users) in the uplink.

Independent users with *independent* transmit channels means much more if the problems of high capacity multiple access system is closely studied. The conventional approach to CDMA systems has considered each user as interferer to another user. It will be shown in chapters 4 and 5 that, the users can do much more than interfere to each other; they can cooperate to transmit their independent and/or mutual information to gain the much needed diversity and subsequently increase user capacity in a practical wireless environments. The MUD or IC techniques play important roles to achieve this gain. New schemes combining user cooperation diversity and MUD is presented in Chapter 4. Another novel collaborative approach to transmit their data while sharing the common resource, i.e. spreading sequence are presented in Chapter 5. To appreciate the problems that have motivated the work in aforementioned chapters, in the next section, general overview of fading channels, diversity and collaborative approach to uplink CDMA, along with review of related work are presented.

2.6 Fading Channels

In this Section, important concepts of fading channels, their characteristics and effects on the error performance are reviewed. The independent fading of channels are also used for collaboration of multiple users that share the common system resources to achieve the higher user capacity in later part of the thesis. Wireless communications systems contrasts with their wired counterparts in two ways: firstly, the transmitted signals propagate through multiple paths and add constructively or destructively, giving rise to large fluctuation in dynamic range of the signals in relatively short period of time, also called fading [67, 5]. Secondly, the transmitted signals are of broadcast nature; therefore, there is always information traveling around whether they are intended for the given receiver or not [9]. These two properties of the transmission channels make wireless communications more challenging and interesting for system engineers. The wireless transmission under mobile user conditions characterizes with multipath propagation. The multipath signals have random distribution of their amplitude and phases, which add constructively or destructively over the channels due to mobility of transmitter and hence causes so called channel fading effect. The fading of wireless channels is long known as being the fundamental bottleneck to reliable communications over such channels. This is due to the dramatic fluctuation in power of the received signal in the range of several tens of dB (see Figure 2.9 for an example) and leads to higher probability of the signal being severely faded, making reliable demodulation very difficult in such events [2]. Following this brief review, some important properties of wireless fading channels and their effects are discussed next.

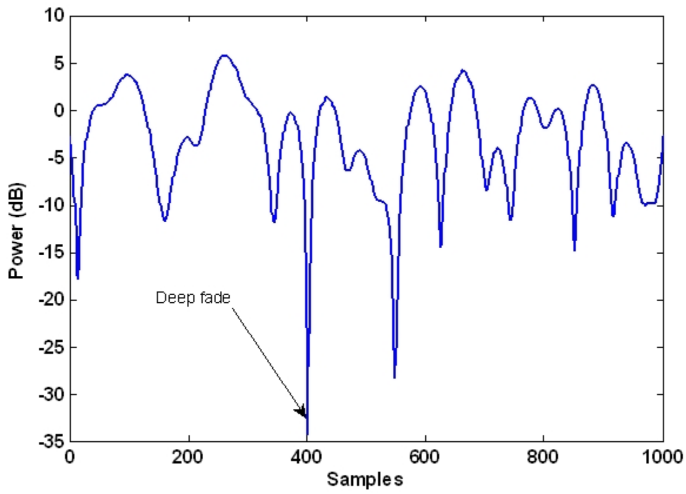


Figure 2.9: Typical envelope of received signal under Rayleigh fading channels

2.6.1 Doppler Spread and Coherence Time

An important measure to assess the channel fading is the coherence time, which is defined as the time period over which the magnitude of the channel changes significantly. This period is determined by the Doppler shift of individual path contributing to the channel, which for n^{th} path can be shown as:

$$D_n = f_c \tau_n(t); 1 \leq n \leq N \tag{2.6}$$

where f_c is the carrier frequency and τ is the path delay. The Doppler spectrum of the channel is the contribution from all N paths and also used to define the Doppler spread as follows

$$D_S = \max_{m,n} f_c |\tau_n(t) - \tau_m(t)|; 1 \leq m \neq n \leq N \tag{2.7}$$

where the maximum is usually taken over the significant paths only. The coherence time of the channel is the period over which its magnitude changes significantly and is inversely proportional to the Doppler spread.

2.6.2 Delay Spread and Coherence Bandwidth

The fading channels are also characterized by another important parameter known as Delay Spread of the multipath signals. This is defined as the time period between the shortest and longest path that significantly contribute to the received signal and can be shown as

$$T_S = \max_{m,n} |\tau_n(t) - \tau_m(t)|; 1 \leq m \neq n \leq N. \tag{2.8}$$

The delay spread of the channel directly effects the frequency coherence or response of the channel. The frequency response of the channel is also given by

$$H(f;t) = \sum_n g_n(t) \exp^{-j2\pi f \tau_n(t)} \tag{2.9}$$

where the effects of multipath is that the phase component is differential i.e. $2\pi f(\tau_n(t) - \tau_m(t))$. This causes the frequency selective fading of the received signal. The channel coherence bandwidth W_c is related to its delay spread as T_s as follows:

$$W_c = \frac{1}{2T_s}. \tag{2.10}$$

The coherence bandwidth also define the channel to be frequency flat or selecting. When the coherence bandwidth is much greater than that of the input signal, the channel is termed as *flat fading*, conversely when the coherence bandwidth is small than of the signal bandwidth it is termed as *frequency selective fading*.

2.6.3 Autocorrelation

Fading channels exhibit significant amount of memory and hence the period in which channel remains in the same state determine various coding design or diversity schemes to ensure more reliable communications under deep faded conditions. An important measure to assess the independence of fading across time/frequency and space is channel autocorrelation $R(t)$. This is calculated here using Jakes model of fading channels consisting of uniform scatters of given carrier frequency f_c and velocity of movement of object v and define the Doppler spectrum with fading rate f_d. Under this model, the normalized autocorrelation function $R(t)$ can be given by

$$R(t) = J_0\Big(2\pi f_d \tau\Big) \tag{2.11}$$

where, $J_0(.)$ is the zero order Bessel function of first kind [30], and τ is the observation period. Using (2.11), the minimum antenna separation distance (coherence distance) for ensuring independent fading is found to be 0.4λ. Hence, independent copies of signals can be obtained by receiving multiple data signal over different channel coherence time, bandwidth or spaces.

2.6.4 Statistics of Fading Channels

The signals received from the fading channels are often modeled by the distribution of their magnitudes for various performance analyses. The random amplitude and phase are often modeled to be complex Gaussian random variables using the well known Central Limit Theorem due to large number of scattered signal components. When the propagation paths from the transmitter to the receiver consists of non line of sight environment, the resulting signal is modeled as zero mean and unit variance complex random Gaussian variable $\mathcal{CN}(0, \sigma^2)$. The probability density function of the such signals follow the Rayleigh model given by:

$$g(t) = \frac{r}{\sigma^2 \exp\left\{\frac{-r^2}{2\sigma^2}\right\}}. \tag{2.12}$$

This Rayleigh model of channel fading is widely used as it very closely represents the practical fading phenomenon of cellular environments. This simple and useful model is also used throughout the thesis to evaluate the performance of various new techniques we propose. When there is some degree of line of light i.e. there is a specular component, another channel model is often used. This is obtained by modeling the specular component as the deterministic signal, which is combined with many weak diffused components. The channel $g(t)$ is modeled as

$$g(t) = \sqrt{\frac{\kappa}{\kappa+1}}\sigma_n e^{j\theta} + \sqrt{\frac{1}{\kappa+1}}\mathcal{CN}(0, \sigma_n^2) \tag{2.13}$$

where the first term is the specular signal and the second term with $\mathcal{CN}(0, \sigma_n^2)$ represents the sum diffused components as a Gaussian and κ also known as Rician factor of the channel is the ratio of power between the specular component and sum of the diffused signals.

2.6.5 Error Performance of Faded Signals

Strictly speaking, the fading can be completely mitigated if the channel state information is known at the transmitter and power control is used to invert the fading amplitude to maintain constant magnitude of the received signal. However, this approach is known to waste all the power to compensate for deep fade and thus incurs large capacity penalty [68]. A better approach to counter the detrimental effects of fading that does not incur capacity penalty is an area of intense research for so many years. In this line of ideas, it is known that the probability of deep fade can be reduced significantly if a measure is taken to ensure that more than one copy of the transmitted signals is available at the receiver. The schemes providing multiple copies of same information over the wireless channels are collectively known as *diversity* [2]. The diversity techniques rely on the fact that the signals propagated through multiple paths are uncorrelated and identically distributed . For example, with the use of two path diversity channels, even if a signal through one path is significantly faded, the other path is most probably in good state and, hence, much reliable demodulation of information is possible than non diversity case by combining them, see Figure 2.10. There are various approaches to provide diversity in fading wireless channels such as, time, frequency and space diversity. These methods are briefly reviewed next to motivate our discussion on a new form of spatial diversity in multiuser systems, namely the user cooperation diversity.

The use of diversity while improves the reliability of reception, can also offer increase in overall spectral efficiency [30]. The effect of diversity on system performance can be explained

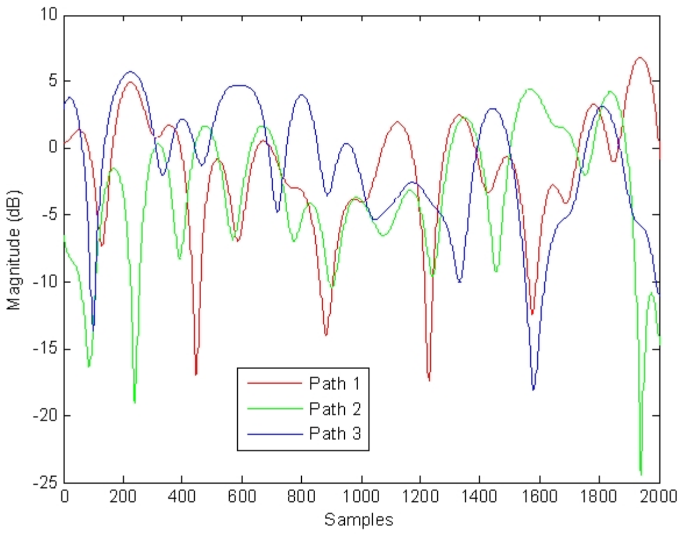


Figure 2.10: Multiple independent Rayleigh fading paths provide diversity

using a simple signal model as follows

$$r_m^l = g_m^l s_m + n_m^l; \qquad l = 1, .., L; m = 1, .., M \tag{2.14}$$

where, r_m^l is the m^{th} symbols of the received signal at l^{th} antenna, L is the number of receiver antennas, $g_m^l \sim \mathcal{CN}(0, 1)$ is the complex Rayleigh fading channel with zero mean and unit variance at l^{th} antenna, s_m is the transmitted symbol and n_m is the receiver noise with two sided power spectral density of $N_0/2$. When single antenna is employed $L = 1$, the probability of r_m being in deep fade can be approximated as the inverse of signal to noise ratio $Pr \approx 1/SNR$, with $SNR =| s |^2 /N_0$. Assuming that deep fade events translate to the most bit errors, the probability of decision error Pe, can be shown to be inversely proportional to the SNR,

$$Pe \approx 1/SNR. \tag{2.15}$$

Assuming fading over L paths are independent, the probability that all paths are under deep fade simultaneously becomes much smaller [30]. Therefore, by transmitting the same information over L independent channels, the probability of error Pe can be decreased significantly; which can also be shown as [2]

$$Pe = \left(\frac{1-\mu}{2}\right)^L \sum_{l=1}^{L} \begin{pmatrix} L-1+l \\ l \end{pmatrix} \left(\frac{1+\mu}{2}\right)^l \tag{2.16}$$

where

$$\mu = \sqrt{\frac{SNR}{1+SNR}}. \tag{2.17}$$

As noted from (2.16), the use of L-order diversity improves the probability of error performance in L^{th} power of SNR. The relationship between the BER and SNR under different diversity order

L is linear and is plotted in Figure 2.11 as an example. It can be clearly seen that by increasing the diversity, the BER improves significantly and approaches that of AWGN channel under $L \longrightarrow \infty$.

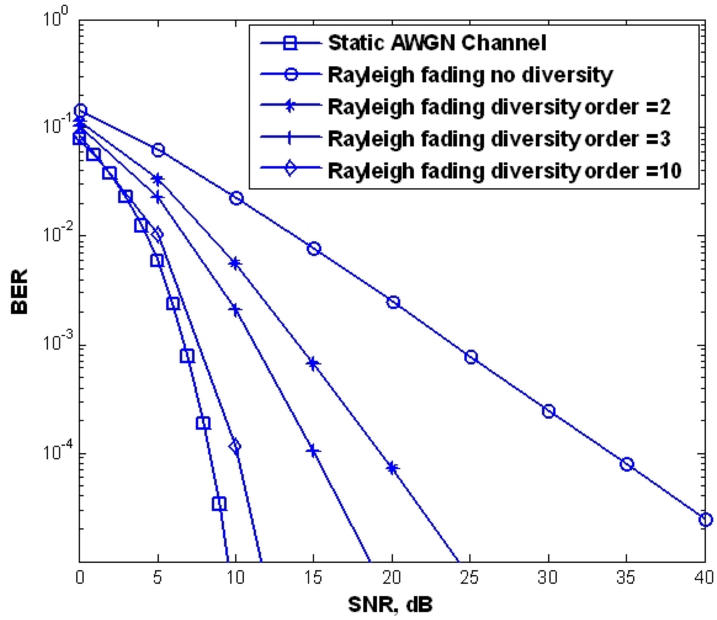


Figure 2.11: BER performance of diversity reception under Rayleigh fading channels

2.7 Diversity Schemes

The diversity is probably one of the most well-investigated topics in the wireless communications literature. Using the idea of independence of fading signals, different diversity approaches are briefly discussed next.

2.7.1 Time Diversity

The fast fading channels can also be regarded as ergodic process [2] and hence, there is a natural time diversity that can be exploited to improve the reliability of signal transmission. By simply transmitting multiple copies of the information over L symbol periods, L^{th} order diversity can be achieved. When the channel is slowly fading, to fully exploit the time diversity, the information should be transmitted over several channel coherence period or average fade durations. A well known approach to exploits time diversity in such channels is to perform coding and interleaving at the transmitter and perform reverse operation at the receiver. Provided that the interleaving depth is sufficiently long to cover L coherence period, full L-order diversity is achieved.

2.7.2 Frequency Diversity

High speed data transmission over the fading multipath channels are often characterized with multiple impulse response within the signal bandwidth. The frequency selective fading behavior of such channels can be exploited to provide frequency diversity. This is dependent upon the measure of correlatedness of channels over different frequency bands. For example, multiple tap equalizers using techniques such as zero-forcing, MMSE [2] can used to invert the effects of the channels to obtain the diversity. In wideband transmissions such as CDMA, multiple delayed copies of the arriving transmitted data can be resolved and coherently combined at the RAKE receiver to give frequency diversity [5]. The use of RAKE receiver allows to gain some frequency diversity by favorably exploiting the multipath signals for the downlink and uplink CDMA. However, as the number of multipath increases their effect on the basestation receiver that has to simultaneously detect large number of users degrades drastically with each additional multipath [65]. This is because each multipath separated by more than a single chip period potentially becomes an interferer if not resolved.

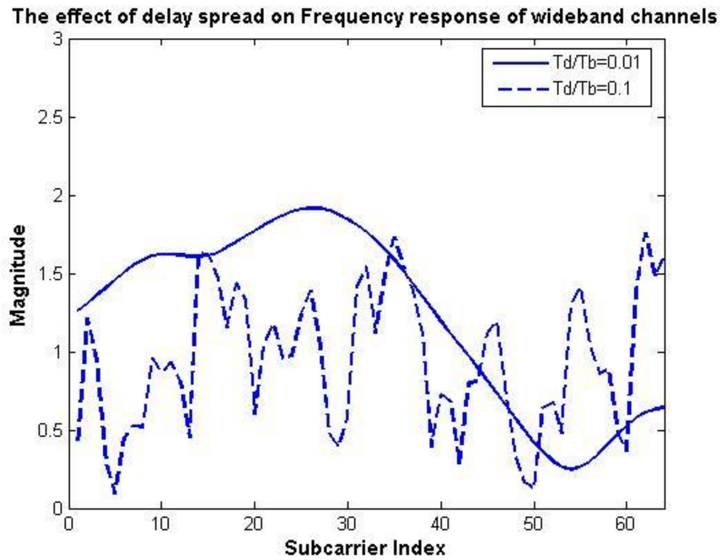


Figure 2.12: The effect of channel delay spread on wideband transmission signals: frequency domain view

The multicarrier or OFDM based transmission effectively solves this problem by reducing the chip rate and transmitting the slower rate information over many subcarriers simultaneously. As an example the frequency response of OFDM transmission using 64 subcarriers for different ratio of channel delay spread to OFDM symbol duration (T_d/T_b) is shown in Figure 2.12. The benefit of this approach is that multipath effects only a single chip signal rendering the fading to be flat on each subcarrier so that low complexity discrete Fourier transform can be applied to recover the signals from the subcarriers. The technique that combines CDMA spreading and OFDM technique are broadly called as Multicarrier CDMA (MC-CDMA) [69], [70], [71]. To exploit the frequency diversity of wideband CDMA with multicarrier transmission different subcarrier combining techniques such as MRC, EGC in the uplink and also MMSE Combining (MMSEC) in the downlink are investigated in the literature [69], [72]. The performance of uplink MC-CDMA is dependent on different system parameters such as: Receiver combining methods, type of sequences, subcarrier fading correlation, system loading, nearfar condition etc. To address the performance limitations of existing receiver approaches such as MRC, EGC, this thesis also proposes a novel blind adaptive subcarrier combining receiver technique in Section 3.4. The blind approach has shown to significantly improve the frequency diversity gains from MC-CDMA transmission and allows much higher number of users to share the same system bandwidth for the same target performance.

2.7.3 Space Diversity

It is shown earlier in (2.14) that, if transmission of information can be carried out over multiple sufficiently separated antenna channels, the information signals experience independent fading. With the use of channel knowledge, the signals can be coherently combined to significantly decrease the probability of deep fade of the resulting combined signal. The spatially separated diversity channels can be obtained either at the transmitter or at the receiver. The use of multiple receive antennas and combining their signals provides the diversity gain as well as so-called array gain as shown in Figure 2.13. Conversely, the transmitter diversity based scheme requires special coding techniques to avoid the received signals to interfere each other. With the use of space-time codes [73], it has been demonstrated that full L-order diversity can be achieved with L transmit antennas without extra time, frequency or power requirements.

Among the diversity schemes described above, the attractive property of the space diversity scheme is that, it offers the diversity gain at no cost of bandwidth expansion. The use of multiple antennas to improve the capacity of multiuser wireless communications has been pioneered by Winters [2] in 1980s. Since then, there has been tremendous research interests recently on exploiting the spatial dimension for both improving the diversity and transmission rates under the research theme of MIMO Spatial Multiplexing [32], [74]. Another stream of research works have been focused at the practical aspects of achieving the spatial diversity gain in multiuser settings by employing technique called 'User Cooperation Diversity', initially proposed by Sendonaris et.al. in [7]. The elements of this new technique and recent developments are briefly reviewed in the next subsection. Where, the motivations for our low complexity user cooperation techniques with successive interference cancellation for a practical uplink of CDMA is also discussed.

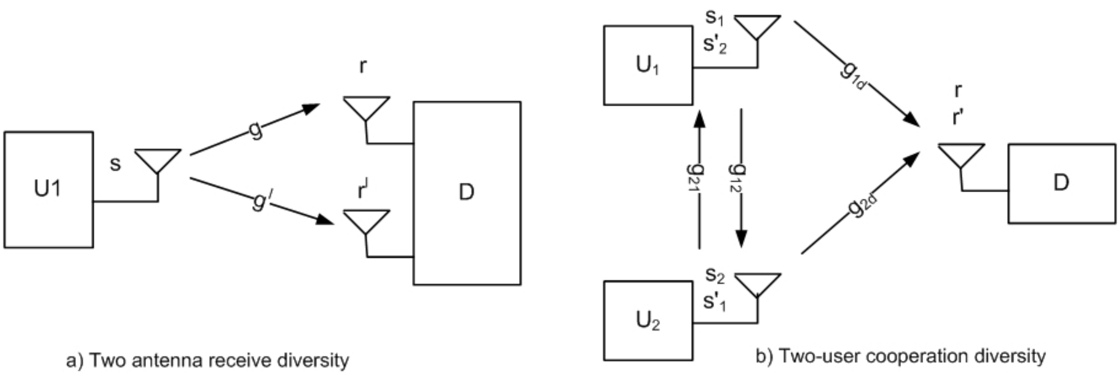


Figure 2.13: Different approaches for achieving spatial diversity

2.8 User Cooperation Diversity

It is well known that channel fading phenomenon brings many challenges for reliable tetherless communications due the significant probability of channels being in deep fade [2]. The information theoretic aspects of fading channels and various approaches for increasing the capacity have been thoroughly investigated in [67]. Despite the challenges, it has been reported in [2] that the information theoretic capacity of fading multiple antenna channels exceeds manyfold the capacity achievable in single antenna Gaussian channels. The asymptotical limit on achievable rates when multiple antennas are used both at transmitter and receivers also called MIMO channels are shown in [35].

Though the employment of antenna arrays is very effective in achieving diversity, mobile nodes however can not afford multiple antennas due to their size and power limitations. This condition also holds true for base-station receivers in many cases, as adequate antenna spacing to ensure uncorrelated fading may not be possible due to limited space available at the basestation. For wideband systems, such as CDMA, frequency diversity of multipath channels can be exploited, however its existence is purely based on the scattering effect of source signal over many paths and hence may not be practical in sparse environments with limited scattering [2]. Furthermore, slow fading environments, such as low mobility of users, may render the use of temporal diversity ineffective. Therefore, independent of any other diversity schemes, the use of some form of space diversity is always desirable for increasing the reliability of communications of fading channels.

Recently, the idea of cooperative diversity is becoming increasingly popular in both cellular and ad-hoc wireless networks due to its ability to provide significant diversity gain without having antenna arrays at the both end of the links [7, 75, 8, 76]. Although the technique is relatively new, the motivation behind it, traces back to early works on the use of inactive user nodes as relays to facilitate the reception of desired node's signals as investigated by Muelen in [77]. The capacity analysis of such networks is investigated by Cover [78]. Sendonaris, Erkip and Aazhang [7] developed the idea of relay channels further for multiuser CDMA cellular networks termed as, 'User Cooperation Diversity' and provided its capacity and outage probability analysis with practical implementations in cellular environments. Laneman investigated different variants of cooperative diversity protocols with space-time coding and time diversity in his dissertation [8]. Furthermore, the use of coding within cooperation is investigated by Hunter and Nosratinia in [75]. A more comprehensive review on different cooperative diversity schemes can be found in a paper by Aggelos et.al. in [79], where they also propose an interesting technique called 'Opportunistic Relaying' based on selecting best path from the group of relays. The diversity multiplexing aspects of the different cooperative schemes is also presented. The basic idea of cooperative diversity that has been receiving so much attention recently, is described next.

The conventional transmit/receive diversity schemes require L antennas to provide L-order diversity. The same L-order diversity can also be achieved if there are L user nodes cooperating among each other so that the information can be transmitted from L antennas of the user nodes. The diversity gain is however achieved at the expense of extra bandwidth as information of each user node has to be exchanged over all user nodes.

An example of a simple two antenna diversity reception and an equivalent two user cooperation diversity scheme that achieves the same diversity gain are shown in Figure 2.13. As it can be seen from the figure that, the cooperative scheme requires at least two symbol periods to complete one cycle of cooperation. The essence of the cooperative scheme using the system model of (2.14) with $L = 1$ can be listed as follows:

- Period 1: User 1(U_1) and User 2 (U_2), also called partners transmit their data $\{s_1, s_2\}$ using different time-slot, frequency or orthogonal sequence to the basestation receiver (D) via their independent channels $\{g_{1d}, g_{2d}\}$; the received signal at D this period is $r = s_1 g_{1d} + s_2 g_{2d}$. At the same time, the users process the received signals from each other via inter-user channels $\{g_{12}, g_{21}\}$to decode the transmitted information $\{s_1, s_2\}$ and obtain their estimates $\{s_1', s_2'\}$

- Period 2: The partners transmit the estimated data $\{s_2', s_1'\}$ via the partners' channels $\{g_{1d}, g_{2d}\}$ to the basestation receiver (D) to form $r' = s_2' g_{1d} + s_1' g_{2d}$. After receiving two signals r, r', the base-station performs maximum ratio combining of signals i.e. $rg_{1d}^* + r'g_{2d}^*$ and $rg_{2d}^* + r'g_{1d}^*$, respectively. If inter-user channel quality is good, symbols of users are decoded perfectly at the partners i.e. $\{s_1', s_2'\} = \{s_1, s_2\}$. The soft estimates of symbols are now $\tilde{s_1} = \left\{|g_{1d}|^2 + |g_{2d}|^2\right\} s_1$ and $\tilde{s_2} = \left\{|g_{2d}|^2 + |g_{1d}|^2\right\} s_2$ and hard decision on the resulting signals to obtain the final data estimates $\{\hat{s}_1, \hat{s}_2\}$

It is noted from the above protocol that, the cooperation achieves full second order diversity as the quality of inter-user channels improves. Note that the protocol assumed no interference between the transmitted signals from User 1 and User 2. A multiuser system such as CDMA uplink however suffers from many practical problems, and hence, simple protocol described above is clearly inadequate to fully exploit the space diversity from the user cooperation.

2.8.1 User Cooperation Diversity: Practical Issues

Bandwidth Efficiency Loss

The significant gain in BER of cooperative diversity scheme however comes at the cost of reduced bandwidth efficiency due to the need to decode and forward process required at the co-operating nodes that usually operate in half duplex mode. The seminal work by Sendonaris et. al. in [7] has mentioned this problem and assumed that it may be possible that the cooperating nodes could simultaneously transmit and receive (full duplex), given that transmit antenna gains are known and canceled from the received signal [78]. Despite the assumption of full duplex operation, the scheme in [7] incurs loss in bandwidth efficiency to achieves the diversity gain from node cooperation. This issue is addressed by Reberio et. al in [80] and Verdhe and Reynolds in [76] by using higher number of additional relays and use of superposition of user's signals.

A new bandwidth cooperative diversity is proposed in this thesis to address the issue of bandwidth efficiency in uplink fading multiple access channels. Like other bandwidth efficient schemes [80, 76], the proposed scheme [81] uses the same total system resources such as time slots, system bandwidth, transmit power, however the scheme makes use of the idea of collaborative transmission and reception [82]. Under this scheme, two or more users (collaborators) are grouped and they share the same spreading sequence for the transmission of their own data and that of the partners. The discussion on the idea of collaborative transmission and reception is carried in the subsection 5.2. The 'Collaborative Diversity' scheme itself is described in details in 4.5.

Multiuser Interference

The cooperative diversity for uplink CDMA proposed by Sendonaris et.al. in [7] assumed the use of orthogonal sequences and, hence, MAI is not an issue in the analysis of their schemes. The cooperation concept has also found widespread application in node to node ad-hoc communications, using relays where orthogonal channels are used, so that simple receivers can be employed [79, 83, 75, 8]. However, the existence of MAI due to non-orthogonal sequences of users in uplink CDMA is an important practical issue that can not be ignored. An investigation of cooperative diversity for a single user in uplink CDMA with multiple relay nodes (idle users) with the use of non-orthogonal spreading sequences at the nodes is carried out by [84]. Where it is concluded that cooperation alone does not ensure that full-order diversity can be achieved in such environments, and hence, more effective MAI suppressing receiver is incorporated in their scheme. The effects of MAI on the error performance and capacity in multiuser CDMA when users cooperate with their partners is an area that is investigated in this thesis. To put our work in context, in the sequel, some recent works carried out in this new area by different researchers will be reviewed and a brief overview and main findings of our schemes employing user cooperation with multiuser SIC receivers will be provided.

2.8.2 Combined User Cooperation and Multiuser Detection

The work of Fang and Hanzo [84] provides some insight into the problem of achieving cooperative diversity in CDMA under MAI effects. To provide an effective solution to the problem, joint study of different CDMA multiuser detection schemes and cooperative diversity is essential. However, there exists very little work addressing the problems of cooperation diversity in realistic multiuser environment with non-trivial MAI [20, 85, 18]. This is because reduction of MAI requires more advanced MUD receivers which complicates the idea of cooperative diversity. Since the complexity of MUD schemes play important roles in feasibility of their implementation in realistic wireless systems, the use of low complexity MUD techniques such as parallel/successive interference cancellation seems more attractive than linear MUD schemes [50].

Considering the issues of cooperative diversity discussed earlier, multiuser cooperative diversity schemes using two different successive interference cancellation techniques are proposed in Chapter 4. The SIC receivers are known as low complexity receivers with proven robustness in fading and nearfar user signal environments [65, 63, 53, 16]. First, the use of conventional SIC [53] is investigated in our cooperative diversity scheme [20]. The performance of this scheme is compared with a SIC without cooperation and that of cooperative diversity without interference cancellation. It has been found that under low user loading condition, the use of SIC with cooperative diversity achieves near full diversity order. However, as the number of users increases, the performance of the scheme shows notable degradation in BER. The main reason for such degraded performance can be explained from the problem of inaccurate MAI estimation and cancellation of convectional SIC as reported in [65]. This motivates us to incorporate the improved CMA-SIC receiver of [16] within the cooperative scheme to obtain better MAI estimates and interference cancellation. It will be shown that, the use of CMA-SIC can significantly improve the performance under high user loading conditions.

2.9 Chapter Summary

In this Chapter, overview of basic elements and important concepts relevant to multiple access wireless communications are presented. Needless to say that the multiple access is a very complex area of digital communications and comes with many practical challenges to achieve the capacity promised by the theory. An attempt is made to cover research efforts made concerning this important area. Following a brief review of information theoretic capacity of multiuser systems, another important performance metric called user capacity is described in Section 2.3. The user capacity of different multiple access schemes is reviewed in Section 2.4.

In Section 2.5 the problem of MAI in uplink of CDMA is investigated, starting with brief review of multiuser detection techniques and interference cancellation receivers in particular are highlighted in 2.5.1. Practical issues of MUD and imperfect channel estimation and cancellation of the IC receivers are discussed in 2.5.2 and different approaches to address these issues are reviewed. Adaptive receiver approaches are also reviewed and blind adaptive interference cancellation approach introduced by this research is briefly discussed.

Simple introduction to fading channels, their impact on system performance and different diversity schemes for mitigating the fading are given in 2.6 and 2.7, respectively. The basic concept of user cooperation diversity is described in 2.8 and its practical issues in wireless systems are discussed in 2.8.1. The emerging research efforts to provide cooperative diversity with multiuser detection is reviewed 2.8.2. The contributions made in the thesis to address the issues of multiuser interference, channel parameter estimation and nearfar conditions for cooperative diversity is also briefly highlighted. Chapter Three addresses some of these problems using a novel blind adaptive approach and will be described next.

Chapter 3

Blind Adaptive Multiuser Interference Cancellation Techniques for CDMA

3.1 Introduction

Conventional MF receivers treat the MAI as random noise, and hence are known to perform much worse when the number of users increases and their received power are unequal. The MAI problem can be alleviated with more advanced multiuser detection techniques [28]. The main objective of this Chapter is to propose simple and efficient techniques to maximize the number of users supported for a specified BER requirement, while keeping the computational complexity of the receiver as low as possible. Therefore among the MUD techniques, Interference Cancellation is the main subject of study in the remaining part of this Chapter. Two novel blind adaptive IC receivers known as constant modulus algorithm SIC (CMA-SIC) and blind adaptive PIC (BA-PIC) are presented in sections 3.2 and 3.3. The proposed receivers using blind adaptive approach to interference estimation and cancellation, can effectively solve the MAI problem and also improve the bandwidth efficiency of the system. Throughout the study, the simulation results are verified by theoretical analysis and vice versa. For this purpose, the approximately Gaussian nature of MAI signals are utilized, that allow simple and fairly accurate performance evaluation of the proposed techniques. The blind adaptive approach is then applied and evaluated for subcarrier combining in MC-CDMA receiver in Section 3.4. In Appendix A, a simple spreading scheme for BPSK DS-CDMA is described, which although is not directly relevant to the blind techniques presented in this Chapter, also addresses the MAI problem in a different manner. Following this brief introduction, in the sequel, the contributions are explained in greater details and main results are presented.

3.2 Improved Successive Interference Cancellation for DS-CDMA using Constant Modulus Algorithm (CMA-SIC)

As mentioned before, the performance of DS-CDMA system is limited by MAI, which is caused by multiple non-orthogonal users' codes sharing the same channel bandwidth. Multiuser receivers [28] that apply advanced signal processing at the outputs of the MF have shown to offer considerable capacity improvement over the single user MF receivers. Extensive research works have been carried out in the literature to improve the performance of the multiuser receivers using different techniques. A good performance comparison of various multiuser receiver techniques can be found in [47]. Among the multiuser receivers, SIC is a low complexity receiver that exhibits robust performance in near far scenarios. Conventional SIC uses MF outputs to sort the users' power and estimate their amplitudes. The estimates generated by the MF are impaired by MAI, leading to incorrect detection and cancellation of desired user signals causing the error propagation in the subsequent stages. There exist many different approaches for improving the performance of SIC in both AWGN and fading channels, e.g. [53], [64].

A simple SIC receiver using the correlation of received signal with users' spreading sequences is analyzed extensively by Patel and Holtzman in [53]. The amplitude estimates of users' signal are obtained directly from the MF outputs and sorting is performed at each user signal stage per symbol. To improve the cancellation process, average of the MF outputs over several symbol periods are also used. Zha and Blostein [64] proposed a SIC that combined hard and linear (soft) decision and cancellation technique employing the amplitude averaging proposed in [53]. It is shown that the performance is very near to the single user bound in AWGN channel with near far scenario.

Other techniques for amplitude estimation of the desired user signal include training sequences or blind methods. CMA is a popular, low complexity blind algorithm used for channel equalization and ISI suppression for constant modulus signals. The application of CMA in blind multiuser detection of CDMA signals has been proposed in [86, 58] and [87]. In [87], a receiver with parallel adaptive filter assigned for each user based on CMA algorithm followed by parallel cancelation units in fast multipath fading channel with power control is described. One limitation of this technique is that, in near far scenarios the CMA algorithm fails to lock on the desired user signal. CMA with cross correlation is presented in [58] to solve such problem in a static multipath channel. However, the technique requires the spreading factor to be at least three times the total number of users as the necessary condition to ensure global convergence of the CMA algorithm. Convergence to undesired local minima is the major problem for the application of CMA in CDMA multiuser receivers.

In this Section, a simple well-known gradient descent CMA algorithm is used, but implemented within each stage of SIC despreader at chip level to address the above issues and provide more effective approach to the user's amplitude estimation in fading channels and error propagation resulting in much improved BER performance and system capacity.

The rest of the Section is organized as follows. In subsection 3.2.1, a generalized system model is presented. In subsection 3.2.2, the principles of the proposed CMA-SIC technique are formulated and the detection and cancellation algorithm are presented in subsection 3.2.3. The theoretical analysis of BER and spectral efficiency analysis of CMA-SIC using Gaussian Approximation are carried out in 3.2.4. In subsection 3.2.5, simulation performance results are generated and compared with the conventional SIC and other separate CMA receivers.

3.2.1 System Model

A synchronous DS-CDMA system of K users in flat fading channel and AWGN is considered. The received composite signal $r(t)$ can be written as:

$$r(t)=\sum_{k=1}^{K}\beta_k(t)s_k(t)+n(t), \tag{3.1}$$

where $s_k(t) = \sqrt{P_k}b_k(t)c_k(t)$ is the transmitted signal of k^{th} user, P_k is the signal power, $\beta_k(t) = \alpha_k(t)e^{-j\pi\phi_k(t)}$ is the complex fading consisting of amplitude $\alpha_k(t)$ and phase $\phi_k(t)$ components, $b_k(t) = \sum_{m=-\infty}^{\infty} b_k(m)p(t-mT_b)$ is the data signal, where $b_k(m)$ is a binary sequence taking values $[-1,+1]$ with equal probabilities, $p(t)$ is rectangular pulse with period T_b. The spreading sequence is denoted as $c_k(t) = \sum_{n=-\infty}^{\infty} c_k(n)p(t-nT_c)$ with antipodal chips $c_k(n)$ of rectangular pulse shaping function $p(t)$ with period T_c and with normalized power over a symbol period equal to unity $|c_k(t)|^2 = 1$. The spreading factor is $N = T_b/T_c$ and $n(t)$ is the AWGN with two sided power spectral density $N_0/2$. The principles and the algorithm of CMA-SIC is given next.

3.2.2 Principles of the Proposed CMA-SIC

In the conventional SIC [53], the detection of user signals are performed based on order of their strength. First the estimation of the desired user is carried out, followed by the cancellation of its MAI contribution from the remaining composite received signal. The relative power estimates of the users are generated at the output of the corresponding users' matched filters and the one with maximum is selected at a time given by

$$z_k(m) = \max\left\{\int_{(m-1)Tb}^{mTb} r(t)c_i(t)dt\right\}, \forall i. \tag{3.2}$$

The estimate of k^{th} user data is taken as

$$\hat{b}_k(m) = sgn\left[Re\Big\{z_k(m)\Big\}\right], \tag{3.3}$$

where sgn and Re denote signum and real function, respectively. The signal $z_k(m)$ is respread using its spreading sequence and subtracted from $r_k(t)$ to remove its MAI contribution as follows

$$r_{k+1}(t) = r_k(t) - z_k(m)c_k(m). \tag{3.4}$$

The processes (3.2), (3.3) and (3.4) are then carried out until all user data signals are detected. Although the conventional SIC improves the detection performance of multiuser CDMA compared to the conventional MF receivers, it suffers from serious error propagation problem. This is due to imperfect MAI estimates generation and cancellation of the conventional SIC, because the MF output introduces some bias which scales linearly with increase in the number of users [47]. Partial cancellation method of Divsalar [52] can be used effectively to reduce the bias and get some improved performance. However, the selection of cancellation weight has to be done very carefully and there is no established theory that suggests the relationship between the weights and the bias reduction. Therefore, another approach is proposed, which incorporates CMA algorithm to address the bias problems of conventional SIC. It provides adaptive and robust operation to the changes in fading channels, number of users and nearfar problems of CDMA system.

In view of the performance of the conventional SIC and the problem of inaccurate MAI estimation and cancellation, it is desirable to generate the weights that doe not allow a decision statistic z_k to revert its sign when the presence of MAI tends to do so. The CMA is a simple algorithm that tries to maintain constant modulus of the signals at the output and its complexity is only $O(N)$ computation per symbol per user, where N is the length of the weight vector. Provided that the CMA is fast enough to track the changes in MAI power variations and the corresponding weights are selected, the decision error due the MAI effects can be completely eliminated. Practical CMA algorithms however may not perform perfectly and there are bound to be some inevitable misconvergence problems. However, the useful properties of the CMA is exploited here and suitably implemented within SIC and shown to minimize the effect of this issue. As will be seen later, the SIC interference cancellation does indeed improve the performance (convergence) of CMA algorithm in adapting weights in different scenarios such as system loads and nearfar problems.

The proposed architecture block diagram is shown in Figure 3.1 for K stages. In this architecture, the effect of strong interferers are removed at each successive stage, which aids the detection and cancellation for weaker user. At every symbol period, $r(t)$ is sampled at the chip rate to form the vector $\mathbf{r}(m)$ of length N chips. Bank of matched filter output based power sorter is employed to order signals according to their strength. The strongest signal is then selected for the first stage

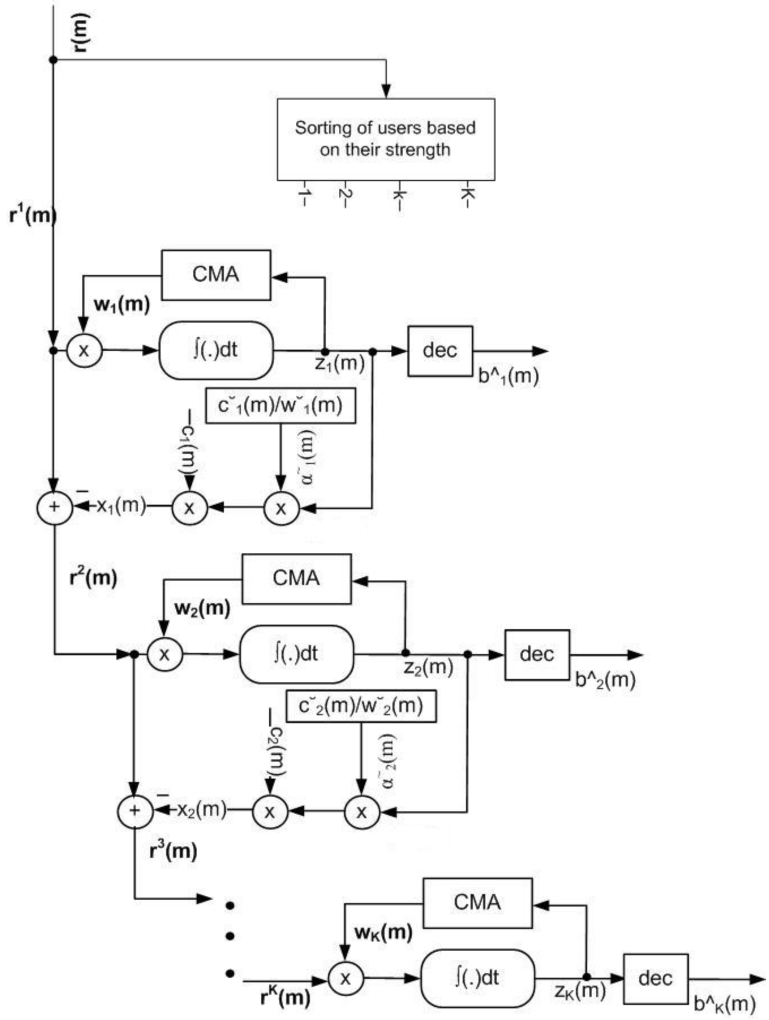


Figure 3.1: CMA-SIC architecture

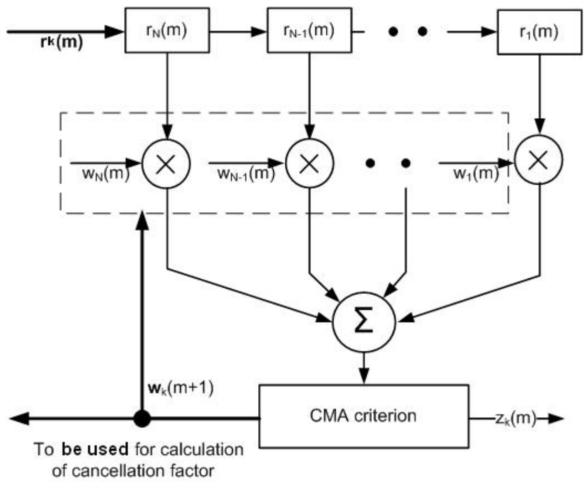


Figure 3.2: CMA-aided despreading process

for the estimation and detection of the desired signal. Adaptive CMA embedded within the despreader is used to adjust the desired incoming signal amplitude at the chip rate. For example, the despreading process for the received signal at the k^{th} user stage is shown in Figure 3.2. The output $z^k(m)$ of the first stage is then weighted utilizing the CMA weights $\mathbf{w}^k(m)$, spread and subtracted from the received signal to form the input to the next stage. This process is repeated for each signal to the weakest signal stage.

3.2.3 CMA-SIC Algorithm

At the first symbol period, the weights of the CMA are initialized with user's spreading sequence $\mathbf{w}_k(1) = \mathbf{c}_k(1)$. Without loss of generality, it is assumed that the first user (strongest among K users) to be detected is user 1. Similarly next strongest user is assigned an index as user 2 and so on. At the first stage, the received signal can be expressed as $\mathbf{r}(m) = \mathbf{r}^1(m)$. The remaining composite signal after cancellation at k stage is expressed as $\mathbf{r}^{k+1}(m)$.

At stage k^{th}, the decision statistic $z_k(m)$ is obtained by multiplying chips of $\mathbf{r}_k(m)$ with the vector of weights $\mathbf{w}_k(m)$ and summed over the symbol period given by

$$z_k(m) = \mathbf{w}_k^T(m)\mathbf{r}^k(m). \tag{3.5}$$

The CMA criterion J_{CM} can be written as minimization of the following cost function

$$J_{CM} = E(z_k(m)^2 - \gamma)^2, \tag{3.6}$$

where $E(.)$ is the expectation operator, γ is the dispersion constant, which is equal to unity for binary phase shift keying (BPSK) signals. The instantaneous error signal $e_k(m)$ is calculated as

$$e_k(m) = z_k(m)\left(z_k(m)^2 - \gamma\right). \tag{3.7}$$

The estimated gradient vector of the error signal is then calculated by

$$\nabla_k(m) = \mathbf{r}^k(m)e_k(m). \tag{3.8}$$

Using the gradient of (3.8), the weight vector at next symbol $\mathbf{w}_k(m+1)$ is updated as follows

$$\mathbf{w}_k(m+1) = \mathbf{w}_k(m) - \mu_k\nabla_k(m), \tag{3.9}$$

where μ_k is the step-size that is used to adapt the elements of the weight vector to minimize the cost function (3.6).

Finally, the output $z_k(m)$ is delivered to the decision making process to perform hard decision

$$\hat{b}_k(m) = dec\{z_k(m)\} \tag{3.10}$$

, where $dec\{.\}$ is taken as simple sign detector for BPSK signals.

The cancellation process also requires amplitude estimate of the detected user signal and spreading. The estimates are obtained using the weights of the CMA algorithm and the known spreading sequence as follows

$$\tilde{\alpha}_k(m) = \frac{\breve{c}_k(m)}{\breve{w}_k(m)}, \tag{3.11}$$

where $\breve{c}_k(m) = 1/N \sum |c_k(mN+i)|$ and $\breve{w}_k(m) = 1/N \sum |w_k(mN+i)|$, $i = 1, 2, ..N$, are the mean amplitude of user's spreading sequence chips and the mean of the weight vector updated by the CMA, respectively. The estimated symbol is then scaled with its new amplitude estimate $\tilde{\alpha}_k(m)$ and spread to generate the cancellation term as follows

$$\mathbf{x}_k(m) = \tilde{\alpha}_k(m) z_k(m) \mathbf{c}_k(m). \tag{3.12}$$

The remaining composite signal after the interference cancellation is given by

$$\mathbf{r}^{k+1}(m) = \mathbf{r}^k(m) - \mathbf{x}_k(m). \tag{3.13}$$

The processes (3.5)-(3.13) are repeated for each stage until the weakest user is detected.

It is important to note what contrasts the proposed SIC with conventional SICs. Here, the adaptive despreading using CMA serves dual purpose of interference suppression for detection and estimation of desired user data as well as amplitude estimation for more accurate cancellation of its MAI contribution to other users' signals compared with the conventional where no such mechanism is available.

In the table below, the step wise procedure of the CMA-SIC multiuser detector is summarized.

For $m = 1, 2, ..,$ perform steps $1 - 10$	
1	At $m = 1$, initialize $\mathbf{w}_k(1) = \mathbf{c}_k(1), \forall k$
2	Initialize CMA algorithm and step size μ_k
3	Despread $\mathbf{r}(m)$ for k^{th} user using spreading sequence/weight vector $\mathbf{w}_k(m)$ and store the sample of $z_k(m)$
4	Evaluate the CMA cost function and calculate gradient vector at the m^{th} symbol, $\nabla_k(m)$
5	Update weights for next symbol, $\mathbf{w}_k(m+1)$
6	Perform decision on $z_k(m)$ to generate $\hat{b}_k(m)$
7	Calculate the amplitude estimate $\tilde{\alpha}_k(m)$ and regenerate the cancellation term $\mathbf{x}_k(m)$
8	Cancel the regenerated signal $\mathbf{x}_k(m)$ from the total composite signal
9	Perform steps 2 to 8 for next strongest user signal
10	Stop after the detection of K user signals

Table 3.1. The CMA-SIC algorithm steps

3.2.4 System Analysis

3.2.4.1 BER

The proposed receiver algorithm keeps track of channel realization of each user at every symbol period during the detection and estimation processes. Therefore, the magnitude of channel autocorrelation over a certain time period is important for the receiver design. The Jake's model of fading channels is used and it consists of sum of uniform scatterers represented by individual sinusoids defined by the carrier frequency f_c, Doppler rate f_d. The channel autocorrelation $R(\tau)$ over a time period τ is obtained as follows:

$$R(\tau) = J_0\Big(2\pi f_d \tau\Big), \tag{3.14}$$

where $J_0(.)$ is the zero order Bessel function of first kind. The variance of output signal of k^{th} user within the proposed CMA-SIC is given by:

$$var\{z_k\} = \frac{K}{MR(t)} \sum_{m=1}^{M} \frac{\{e_k(m) \pm \mu\}^2}{N}; 1 \leq k \leq K. \tag{3.15}$$

The average SINR for k^{th} user's signal is obtained as follows

$$SINR_k = \frac{E^2\{z_k\}}{var\{z_k\}}. \tag{3.16}$$

Assuming equal average power of the users $P_k = P; \forall k$, the average SINR of the system is obtained

$$\overline{SINR} = \frac{1}{K} \sum_{k=1}^{K} SINR_k, \tag{3.17}$$

where $\overline{SINR}$ is the average SINR obtained from the channel realization $g_k(m), 1 \leq k \leq K$. The average probability of error is obtained by calculating the BER from averaged SINR from fading distribution of each user's signal and given by

$$Pe = \frac{1}{2}\left[1 - \sqrt{\frac{\overline{SINR}}{1+\overline{SINR}}}\right]. \tag{3.18}$$

3.2.4.2 Spectral Efficiency

This Section analyzes the spectral efficiency (SE) of CDMA with CMA-SIC in single isolated cell based on the previous studies on CDMA channel capacity. The SE of CDMA with conventional receiver using random spreading sequences has been shown by Verdu and Shamai in [6]. The employment of SIC further improves the SE and in ideal case, this has shown to approach the capacity of an Gaussian Multiple Access channel [65], [88], [23]. Practical systems come with inevitable channel (amplitude) estimation error problem and the achieved SE is far from ideal. In this Section, a simple analysis of achievable SE of practical SIC receivers (Conventional SIC and the proposed CMA-SIC) is provided and compared with that of SE of single user under Rayleigh fading with perfect channel state information (CSI). The capacity results of the receivers are taken over the expected SINR values from the results of simulations. Using the well known Jenssen's inequality theorem [30] that the averaged value of a random distribution does not exceed the expected value itself, the obtained capacity results represent as the upper bounds.

The SE C_0 of a single user transmission over the fading Gaussian channel with perfect CSI at receiver is simply the function of SNR, and is given by [6].

$$C_0 = E\Big[\log_2\{1 + \alpha^2 SNR_0\}\Big]. \tag{3.19}$$

In CDMA systems, the system bandwidth is equally divided among K users each occupying the fraction of $1/N$ bandwidth. Therefore, the normalized SE of a K user CDMA system is obtained as follows

$$C^{CDMA} \leq \frac{K}{N}\log_2\Big[1 + \overline{SINR}\Big], \tag{3.20}$$

where $\overline{SINR}$ is given is (3.17) above for a typical CDMA signals. The SE for CDMA system with conventional MF receivers employing random sequences under Rayleigh flat fading channels can be obtained using the SINR at the output of decision variables z_{conv} of MF receivers. The signal z_{conv} consists of contribution of desired user signal a_k and sum of correlated MAI users' signal components, given by

$$\begin{aligned} z_{conv} &= \int_0^{T_b} r(t)c_k(t)dt, \\ &= \sqrt{P}_k a_k b_k + \sum_{i=1, i\neq k}^{K} \sqrt{P}_i \rho_{ki} a_i b_i + n_k \\ &= D_k + I_k + n_k, \end{aligned} \tag{3.21}$$

where ρ_{ki} is the magnitude of cross-correlation between k^{th} and i^{th} users' spreading sequences, D_k and I_k are denoted as desired and sum of interfering users' data signals, respectively, n_k is correlated AWGN term. Therefore, the SE of conventional MF receivers are given by:

$$C^{MF} \leq \frac{K}{N}\log_2\Bigg[1 + E\{SINR_{MF}\}\Bigg] = \frac{K}{N}\log_2\Bigg[1 + \frac{E^2\{z_{conv}\}}{var\{z_{conv}\}}\Bigg], \tag{3.22}$$

where the MAI effects from $K-1$ users is assumed as Gaussian noise to obtain the SINR in (3.17). For systems employing SIC, the analysis becomes more complicated as ordering of users' signal power is required at every symbol period. For more detailed analysis of SIC receivers, an excellent paper by Patel and Holtzman [53] can be used. Below, the SE analysis of SIC and CMA-SIC is presented, based on the SINR obtained from our previous analysis of system SINR. The most important factor affecting the SE of SIC receivers is the accuracy of amplitude estimation. For the conventional SIC, the estimation of amplitude $\hat{\alpha}_k$ is simply taken as the output of matched filter $z_{conv}$for the user

$$\hat{\alpha}_k = |z_{conv}|; k = 1, .., K. \tag{3.23}$$

The SE of SIC can be obtained as follows

$$C^{SIC} \leq \frac{K}{N} \log_2 \left[1 + E\{\hat{\alpha}_k^2 SNR_0 \eta\} \right], \tag{3.24}$$

where expectation is taken over the fading power distribution of desired k^{th} user, $\eta = SNR_k/SNR_0$ is the asymptotic multiuser efficiency as defined in [28] to assess the multiuser receiver techniques. The SE of system employing CMA-SIC can be obtained as follows

$$C^{CMA-SIC} \leq \frac{K}{N} \log_2 \left[1 + \frac{E^2\{z_k\}}{var\{z_k\}} \right], \tag{3.25}$$

where the expectation is taken over the fading power distribution of the desired k^{th} user, $var\{z_k\}$ is the variance of decision variable of k^{th} user by employing the CMA-SIC. The reduced variance of decision variable for CMA-SIC is due to the improved interference estimation used for cancellation at each stage. Therefore, it is expected that, among the receivers analysed, CMA-SIC provides highest spectral efficiency.

3.2.5 Performance Results and Comparisons

A model of K user synchronous uplink DS-CDMA system employing BPSK and short binary Gold sequences [89] of length N = 31 is used. The initial code acquisition for such sequences are investigated in [90] and can be used here without much difficulties. The channel is Rayleigh flat fading channel with Doppler shift of 185Hz. A fixed step size of $\mu_k = 0.0001$, for all k is assumed in the CMA algorithm. The selection of step size in CDMA is generally based on the spreading factor used, the dynamic range of the received signal and effects the convergence of the algorithm [91], [59].

The BER performance of the proposed receiver (CMA-SIC) is shown in Figure 3.3 and compared with conventional receiver (Matched Filter), CMA receiver without interference cancellation (CMA only) [59], conventional SIC (SIC), and SIC with sorting each cancellation (SIC-Sorting)[53]. The proposed CMA-SIC showed a superior BER performance reaching the single user bound with 10 users. This also explains the accuracy of the proposed CMA-SIC algorithm in detection and MAI cancellation of the desired user signals.

The significant result of CMA-SIC compared to other receivers, can be explained as follows. As the signal-to-noise ratio increases, MAI becomes dominant source affecting the error performance. The error is introduced whenever the magnitude of the decision variable z^k becomes such that the polarity of transmitted user signals is reverted. Conventional SIC can not correct the magnitude of the signal z^k and, hence, suffers from error propagation. The CMA-SIC considers the decision variable of previous symbol period to adapt weights to make the variable closer to $+1, -1$. It is intuitive that with high probability the magnitude of the adapted output signal will be closer to the correct polarity of the transmitted data than that without adapting, as done in conventional despreading. The CMA only receiver has this capability, but it suffers from the problem of locking interferer than to the desired user signal due to dissimilar power profiles of the users.

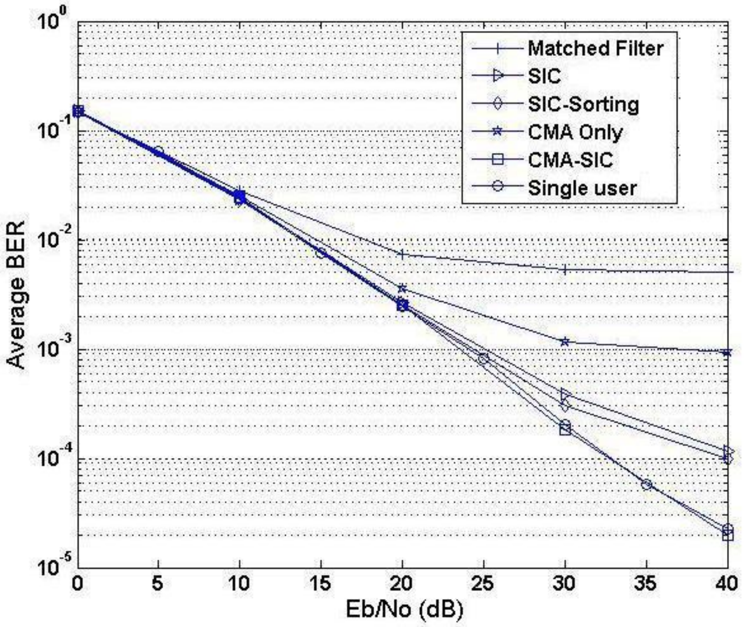


Figure 3.3: Performance of CMA-SIC in flat Rayleigh fading channel with K=10 and Gold sequences of N=31

The conventional SIC showed better performance than the Matched Filters and the CMA only receivers. The SIC-Sorting slightly outperformed SIC. It can also be seen that the CMA only receiver provides much improvement performance compared to Matched Filters. This clearly indicates the robustness of the CMA in suppressing the MAI even in fading channel.

The BER of CMA-SIC and conventional SIC are compared under higher loading of $K = 20$ users in Figure 3.4. The CMA-SIC retains near single user performance even with increase in number of users. The conventional SIC however is shown to perform far worse in this case and only slightly better than MF receivers. The reason for improved performance of CMA-SIC is due to both reliable interference estimation/cancellation and use of adaptive despreading. Figure 3.5 shows the performance of the proposed CMA-SIC in different system loading. The CMA-SIC does not exhibit error propagation when the number of users increase. It shows near single user performance even under heavily loaded system, while the error performances of SIC and SIC-

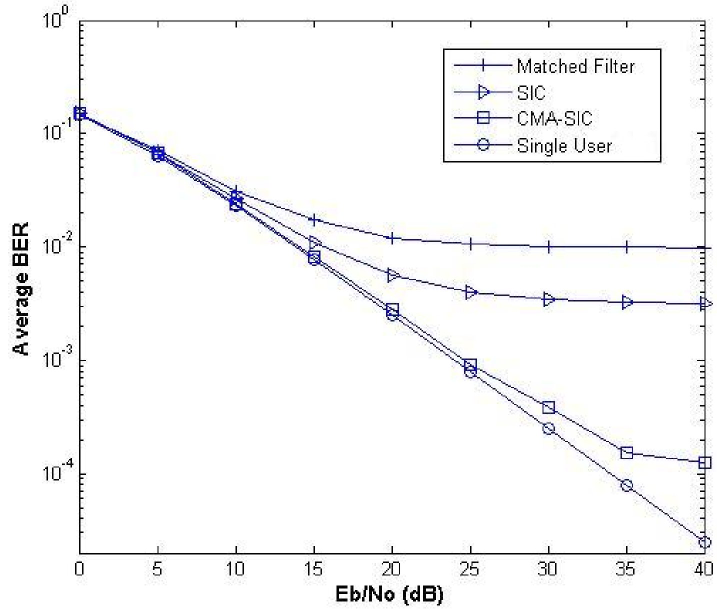


Figure 3.4: Performance of CMA-SIC in flat Rayleigh fading channel with K=20 and Gold sequences of N=31

Sorting and CMA only receivers start to degrade as the number of users increase above 15 users. SIC and SIC-Sorting showed good performance when system is lightly loaded, however, at high load their performance degraded approaching that of conventional Matched Filters. The CMA based receivers showed to offer comparatively better performance under such conditions.

In Figure 3.6, the output signal waveform of CMA-SIC and conventional SIC receiver are shown for $K = 20$ and under a) for $E_b/N_0 = 20$ and b) for 40 dB, respectively. For reference, the output of users' true fading channel output is also shown. It can be clearly seen from the figure that the CMA-SIC provides much accurate estimation of users' channel magnitude and hence leading to improved BER and spectral efficiency. The magnitude of channel estimation error is evaluated in terms of mean square error (MSE), $\varepsilon^2 = E\Big\{|\alpha - z\tilde{\alpha}|\Big\}^2$. The CMA-SIC shows much improved MSE performance compared with conventional SIC as the E_b/N_0 increases as can be seen in Figure 3.7.

Figure 3.8 shows the performance of CMA-SIC in fading channels and nearfar ratio of 10 dB. Here, the step size $\mu_k = 0.00001, \forall k$ is assumed in the CMA algorithm. The desired user (weakest user) has unity power, while all other users have been assigned powers uniformly distributed between 0 and 10dB. It can be clearly seen from the figure that CMA-SIC is more robust compared to all other receivers. Performance of SIC and SIC-Sorting receivers degraded dramatically as the system load increased above 15 users. CMA only receiver performed worse than in equal power.

The nearfar resistance of the CMA-SIC is also investigated for a highly loaded system with 20

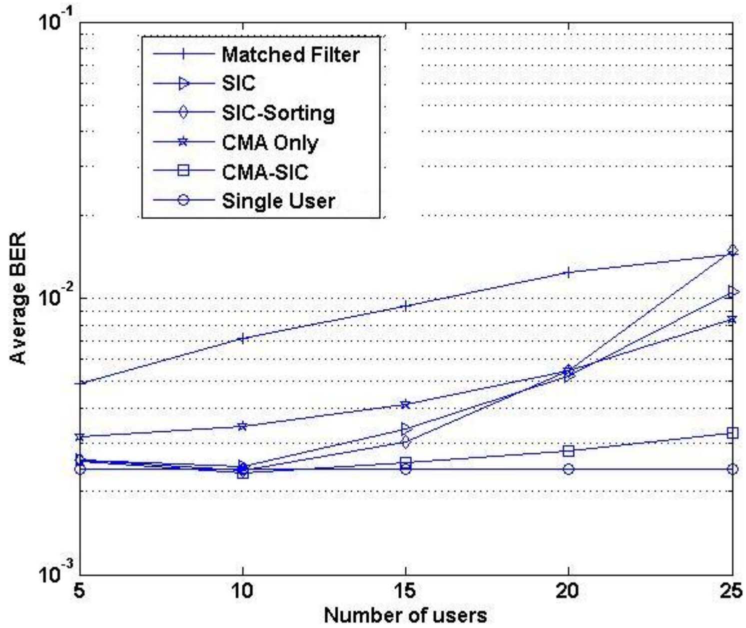


Figure 3.5: BER vs. number of users of CMA-SIC in flat Rayleigh fading channel with $E_b/N_0 = 20dB$ and Gold sequences of N=31

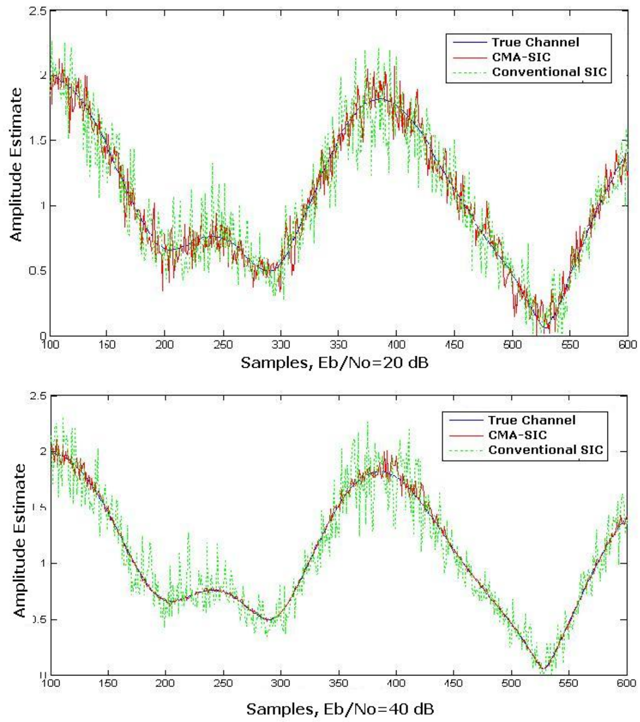


Figure 3.6: Amplitudes estimates at the output of CMA-SIC in flat Rayleigh fading channel with K=20 and Gold sequences of N=31

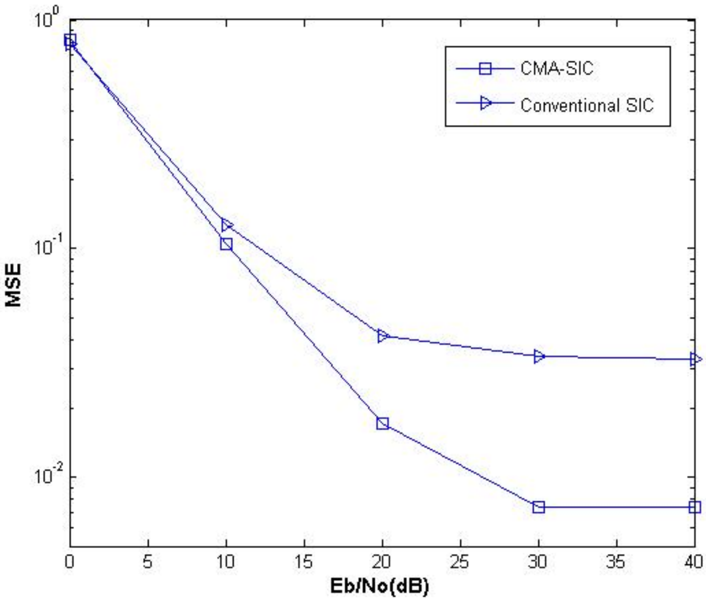


Figure 3.7: MSE performance of amplitude estimation of CMA-SIC in flat Rayleigh fading channel with K=20 and Gold sequences of N=31

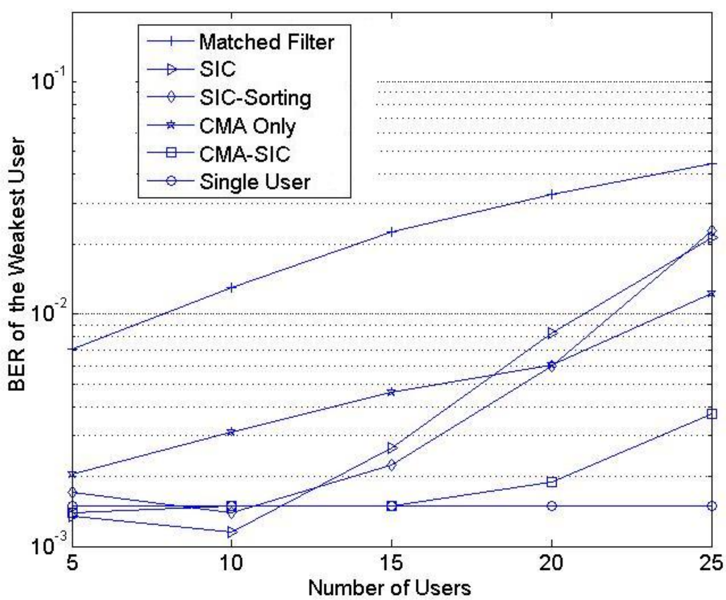


Figure 3.8: Performance of CMA-SIC in flat Rayleigh fading channel with nearfar ratio of 10 dB; E_b/N_0 of the weakest user=20dB and Gold sequences of N=31

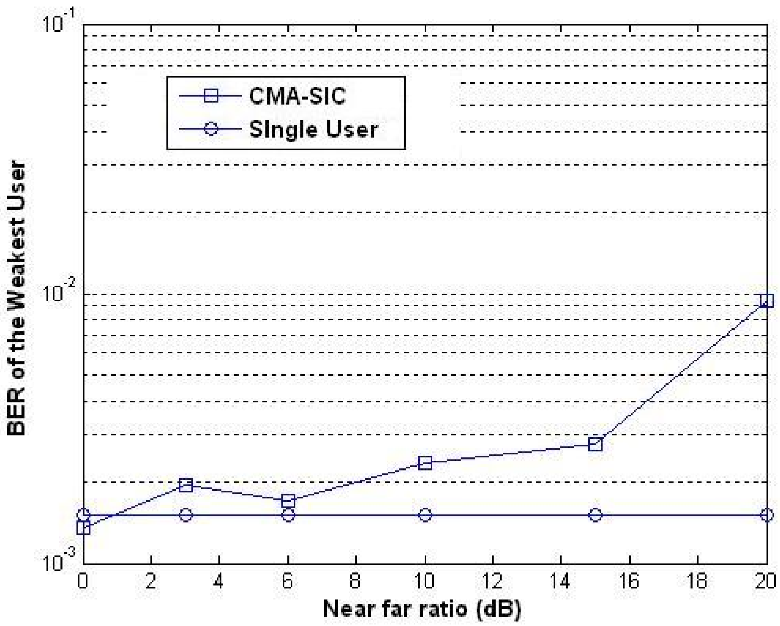


Figure 3.9: Performance of CMA-SIC in flat Rayleigh fading channel with near far ratio of 0-20 dB; $K = 20$, E_b/N_0 of the weakest user=20dB, and Gold sequences of N=31

users. The power of desired weak k^{th} user to unity is fixed and other users are let to transmit at higher power. The power of other users are uniformly distributed between $0 - \lambda dB$ as shown in Figure 3.9. The step size chosen for the algorithm is made variable according to considered nearfar ratio of λ, which is given by $\mu_k = 0.0001/\lambda$. It can be seen in Figure 3.9 that BER performance of CMA-SIC does gracefully degrade with increase in nearfar conditions. However, its near single user performance is retained at the nearfar ratio of as high as 15dB.

In Figure 3.10 the BER performance of CMA-SIC is compared in an AWGN environment with nearfar conditions. The desired user (weakest user) has unity power, while all other users have been assigned powers uniformly distributed between 0 and 10dB. It clearly shows that CMA-SIC is more robust compared to all receivers in this condition.

The Figure 3.11 shows the normalized spectral efficiency performance of CMA-SIC in Rayleigh fading channels with $K = 20$ users using the expression in (3.25). The variances of noise and interference components are obtained using Monte Carlo Integration of output signals from different receivers (MF, SIC and CMA-SIC) using simulations. The capacity of a single user transmission under Rayleigh fading channel with AWGN is also shown for comparison. As expected, the CMA-SIC shows considerable improvement in the achieved spectral efficiency compared with MF and conventional SIC. The conventional SIC that obtains the amplitude estimates from the output of matched filters are shown to achieve only slightly higher spectral efficiency than MF due to estimation error and imperfect cancellation contributing to noise term within the expression in (3.24).

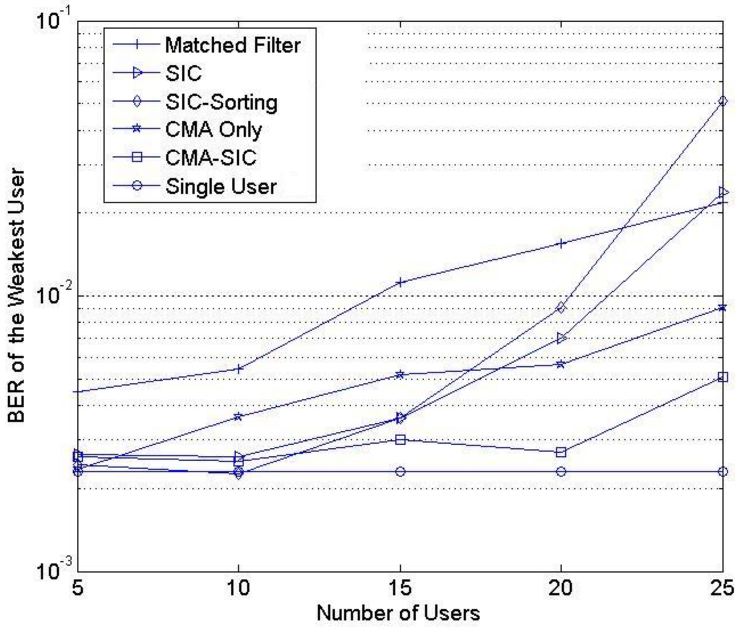


Figure 3.10: Performance of CMA-SIC in AWGN with a nearfar ratio of 10 dB; E_b/N_0 of the weakest user=6dB and Gold sequences of N=31

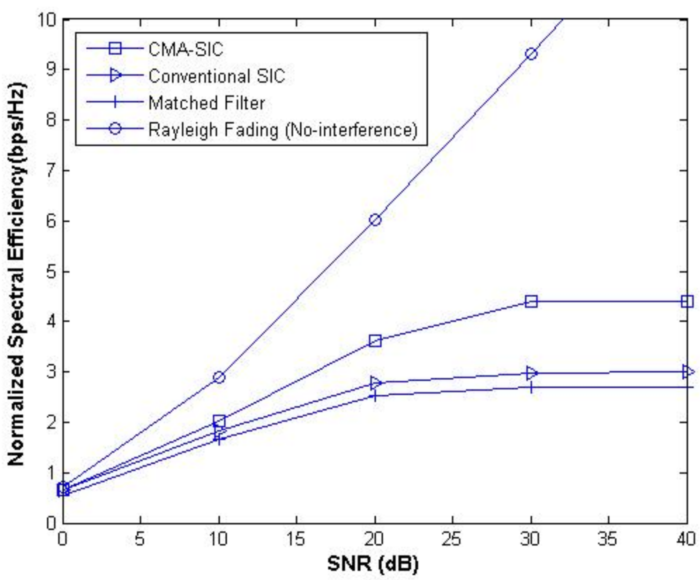


Figure 3.11: Spectral Efficiency of CMA-SIC in Rayleigh flat fading channel, K=20 and Gold sequences of N=31

The effect of different step-size $\mu = 0.00001 - 0.001$ on the BER in Rayleigh flat fading channels is investigated in Figure 3.12. The system loading of $K = 20$ equal power $P_k = 1, \forall k$ users are used and operating of $E_b/N_0 = 20$ and 30 dB are selected. It can be seen that the performance of CMA-SIC is effected by the step-sizes used. Also noted that the $\mu = 0.0001$ used in previous simulations is not the optimum step-size and there is still room for further performance improvement.

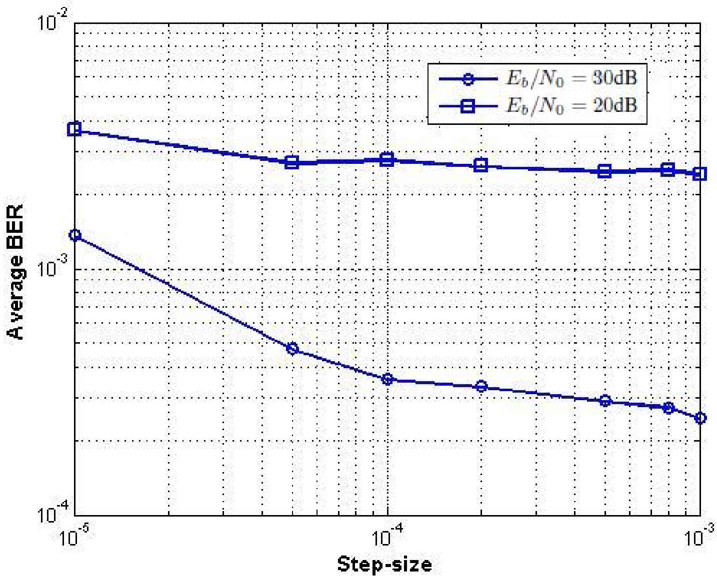


Figure 3.12: The effect of step-size on the performance of CMA-SIC in flat Rayleigh fading channel under different E_b/N_0; K=20 and Gold sequences of N=31

As can be seen from the results, the blind adaptive approach for a SIC has shown to address the imperfect MAI estimation and cancellation very well. In the next Section, the idea is further refined and carefully applied to address the unreliable MAI estimation and nearfar problems that limit the performance of PIC receivers.

3.3 Robust Blind Parallel Interference Cancellation for DS-CDMA with Adaptive Despreader and Improved Interference Estimator

Among the type of IC receivers, PIC is long known as an effective receiver technique that greatly alleviates the MAI problem using multistage architecture and has reasonable latency. There has been large number of work on improving the performance of PIC in the past decade. The PIC proposed by Varanasi et.al. in [51], also known as 'Full PIC', performs complete interference cancellation to obtain final decision for a user. The Full PIC is known to show dramatically degraded BER performance when the system load increases. Identifying that the cancellation terms for such PIC that are generated from the bank of MF output, do not provide reliable MAI estimates for the decision statistics, partial cancellation approach for PIC is proposed by Divsalar et.al. [52]. The partial PIC approach employs cancellation factors less than unity at the initial stages to prevent excessive error propagation problem of the full PIC. At each subsequent stage the factors are increased, approaching to unity, assuming that the reliability of MAI estimates has improved gradually. It is well known that selecting single weight for all users is not optimal way and the weights need not be necessarily less than one [62, 61]. Therefore, adaptive multistage PIC [61] is proposed, that uses set of weights at each stage for each individual user that are calculated from the mean square error criteria of LMS algorithm for the given received signal and sum of the users' signal estimates.

The works in [52, 61, 51] have considered the design of PIC, primarily, in AWGN channel. When the PIC is applied in fading channels, its performance is significantly degraded due to additional impairment, which the fading process causes on top of MAI [92]. The PIC that uses soft (linear) cancellation obtained from MF output at each stage [47, 52] are used in fading channel conditions and referred to here and afterwards as 'Conventional PIC'. Recently weighted linear PIC [62] in Rayleigh fading channels is proposed to improve the Signal to Interference Ratio (SIR) at each stage of the PIC. Optimum weights for interference cancellation for each user is derived using the knowledge of users' signal cross correlations for maximizing the average SIR. This has shown to provide considerable performance improvement compared to the conventional PIC in the same system environments. As the scheme only considers the statistics of users' channels, it may not perform well in nearfar user conditions. It is very important to note that a partial weight is not what is essential for improved performance of the IC receivers, but the accurate amplitude estimate of the desired user's signal. The difficulty in obtaining accurate channel estimates and bandwidth expansion involved in sending training sequences motivates us to consider blind techniques that do not require channel parameter estimation.

In this contribution, a novel blind adaptive PIC, also termed as BA-PIC, is proposed to address the error propagation and amplitude estimation problems of existing PICs for operation in Rayleigh flat fading and nearfar channel environments. The BA-PIC employs constant modulus algorithm within its despreaders and uses the adaptive weights of CMA at each stage to obtain the amplitude estimates for the users at every symbol period. Both CMA and PIC are not new techniques and the literature on these techniques is abundant and are listed in [17] and references therein. Some work employing CMA in blind multiuser detection for CDMA have been reported in the past. However, in fading and nearfar channel environments, the CMA receivers suffer from serious problems of ill convergence, locking to an interferer's rather than desired user's signal [58]. The IC receivers, on the other hand, are known to operate very well in such environments.

Observing the shortcomings of both CMA and PIC techniques and also their advantages, a new approach that uniquely combines these techniques has been identified to achieve significantly enhanced performance for multiuser detection in CDMA compared to either techniques. More specifically, the knowledge of finite alphabet property of users' data symbols is exploited to improve the detection performance at each PIC stage, showing some resemblance to the principles of iterative multiuser receiver techniques for CDMA [93]. The proposed receiver has many desirable properties: besides being blind and low complexity PIC, it also supports large number of users and is extremely nearfar resistant. To assess the performance of the proposed PIC approach, an error probability analysis is presented, using Gaussian Approximation of signals generated at each stage at the CMA output. Simulation results are shown to confirm the analysis and to show that significant performance gain compared to the conventional PIC can be achieved under various system loads and channel conditions considered. Novel contributions of the proposed PIC receiver design approach compared with existing PICs such as [51, 52, 61] is highlighted in the next paragraph.

Adaptive despreading at chip level is proposed to minimize the error variance of each user's symbol decision. This is unlike all existing PICs that use conventional fixed correlation for despreading. Furthermore, the receiver algorithm uses the despreaders' weights to continuously track the power variation of each user's received signal, which is essential in PIC processing for more reliable detection of users' data, specially under time varying fading environments. Adaptive pre-respreader interference estimation is proposed, in which online adaptive scaling factors are generated during every symbol period using the average magnitude of user's spreading sequence and weights of the adaptive despreader. This method provides more accurate cancellation of users' MAI contributions at each stage and unlike conventional and partial cancellation based approaches, does not require optimization of cancellation weights for different system loading environments while also performing very well even under severe near far conditions.

The rest of the contribution is organized as follows. In subsection 3.3.1, a generalized system model of the proposed PIC is presented. In subsection 3.3.2, the principles of the proposed Blind PIC receiver technique are formulated. The detection and cancellation algorithm and complexity analysis are carried out in subsection 3.3.3. In subsection 3.3.4, the BER performance and spectral efficiency analysis at different stages of the proposed PIC is presented. The simulation performance results and comparisons under different system conditions are shown in subsection 3.3.5.

3.3.1 System Model

A synchronous DS-CDMA system of K users under Rayleigh flat fading and AWGN channel conditions is considered. The received composite signal $r(t)$ can be written as:

$$r(t)=\sum_{k=1}^{K}\beta_k(t)s_k(t)+v(t), \tag{3.26}$$

where $s_k(t) = \sqrt{P_k}b_k(t)c_k(t)$ is the transmitted signal of k^{th} user, P_k is the signal power, $\beta_k(t) = g_k(t)e^{-j\pi\phi_k(t)}$ is the complex fading channel of unit variance $\sigma^2 = 1$ and consisting of amplitude $g_k(t)$ and phase $\phi_k(t)$ components, respectively. $b_k(t) = \sum_{m=-\infty}^{\infty} b_k(m)p(t - mT_b)$ is the data signal, where $b_k(m)$ is a binary sequence taking values $[-1, +1]$ with equal probabilities, $p(t)$ is rectangular pulse with period T_b. The spreading sequence is denoted as $c_k(t) = \sum_{n=-\infty}^{\infty} c_k(n)p(t - nT_c)$ with antipodal chips $c_k(n)$ of rectangular pulse shaping function $p(t)$ with period T_c and normalized power over a symbol period equal to unity $\int_0^{T_b} c_k(t)^2 dt = 1$. The spreading factor is $N = T_b/T_c$ and $v(t)$ is the AWGN with two sided power spectral density $N_0/2$. Throughout the Chapter, it is assumed that coherent phase reference of all users' signals is available. Without loss of generality, it is assumed that k^{th} user is the desired user. To gain better exposition of our blind PIC approach, the operation of conventional and weighted PIC is briefly described next.

Conventional and Weighted PIC Techniques

Although we are referring to PIC used in [47] and [52] as described earlier to obtain a conventional PIC, it's all the main functional blocks are the same with other PICs including the proposed receiver. Therefore, the signal model described here applies equally well to the proposed PIC.

At every symbol period m, the received signal is first chip matched filtered and sampled to form received signal vector $\mathbf{r}(m)$. In a conventional PIC, first the initial estimation of k^{th} user's data signal at stage 0 ($l = 0$) is carried out by obtaining a decision variable signal $z_k^0(m)$ from the output of bank of MFs matched to users' spreading sequences as follow:

$$\begin{aligned} z_k^0(m) &= \int_{(m-1)T_b}^{mT_b} \{\mathbf{r}(m)\}^T \mathbf{c}_k; \forall k \\ &= \sqrt{P}_k(m)\beta_k(m)b_k(m) + \sum_{i=1, i\neq k}^{K} \sqrt{P}_i \rho_{ki}\beta_i(m)b_i(m) + v_k(m) \\ &= D_k(m) + I_k(m) + v_k(m) \end{aligned} \tag{3.27}$$

where $\{\mathbf{r}(m)\}^T$ denotes transpose of $\mathbf{r}(m)$, $\mathbf{c}_k$ is a vector consisting of the k^{th} user's spreading sequence, ρ_{ki} is the magnitude of cross-correlation between k^{th} and i^{th} users' spreading sequences, $D_k(m)$ and $I_k(m)$ are denoted as the desired and interfering users' received data signals, respectively, $v_k(m)$ is the correlated AWGN term. The estimation of the desired k^{th} user's data $b_k(m)$ in the l^{th} stage $1 \leq l \leq L$, where L is the total number of PIC stages, is carried out by subtracting from $\mathbf{r}(m)$ the sum of MAI contribution of other users. The MAI estimates are formed by summing the remaining users' decision variables from previous stage $z_i^{l-1}(m); i \neq k$ multiplied by their spreading sequences $\mathbf{c}_i$. This process is shown as follows:

$$z_k^l(m) = \int_{(m-1)T_b}^{mT_b} \Big\{\mathbf{r}(m) - \lambda \sum_{i=1, i\neq k}^{K} z_i^{l-1}(m)\mathbf{c}_i\Big\}^T \mathbf{c}_k; \forall k, \tag{3.28}$$

where λ is the weighting factor used for the MAI cancellation. The conventional PIC as given in [47] and [52] is obtained from (3.28) by setting $\lambda = 1$. The conventional MF receiver can be derived as a special case of the PIC described here by setting $\lambda = 0$. The Partial PIC [52] and also the Weighted PICs as described in [62] can be obtained from (3.28) by setting a positive and non-zero value of λ. The estimate of the k^{th} user data is then obtained as follows:

$$\hat{b}_k^l(m) = sgn\Bigg[Re\Big\{z_k^l(m)\beta_k^*(m)\Big\}\Bigg]; \forall k, \tag{3.29}$$

where $sgn\{.\}$, $Re\{.\}$ and * denote sign, real and complex conjugation operation, respectively. The processes (3.28) and (3.29) are carried out at each stage of the PIC for all users' data signals to be estimated/detected.

Although the conventional PICs improve the detection performance of CDMA as compared to the simple MF receivers, they often suffer from serious error propagation problem [47]. This is due to imperfect MAI estimates generation and cancellation of the conventional PICs; because the MF outputs employed at each stage introduce bias into decision variables which increases linearly with increase in number of users. The problem becomes more significant when the users' received signal power are different; giving rise to the nearfar problem. Therefore, to overcome the shortcomings of conventional PICs and also to avoid the need for channel parameter estimation, a novel blind PIC approach is proposed, which not only is practical, but also offers robust operation

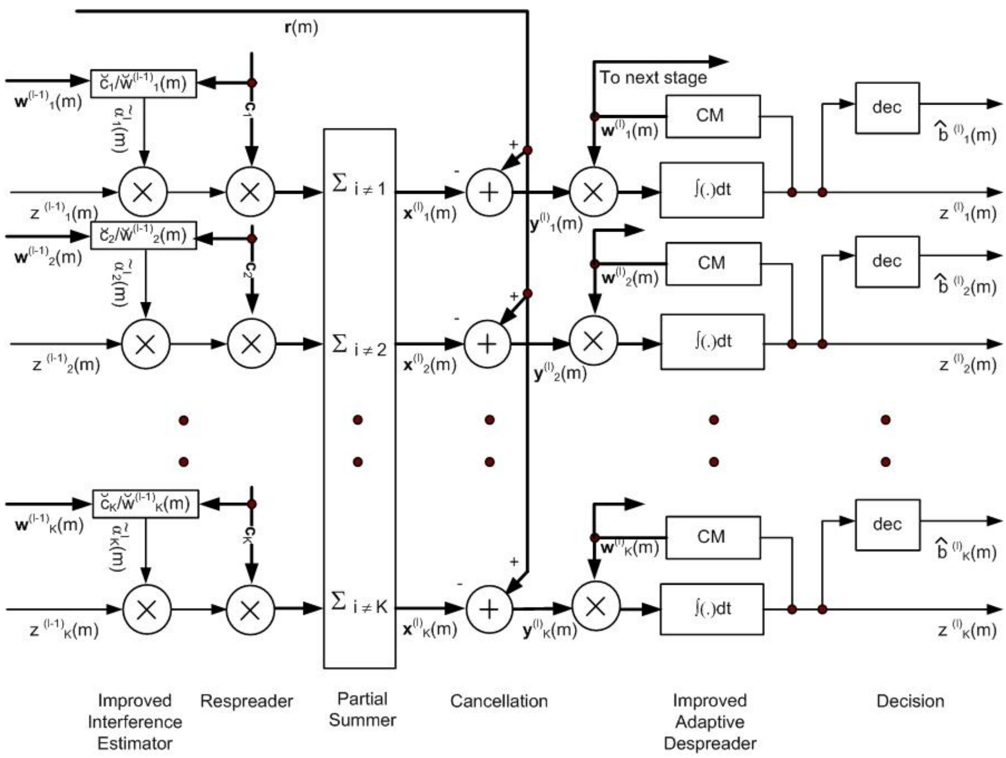


Figure 3.13: Architecture of l^{th} stage of BA-PIC

in different system load and channel conditions. It employs an adaptive algorithm using the constant modulus property of users' signals to address the problems of conventional PIC in two ways and will be described next.

3.3.2 Principles of the Proposed Blind Adaptive PIC (BA-PIC)

Basic idea of the proposed PIC is to first detect desired user' data signal by blindly suppressing interfering users' signals during the despreading process and then employ interference cancellation stages, using blindly obtained amplitude estimates and spreading sequences of the interfering users to further improve the detection performance. The estimation and detection process of the proposed PIC is different from other PICs [47, 52, 62, 61]. In addition to employing adaptive weighted interference cancellation, it also exploits the finite alphabet of users' signals to further improve the detection performance of individual user's signal. More specifically, the CM algorithm is utilized here for the dual purpose of interference suppression for the initial detection and also amplitude estimation for interference cancellation for the desired user. To reduce severe error propagation due to bias problem of conventional fixed correlation based despreading, adaptive despreading is considered. This process attempts to restore the constant modulus of each user's output signal by multiplying the input signal vector with adaptively weighted chips of spreading sequence of each user updated by the CM criterion. Also, to minimize the risk of wrong interference cancellation, the soft output signals from previous stage of the users that serve as inputs to respreaders at current stage are also individually weighted. Unlike partial weights used for interference cancellation in other PICs [47, 52, 62, 61], the MMSE amplitude estimate of each user's signal is used in the proposed PIC. This is obtained by using the soft output signal at the given stage and weights of the CM algorithm from the previous stage.

The system architecture of (l^{th}) stage of the proposed PIC is shown in Figure 3.13. In this architecture, the effect of all interferering users' signals are first removed to aid the detection of the desired user. At every symbol period, $r(t)$ is sampled at the chip rate to form the vector $\mathbf{r}(m)$ of length N chips. Adaptive CM algorithm embedded within the despreader is used to adjust the desired user's signal amplitude using a weight for each chip signal; thus providing N degree of freedom for amplitude adjustment. A block diagram of an adaptive CMA aided despreader of the proposed PIC is shown in Figure 3.14. The initial stage $l = 0$ of the PIC is obtained from a bank of K adaptive despreaders with their common input being the received signal $\mathbf{r}(m)$. In the successive stages $l > 1$, the input signal vector $\mathbf{y}_k^l$ for each user is obtained by canceling from $\mathbf{r}(m)$ the sum of MAI contributions of interfering users' signals $\mathbf{x}_k(m)$. The output signal of each user $z_k^l, \forall k$ is also weighted utilizing the estimate $\tilde{\alpha}_k^l(m)$ obtained from the CM criterion from the previous stage, spread and subtracted from the received signal to form the input to the next stage $\mathbf{y}_k^{l+1}$. The same process can be repeated for multiple stages with each additional stage contributing to further improved performance.

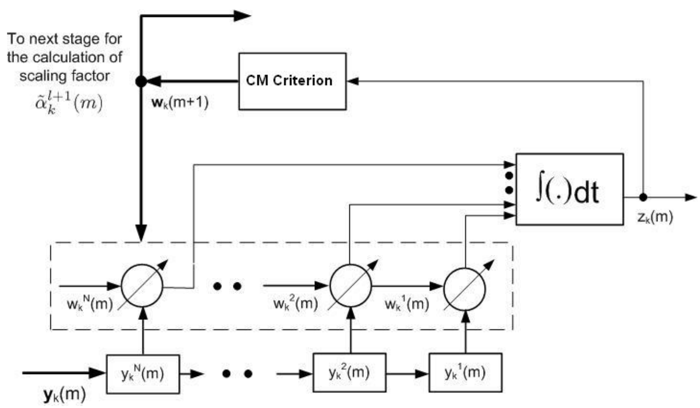


Figure 3.14: Adaptive despreading and weights generation methods of BA-PIC for k^{th} user

In any CDMA receiver, the decision error for the desired user signal occurs when the effects of the total MAI contributions flip the polarity of the user's decision variable signal $z_k(m)$ or causes the combined MAI and user's signal magnitude to be very near to zero so that receiver noise becomes the dominant source of the decision error. The two conditions for probability of decision error can be shown as:

$$Pr\{\hat{b}_k(m) \neq b_k(m)\} \longrightarrow \begin{cases} Pr\{D_k(m) - \{I_k(m) + v_k(m)\} < 0\} & \text{, if } b_k(m) = +1 \\ Pr\{D_k(m) - \{I_k(m) + v_k(m)\} > 0\} & \text{, if } b_k(m) = -1 \end{cases} \tag{3.30}$$

Since the polarity of $D_k(m)$ can take only two values, from (3.30), assuming that the CM criterion at each stage always forces the output signal $z_k(m)$ towards the correct polarity of transmitted signal $b_k(m)$, conditions for minimizing the decision error can be set for the proposed blind receiver as follows:

$$Pr\{\hat{b}_k(m) \neq b_k(m)\} \longrightarrow 0, \text{if} \begin{cases} 1) \mid D_k(m) \mid^2 \longrightarrow 1 \\ 2) \mid I_k(m) + v_k(m) \mid^2 \longrightarrow 0 \end{cases} \tag{3.31}$$

The first condition in (3.31) can be met asymptotically for static and noiseless channel conditions by a receiver employing any form of CM criterion for the given input signal [87, 58, 86, 55]. In wireless environments, the users' signal power vary from near zero to several (or tens) dB. In such conditions, the use of CM criterion does not guarantee stable recovery of desired user's data signal, particularly when the strong interference signals are present in the system. Our objective is to recover the users' data signals in realistic wireless environment with fading and nearfar conditions. Therefore, it is intuitive to incorporate an IC stage to enhance the detection of the despreader output. Since all the users' signals are detected simultaneously in a PIC, accurate estimation of both the desired user's and MAI users' signal power variations need to be taken into account for the detection of desired users' data signals. The proposed receiver attempts to satisfy both conditions of (3.31) for the minimum decision error with its unique design approach and is described next.

The idea behind the detection and amplitude estimation process of the proposed blind PIC is elaborated by considering a typical scenario for a user. At a given symbol period m, it is assumed that due to the channel effect the magnitude of desired k^{th} user's data signal is positive but very near to zero $D_k(m) \simeq 0$. When a simple MF (correlation) receiver is employed, the probability that the magnitude of the receiver's output signal $z_k^l(m)$ taking the correct symbol polarity is low because of its inability to counter the MAI effects. When an additional CM criterion to the receiver is applied, it keeps track of previous signal magnitudes and multiplies the current input signal $y_k^l(m)$ with larger weights to make the absolute magnitude of the output signal to approach unity i.e. $|z_k^l(m)| \longrightarrow 1$. It is intuitive that with high probability, the weighted output signal will be forced towards the correct polarity rather than to the opposite polarity (which will introduce the decision error). Due to slow convergence of CM algorithm, this process of weighting the signal alone does not ensure that the symbol decision is correct. The terms $\{I_k(m) + v_k(m)\}$ in (3.31) must also be minimized. The use of interference cancellation is appropriate for such user signal environments [47, 52, 92, 62, 61, 50]. The cancellation technique when applied accurately, eliminates the effects of strong MAI signals and leads to ideal recovery of desired user's data signal. However, inaccurate cancellation seriously degrades the performance of a PIC and its performance can be even worse than when not performing any cancellation [61].

The cancellation weights in the proposed PIC are chosen, considering the conditions of decision error due to each users' independent and varying amount of MAI contribution as shown in (3.30). Since the desired user's signal is weak, the cancellation weight for this user must also be small as compared to unity so that its individual MAI contribution can be accurately canceled. Conversely, assuming that in another time interval, the magnitude of the user's signal has increased as such to become greater than unity $D_k(m) > 1$, then CM algorithm generates smaller multiplication weights to make the output signal magnitude $|z_k^l(m)|$ near to unity. The cancellation weight to be applied now must be larger than unity so that larger amount of signal proportional to the magnitude of user's signal $\mid D_k(m) \mid$ will be canceled from the total received signal. The weight calculation process of the CM algorithm within the proposed PIC can be shown below:

$$\mathbf{w}_k^l(m+1) = \begin{cases} a_k^l(m)\mathbf{w}_k(1) & \text{, if } z_k^l(m) < 1 \\ \frac{1}{a_k^l(m)}\mathbf{w}_k(1) & \text{, if } z_k^l(m) > 1 \end{cases} \tag{3.32}$$

where $a_k^l(m)$ is a positive scaling factor that the CMA uses to adjust the elements of the weight vector $\mathbf{w}_k(m)$ during each symbol period, $\mathbf{w}_k(1)$ is the initial weight vector of the k^{th} user, which is set as a vector of unit norm consisting of chips of the k^{th} user's spreading sequence $\mathbf{c}_k$ i.e. $\mathbf{w}_k(1) = \mathbf{c}_k; \forall k$. The estimate $\tilde{\alpha}_k^l(m)$ of the $a_k^l(m)$ is used in the receiver for interference cancellation. When $\tilde{\alpha}_k^l(m) z_k^l(m) \longrightarrow D_k(m)$, the cancellation process completely removes the effect of k^{th} user's MAI contribution so that other users' signals will be detected and estimated more accurately. Practical CM algorithms, however, does not perform perfectly and there are bound to be some inevitable misconvergence problems. Also in the presence of strong interferering signals, CM algorithm may not be able to lock to the desired weak user. However, the useful properties of the CM algorithm is exploited and its performance is enhanced using PIC stages in the proposed receiver. Their joint operation within the receiver has shown to minimize the degrading effects of both techniques. Since obtaining exact closed-form expressions for the analyzing the performance of stochastic process of CM algorithm in the proposed receiver is very much involved, we have resorted to approximations and simplifying assumptions to evaluate the performance of the receiver at each stage. As will be shown later, the performance results obtained clearly show that they are very much indicative of our design approach as expected. The detail algorithm for each stage of the proposed blind multistage PIC receiver is described next.

3.3.3 BA-PIC Algorithm

At the first symbol period, the weights of CM criterion used for all users at all stages are initialized with vectors of the users' spreading sequences $\mathbf{w}_k^l(1) = \mathbf{c}_k, \forall k, \forall l$. The chip matched filtered received signal for reference at each stage is the same and is denoted as the vector $\mathbf{r}(m)$. The MAI signal for the desired k^{th} user consisting of sum of interferering users' signal estimates at l^{th} stage is denoted as $\mathbf{x}_k^l(m)$. The less interfered input signal vector $\mathbf{y}_k^l(m)$ at each stage is obtained by subtracting $\mathbf{x}_k^l(m)$ from the receiver signal $\mathbf{r}(m)$. The output signal from l^{th} stage serves as input to $l+1$ stage and so on. This iterative detection process is also extended to multiple stages. The input to the initial stage of the proposed PIC, is set as the received signal vector

$$\mathbf{y}_k^0(m) = \mathbf{r}(m). \tag{3.33}$$

For the purpose of clarity, the processes of the proposed PIC algorithm is described for l^{th} stage only. The processes for other stages are obtained similarly with appropriate modifications.

At the l^{th} stage, the decision statistic for k^{th} user $z_k^l(m)$ is obtained by multiplying elements of input signal vector $\mathbf{y}_k^l(m)$ with the elements of weight vector $\mathbf{w}_k^l(m)$ and the products summed over a symbol period (The procedures described apply to all other users $\forall i, i \neq k$ and their corresponding signals and are performed in parallel with k) is given by

$$z_k^l(m) = \left\{\mathbf{w}_k^l(m)\right\}^T \mathbf{y}_k^l(m). \tag{3.34}$$

The CMA criterion J_{CM} applied to $z_k^l(m)$ can be written as minimization of the following cost function

$$J_{CM} = E\{z_k^l(m)^2 - \gamma\}^2 \tag{3.35}$$

where $E(.)$ is the expectation operator, γ is the dispersion constant, which is equal to unity for binary phase shift keying (BPSK) signals. The instantaneous error signal $e_k^l(m)$ is calculated as

$$e_k^l(m) = z_k^l(m)\{z_k^l(m)^2 - \gamma\}. \tag{3.36}$$

The estimated gradient vector of the error signal is then calculated as

$$\nabla_k^l(m) = \mathbf{y}_k^l(m) e_k^l(m). \tag{3.37}$$

Using the gradient, the weight vector for next symbol $\mathbf{w}_k^l(m+1)$ is updated as follows

$$\mathbf{w}_k^l(m+1) = \mathbf{w}_k^l(m) - \mu_k \nabla_k^l(m), \tag{3.38}$$

where μ_k is a small step-size that is used to adapt the elements of the weight vector to minimize the cost function (3.35). Finally, the output signal $z_k^l(m)$ is delivered to the decision making process to perform decision on k^{th} user data at l^{th} stage as follows

$$\hat{b}_k^l(m) = dec\{z_k^l(m)\}, \tag{3.39}$$

where $dec\{.\}$ is simply a sign function for BPSK data.

The cancellation process requires amplitude estimate of the detected user's signal along with the user's spreading sequence. An estimate of the scaling factor is first obtained, using the weights of the CM algorithm and the known spreading sequence of the user as follows

$$\tilde{\alpha}_k^{l+1}(m) = \frac{\breve{c}_k(m)}{\breve{w}_k^l(m)}, \tag{3.40}$$

where $\breve{c}_k(m)$ and $\breve{w}_k^l(m)$ are the mean amplitude of chips of user's spreading sequence and elements of the weight vector updated by the CM algorithm, respectively and are given by

$$\breve{c}_k(m) = \frac{1}{N}\sum_{n=1}^{N}\Big|c_k\{(m-1)N+n\}\Big| \tag{3.41}$$

$$\breve{w}_k(m) = \frac{1}{N}\sum_{n=1}^{N}\Big|w_k\{(m-1)N+n\}\Big|. \tag{3.42}$$

The estimated data signal of a user $z_k^l(m)$ is then scaled with $\tilde{\alpha}_k^{l+1}(m)$ and spread with $\mathbf{c}_k$ to generate the cancellation term for k^{th} user in the next stage. In similar manner, the MAI estimates for the k^{th} user is obtained by summing the cancellation signals from all other users $i, i \neq k$ spread with their corresponding sequences to form the signal $\mathbf{x}_k^l(m)$ as follows

$$\mathbf{x}_k^{l+1}(m) = \sum_{i=1,i\neq k}^{K} \tilde{\alpha}_i^{l+1}(m) z_i^l(m)\mathbf{c}_i(m). \tag{3.43}$$

The improved signal estimate for k^{th} user for the next $l+1$ stage is obtained after the interference cancellation as follows:

$$\mathbf{y}_k^{l+1}(m) = \mathbf{r}(m) - \mathbf{x}_k^{l+1}(m). \tag{3.44}$$

The signal vectors of the users $\{\mathbf{y}_1^{l+1}, \mathbf{y}_2^{l+1}, ...\mathbf{y}_K^{l+1}\} \in \mathbf{Y}^{l+1}$ serve as inputs to the next stage for further improvement of each user's data signal estimation. The processes derived in (3.34)-(3.44) are repeated with new signal sets for desired number of stages.

3.3.4 System Analysis

3.3.4.1 BER

In this subsection, a simplified analysis of bit error probability for the proposed PIC is provided, using the GA method [36] and compares with that of Conventional PIC under flat fading channel conditions. Under equal average users' power conditions, the BER for all users are approximately equal at any stage of a PIC . Therefore, BER analysis is presented for k^{th} user in the system for clarity in expressions.

The average probability of error for the desired user at the output of the l^{th} stage of PIC is obtained by calculating the BER corresponding to each fading SINR distribution of the user's signal given by

$$Pe^l = \frac{1}{2}\left[1 - \sqrt{\frac{\overline{SINR}^l}{1+\overline{SINR}^l}}\right], \tag{3.45}$$

where $\overline{SINR}$ is the average SINR calculated from the channel realization $g_k(m)$ of the desired k^{th} user at the l^{th} stage, given by

$$SINR^l = \left[\frac{E\{(z^l)^2\}}{var\{z^l\}}\right] \tag{3.46}$$

where, $var\{z^l\}$ is the variance of the decision variable signal at the output of l^{th} stage of the PIC.

The proposed PIC receiver algorithm keeps track of channel realization of each user at every symbol period during the detection and estimation processes. Therefore, the magnitude of channel autocorrelation over a certain time period is important for the receiver design. The Jake's model [30] of fading channels is used, which consists of sum of uniform scatterers represented by individual sinusoids defined by the carrier frequency f_c, Doppler rate f_d. The channel autocorrelation $R(\tau)$ over a time period τ is obtained as follows:

$$R(\tau) = J_0\Big(2\pi f_d \tau\Big), \tag{3.47}$$

where $J_0(.)$ is the zero order Bessel function of first kind. For the clarity in the analyses to follows, subscripts 'ba-pic' and 'conv-pic' are used to distinguish the signals of the proposed and conventional PIC respectively.

Stage 0

Observing the expression of (3.45), it is noted that SINR value is maximized when the variance of the decision variable $z^l(m)$ is minimized. The use of CM criterion tries to achieve this goal by adapting the weights recursively using the weights of previous symbols as such to maintain the absolute magnitude of the output signal equal to the dispersion constant γ. This minimizes the instantaneous error signal $e^l(m)$ as shown in (3.35), attempting to satisfy the first condition of (3.31). The continuous process of learning and the use of N adaptive weight elements for despreading during every symbol, allows the proposed receiver to improve the detection performance compared to conventional MF based correlation receiver. The blind nature of the CM algorithm does not require any information about the statistics of the interfering users' signals. Instead, the composite response due to noise and interference are embedded within the error signal $e^0(m)$. The task of our analysis is to prove that the variance of the despreader output is less than that of the MF output, shown as follows:

$$\frac{var\Big\{z^0_{ba-pic}\Big\}}{var\Big\{z^0_{conv-pic}\Big\}} < 1. \tag{3.48}$$

The despreading process used in conventional PIC can be considered as special case of our CM adapted despreading given by $\mathbf{w}_k(m) = \mathbf{c}_k, \forall m$. The degree to which despreading within a conventional PIC is resilient to MAI impairments is dependent on the length of the vector $\mathbf{c}_k$ also called the processing gain N. In the despreading process of the proposed PIC, each chip elements of a user's sequence is adapted to meet the constant modularity of the output signal $z^l_k(m)$. To see the effect of CM adaptation on the SINR improvement, the case of conventional dispreading is first analysed, i.e. $\mathbf{w}_k(m) = \mathbf{c}_k, \forall m$, the decision variable $z^l_k(m)$ obtained consists of desired data signals and sum of noise and interference components shown as in (3.27).Assuming that the channel correlations do not change significantly over the given observation period of M symbols, the variance at the output of MF (initial stage of conventional PIC) for the desired user k can be shown as:

$$var\{z^0_{conv-pic}\} = \frac{1}{MNR(\Delta t)} \sum_{m=1}^{M} \sum_{i=1, i\neq k}^{K} \rho_{ki}\{g_i^2(m)\} + \frac{N_0}{T_b}. \tag{3.49}$$

It is noticed that the variance of the MF is dependant on the average variances of the users' respective channels and partly on the channel correlations R at the end of the period $\Delta t = MT_b$. In the case of adaptive despreading using CM criterion, the MAI and noise of (3.27) are replaced by instantaneous error values $e^l(m)$ and associated parameters such as vector length N and step size μ. For the observation period of M symbols, the variance of the output signal of CM despreader is given by

$$var\{z^0_{ba-pic}\} = \frac{K}{MR(\Delta t)} \sum_{m=1}^{M} \frac{\{e^0(m) \pm \mu\}^2}{N}. \tag{3.50}$$

From (3.50), it is observed that the error variance $var\{e^0\}$ for a given time period depends upon system parameters, such as number of users K in the system, length of observation period M, the channel fading rate defined by the Doppler shift (frequency), the degree of freedom for weight adaptation N and the choice of step size μ [16, 59]. Therefore, the variance will be minimized for a given period of M symbols, when N is sufficiently large and $R(\Delta t)$ does not change significantly for the period. Assuming $R(\Delta t) \approx 1$ during the period of M symbols, due to recursive nature of the CM cost function, $var\{z^0_{ba-pic}\}$ for a given period M can also be represented as the averaged value of a geometric series consisting of sum of decaying numbers $e^0(m)$ (the rate of decay can be assumed constant for simplicity)

$$var\{z^0_{ba-pic}\} = \frac{1}{M}\left\{var\{z^0_{conv-pic}(1)\} + K\sum_{m=2}^{M} \frac{\{e^0(m) \pm \mu\}^2}{N}\right\} < var\{z^0_{conv-pic}\}. \tag{3.51}$$

Here it is observed that the variance of the adaptive despreader output is exactly the same as that of MF output when $M = 1$ (due to CM algorithm initialization) and reduces monotonically (although in practice it is stochastic from sample to sample) with time within the period of M. Since the fading channels characterized by time varying $R(\Delta t)$ are being considered, the reduction in variance will not be constant. However, it can be concluded that provided that the channels do not decorrelate significantly over a period of several symbols , the application of CM criterion always reduces the variance of decision statistics z^l as compared to conventional despreading and, thus, leads to higher output SINR. Assuredly the despreader output (error) can be assumed Gaussian distributed. For ease in analysis, it is assumed that the true polarity of desired k^{th} user's signal is known to CM criterion i.e. $\left\{\left|sgn\{E\{g_k(m)b_k(m)\}\right| = \gamma\right\}$ and $var\{z^0_{ba-pic}\} =$

$E\{e^0(m)\}^2$, then the SINR at the output of CMA can also be shown as:

$$\overline{SINR}^0_{ba-pic} = \frac{E^2\{z^0_{ba-pic}\}}{var\{z^0_{ba-pic}\}} \approx \frac{\gamma^2}{var\{z^0_{ba-pic}\}}. \tag{3.52}$$

The average probability of error Pe^0_{ba-pic} of a desired k user using the proposed blind receiver at the output of initial stage is obtained using the derived SINR (3.52) and BEP expression of (3.45) conditioned on user's channel g_k as derived earlier.

Stage 1

The BEP analysis for signals at the output of the first stage of PIC $l = 1$ is now shown. The IC stages $l \geq 1$ in the proposed receiver are used to ensure that the second condition in (3.31) is met. The IC process regenerates the MAI estimates $\hat{I}_k(m)$ using the output signals of previous stage $z_i^{l-1}(m)$, spreading sequence of the users $\mathbf{c}_i$ and the amplitude estimates of users $\tilde{\alpha}_i(m), \forall i, i \neq k$ to minimize the effects of $I_k(m)$. The process can be shown as:

$$z^1_{ba-pic}(m) = v_k(m) + \sqrt{P}_k(m)\tilde{\alpha}^0_k(m)z^0_k(m) + \left\{ \sum_{i=1, i\neq k}^{K} \sqrt{P}_i \rho_{ki} g_i(m) b_i(m) - \sum_{j=1, j\neq k}^{K} \sqrt{P}_j \rho_{kj} \tilde{\alpha}^1_j(m) z^0_j(m) \right\} \tag{3.53}$$

where subscripts j denote the indices of regenerated users' signals. As the necessary condition to minimize the probability of error, very accurate estimates of amplitudes g_k for all users are required. In other words, following minimum mean squared error (MSE) ε^2_k condition for amplitude estimation is satisfied:

$$\varepsilon^2_k = E\Big\{(\tilde{\alpha_k}\big|z_k\big| - g_k)^2\Big\} \longrightarrow 0, k = 1, 2, .., K. \tag{3.54}$$

It will not be difficult to see, when (3.54) is satisfied, this also leads to perfect cancellation of interfering users' signals i.e. $j = i; \forall j$. To assess the accuracy of the amplitude estimation of the proposed PIC from $\tilde{\alpha_k}$ as shown in (3.40), it is first assumed that the CMA criterion for k^{th} user signal has converged by generating perfect weights $\bar{\mathbf{w}}_k(m)$ for the given input signal $z_k(m) = \bar{\mathbf{w}}_k(m)^T \mathbf{y}_k(m)$ as follows:

$$e_k(m) = z_k(m)(z_k(m)^2 - \gamma)^2 \longrightarrow 0. \tag{3.55}$$

It is known that when CM algorithm is used for blind channel estimation in an interference free channel, the estimate $\hat{g}_k(m)$ is simply obtained as the inverse of the weights [94]. Since the users' signals are sampled at chip rate in CDMA systems, the estimate can be obtained as:

$$\hat{g}_k(m) = \frac{\gamma}{\hat{w}_k(m)} \longrightarrow \frac{\sum \mathbf{c}_k}{\sum \mathbf{w}_k(m)}. \tag{3.56}$$

The weight vector $\mathbf{w}_k(m)$ consists of N adaptive elements, which serve the purpose of both suppressing the MAI and also provides unbiased estimates of users' data signals [95]. In this work, a scaling factor $\tilde{\alpha}_k(m)$ as shown in (3.40) is used, along with CM despreader output at previos stage $z_k^{l-1}(m)$ for obtaining the joint data and amplitude estimate of k^{th} user [16].

Now the task of analysis is to show that the variance of the proposed PIC at stage 1, $var\{z_{ba-pic}^1\}$ is smaller as compared to that of the Conventional PIC. This depends on the accuracy of amplitude estimation i.e. ε_k^2 and variance of decision variable in previous stage $var\{z_{cmmv-pic}^0\}$. In conventional PIC the data signal estimate for k^{th} user $D_k(m)$ is simply taken as the absolute magnitude of MF output $z_k^0(m)$ while ignoring $I_k(m) + v_k(m)$ as in (3.27). Therefore, the variance of decision variable the desired user $var\{z_{conv-pic}^1\}$ can be shown as:

$$\begin{aligned} var\{z_{conv-pic}^1\} &= \sum_{i=1, i\neq k}^{K} \Big\{\rho_{ki}^2 - \kappa_{ij}\rho_{kj}^2\Big\} var\Big\{z_{conv-pic}^0\Big\}; \\ \kappa_{ij} &= \begin{cases} 1 & \text{, if } i = j \\ -1 & \text{, if } i \neq j \end{cases} \; ; j = 1, .., K; j \neq k. \end{aligned} \tag{3.57}$$

In the case of BA-PIC, the variance of decision variable of desired user z_{ba-pic}^1 is dependent upon individual user's amplitude estimation error $\varepsilon_k(m), \forall k$ and also the variance of decision variable for the desired in the previous stage $var\{z_{ba-pic}^0\}$

$$\begin{aligned} var\{z_{ba-pic}^1\} &= \varepsilon_k^2 + \sum_{i=1, i\neq k}^{K} \varepsilon_i^2 \Big\{\rho_{ki}^2 - \kappa_{ij}\rho_{kj}^2\Big\} var\Big\{z_{cmmv-pic}^0\Big\}; \\ \kappa_{ij} &= \begin{cases} 1 & \text{, if } i = j \\ -1 & \text{, if } i \neq j \end{cases} \; ; j = 1, .., K; j \neq k. \end{aligned} \tag{3.58}$$

From (3.57) and (3.58), the following is observed. The reduction in variance of decision statistic of conventional PIC $var\{z_{conv-pic}^1\}$ is dependent on the variance in the previous stage $var\{z_{conv-pic}^0\}$ and how reliable the current stage generates the MAI estimates for cancellation quantified by the residual signal variance $\chi^2 = \rho_{ki}^2 - \rho_{kj}^2$. As noted in (3.57), the variance χ^2 could be high due to inaccurate estimation of the channel power variation $var\{g_i\}$ of interfering users signals. The convential PIC detection becomes vulnerable to MAI particularly when desired user's signal is weak and there are strong interferers. Unlike the conventional and other PIC found in the literature [47, 52, 62, 61], where the cancellation weights are selected based on random guess or signal statistics, in the proposed PIC the cancellation weights are generated using the algorithm that is adaptive in MMSE sense to instantenous changes in the power variations of both desired and interfering users' signals. Based on this heuristic (3.53)-(3.58) and analysis of previous PIC stage, it is not difficult to conclude that the SINR of the proposed PIC is always higher than the conventional PIC and can be shown as below

$$\overline{SINR}_{ba-pic}^1 = \frac{E^2\left[z_{ba-pic}^1\right]}{var\{z_{ba-pic}^1\}} \approx \frac{\gamma^2}{var\{z_{ba-pic}^1\}}. \tag{3.59}$$

The $\overline{SINR}_{ba-pic}^1$ value can be used in (3.45) to obtain the BEP for the desired user signal at the output of stage 1. The extension of the analysis to more stages is straightforward. It can be expected that, each additional stage will further reduce the variance of decision variables of each user and thus lead to further performance improvement.

3.3.4.2 Spectral Efficiency

In this Section, a simple analysis of achievable SE of PIC receivers (Conventional PIC and the proposed BA-PIC) in single cell condition is provided and compared with that of SE of single user under Rayleigh fading with perfect CSI. The SE analysis is carried out from the SINR values obtained at the output of each stage of PIC derived earlier in (3.52)-(3.59).

Stage 0

The SE of conventional PIC at the initial stage $l = 0$, is obtained directly from output of correlators. The signal $z^0_{conv-pic}$ consists of contribution of desired user signal g_k and sum of correlated MAI users' signal components, given by

$$\begin{aligned} z^0_{conv-pic} &= \int_0^{T_b} r(t)c_k(t)dt, \\ &= \sqrt{P}_k g_k b_k + \sum_{i=1, i\neq k}^{K} \sqrt{P}_i \rho_{ki} g_i b_i + n_k \\ &= D_k + I_k + n_k, \end{aligned} \tag{3.60}$$

where ρ_{ki} is the magnitude of cross-correlation between k^{th} and i^{th} users' spreading sequences, D_k and I_k are denoted as desired and interfering users' data signals, respectively, n_k is correlated AWGN term. The most important factor affecting the SE of PIC receivers is the accuracy of amplitude estimation. For the conventional PIC, the estimation of amplitude $\hat{\alpha}_k$ is simply taken as the output of matched filter for the user

$$\hat{\alpha}_k = |z^l_{conv_pic}|; k = 1, .., K. \tag{3.61}$$

Therefore, the SE of conventional PIC receivers is given by:

$$C^0_{conv-pic} \leq \frac{K}{N}\log_2\left[1 + E\{SINR^0_{conv-pic}\}\right] = \frac{K}{N}\log_2\left[1 + \frac{E^2\{z^0_{conv-pic}\}}{var\{z^0_{conv-pic}\}}\right], \tag{3.62}$$

where the MAI effects from $K-1$ users is assumed as Gaussian noise to obtain the SINR in (3.62), the expectation is taken over the fading power distribution of desired k^{th} user. The SE of system employing BA-PIC can be obtained as follows

$$C^0_{ba-pic} \leq \frac{K}{N}\log_2\left[1 + E\{SINR^0_{ba-pic}\}\right] = \frac{K}{N}\log_2\left[1 + \frac{E^2\{z^0_{ba-pic}\}}{var\{z^0_{ba-pic}\}}\right], \tag{3.63}$$

where the expectation is taken over the fading power distribution of desired k^{th} user, $var\{z^0_{ba-pic}\}$ is the variance of decision variable of desired k^{th} user by employing the BA-PIC. The reduced variance of decision variable for BA-PIC is due to the improved interference estimation used for cancellation. Therefore, it is expected that, among the receivers analysed, BA-PIC provides higher spectral efficiency than conventional PIC even at stage 0.

Stage 1

In a similar manner as above, the SE of PIC receivers at stage 1 are obtained as follows:

The SE of conventional PIC receivers is given by:

$$C^1_{conv-pic} \leq \frac{K}{N} \log_2 \left[1 + E\{SINR^1_{conv-pic}\} \right] = \frac{K}{N} \log_2 \left[1 + \frac{E^2\{z^1_{conv-pic}\}}{var\{z^1_{conv-pic}\}} \right], \tag{3.64}$$

where the residual MAI effects from $K-1$ users after interference cancellation is assumed as Gaussian noise to obtain the SINR in (3.64), the expectation is taken over the fading power distribution of desired k^{th} user. The SE of system employing BA-PIC can be obtained as follows

$$C^1_{ba-pic} \leq \frac{K}{N} \log_2 \left[1 + E\{SINR^1_{ba-pic}\} \right] = \frac{K}{N} \log_2 \left[1 + \frac{E^2\{z^1_{ba-pic}\}}{var\{z^1_{ba-pic}\}} \right], \tag{3.65}$$

where the expectation is taken over the fading power distribution of desired k^{th} user, $var\{z^1_{ba-pic}\}$ is the variance of decision variable of desired k^{th} user by employing the BA-PIC. It is not difficult to predict that the SE of BA-PIC at stage 1 is much higher than conventional PIC as it has been observed that much improved amplitude estimation for interference cancellation is obtained at this stage.

3.3.4.3 Computational Complexity

It is well known that the CMA is a low complexity algorithm with computational complexity in order of $O(N)$ per symbol. Table 3.2 shows that the proposed Blind PIC employing the CM algorithm does not significantly increase the receiver complexity and is still within the range of $O(KN)$ computations per symbol per stage. Though the Blind PIC is more computationally complex than conventional PIC, its operation is based on very low complexity algorithm and could be significantly less complex than that of the SIR-Optimized Weighted LPIC [62], which requires the computation of cross correlation of each user's sequence and comes in order of $O(K^2N)$ floating point operations (FLOPs) for each stage and increasing exponentially for each additional stage. This complexity is equivalent to matrix inversion decorrelation multiuser detectors [29], which is still too high for the practical implementation. The complexity reduction techniques for a PIC such as Reduced PIC and Differential PIC proposed in [96] can be applied in straightforward manner to further reduce the complexity of the proposed Blind PIC. Another approach for the complexity reduction would be to implement block update of CM algorithm weights, where the weight vectors are updated once every block (several/tens of symbols)[97].

PIC Operations	**Conventional PIC**	**BA-PIC**
• Despreading • Error Calculation • Gradient Vector Estimation • Weight Vector Update	KN - - -	KN K KN KN
• Calculation of Scaling Factors	-	KN
• Respreading • Partial Summer • Interference Cancellation	KN KN KN	KN KN KN
Total Computations	~ 4KN	~ 7KN

Table 3.2: Number of FLOPs required per stage for PIC receivers

3.3.5 Theoretical and Simulation Results

A baseband synchronous uplink DS-CDMA system with K users and coherent BPSK modulation is assumed. Short binary Gold sequences [89] of length N = 31 are used for spreading users' data. The initial code acquisition scheme for a system using such sequences are investigated in [90] and can be used here without much difficulties. The channel is Rayleigh flat fading channel with normalized Doppler rate $f_d T_b$ of 0.003. A fixed step size of $\mu_k = 0.0001$, for all users is assumed in the CMA algorithm. The selection of step size in CDMA is generally based on the spreading factor used, the dynamic range of the received signal as described earlier and also effects the convergence of the algorithm [16, 59].

In Figure 3.15 , BER performance of 10 users with the proposed PIC at the output of initial (BA-PIC Stage 0) and first stage (BA-PIC Stage 1) obtained from the analysis using Gaussian Approximation is shown. For comparison, the BER simulation results under the same settings are also shown. As predicted, the analytical results closely match with the simulation results.

The BER performance results of the proposed blind receiver with $l = 0$ (BA-PIC Stage 0) and $l = 1$ (BA-PIC Stage 1) is compared with conventional PIC (Conventional PIC Stage 0 and Conventional PIC Stage 1) and shown in Figure 3.16. It is evident from the results that by employing CMA for adaptive despreading at the initial stage of the proposed PIC, much improved BER performance than the conventional PIC can be achieved. This result confirms our analysis, as the improvement in BER is directly related to the SINR improvement using our PIC approach. For 10 users, with an extra PIC stage, the BA-PIC showed a superior BER performance compared to the conventional PIC reaching very near to the single user BPSK signal bound. This is due to the improved accuracy of amplitude estimates for the interference cancellation of the users' signals of the BA-PIC.

In Figure 3.17, the error performances at different stages of the proposed BA-PIC and con-

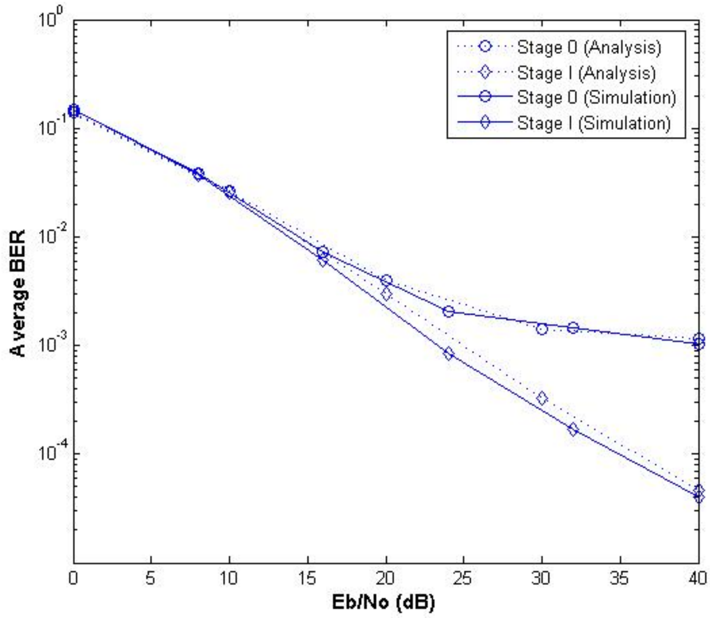


Figure 3.15: BER performance of BA-PIC in flat Rayleigh fading channel with K=10 and Gold sequences of N=31

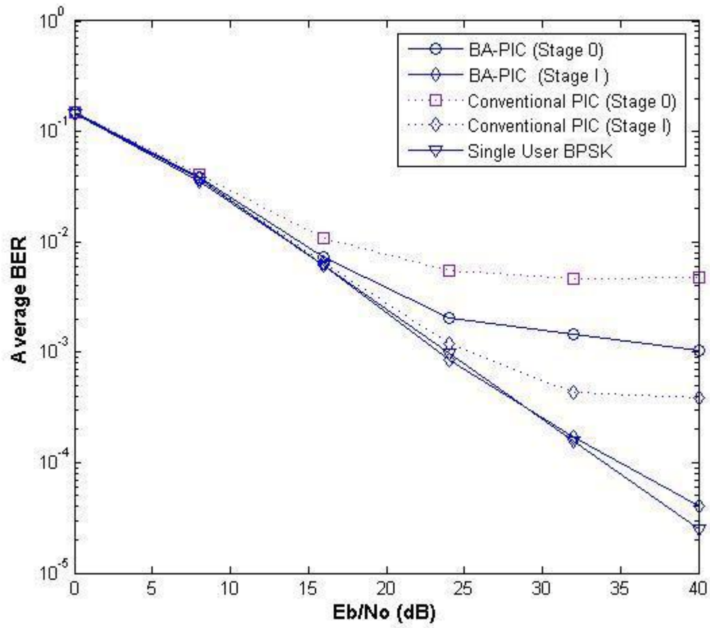


Figure 3.16: BER performance of BA-PIC in flat Rayleigh fading channel with K=10 and Gold sequences of N=31

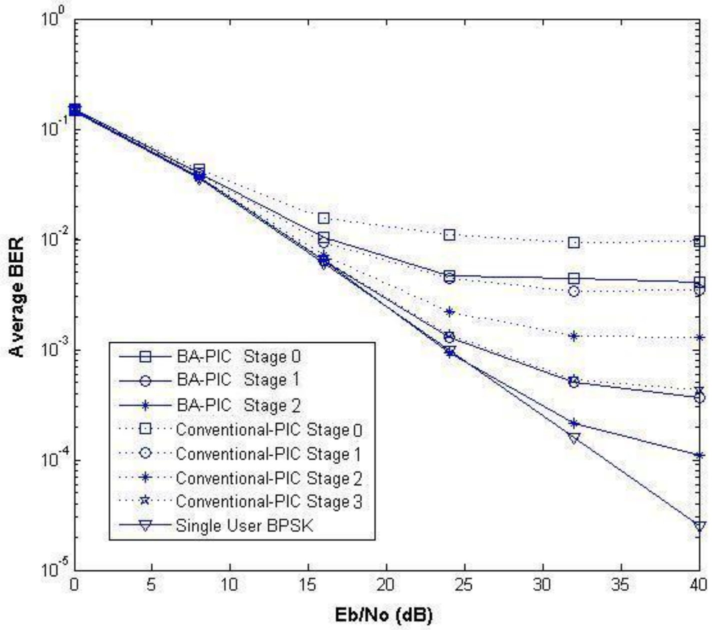


Figure 3.17: Performance of BA-PIC in flat Rayleigh fading channel with K=20 and Gold sequences of N=31

ventional PIC receivers in fading channels under higher system load of 20 users are shown. The performance of both receivers are shown to degrade compared to the 10 user system in 3.15. Therefore, the additional stages are employed to cope with the situation. The BA-PIC shows significant improvement in the performance with increase in number of PIC stages. As noted from the figure, the performance of the BA-PIC receiver at stage 1 has identical performance to that of conventional PIC with stage 3. The performance of the BA-PIC for stage 2 is shown to approach very near to that of a single user, thus giving impressive capacity advantage our approach. The BER performance obtained confirms our expectation and corresponds very well to the improvement in MSE of amplitude estimation plotted in Figures 3.18, 3.19 and 3.20.

In Figure 3.18, for the purpose of visualization, the signals at the output of the initial and first stage of PIC receivers for 20 users are shown. The MSE performance metric derived in (3.54) are used to quantify the accuracy of the estimation processes. The BA-PIC shows much improved MSE performance as compared to conventional PIC for both stages. The BER performance of PICs in Figure 3.17 are closely related to MSE of amplitude estimation. In Figure 3.19, relative MSE performance at the output of different stages of PICs is shown. As expected the MSE performance of the receivers improve with increase of number of PIC stages. For all stages, the BA-PIC shows significantly improved MSE compared to conventional PIC. How additional stage improve the MSE of amplitude estimation in BA-PIC is shown in Figure 3.20 and compared with conventional PIC for $K = 20$.

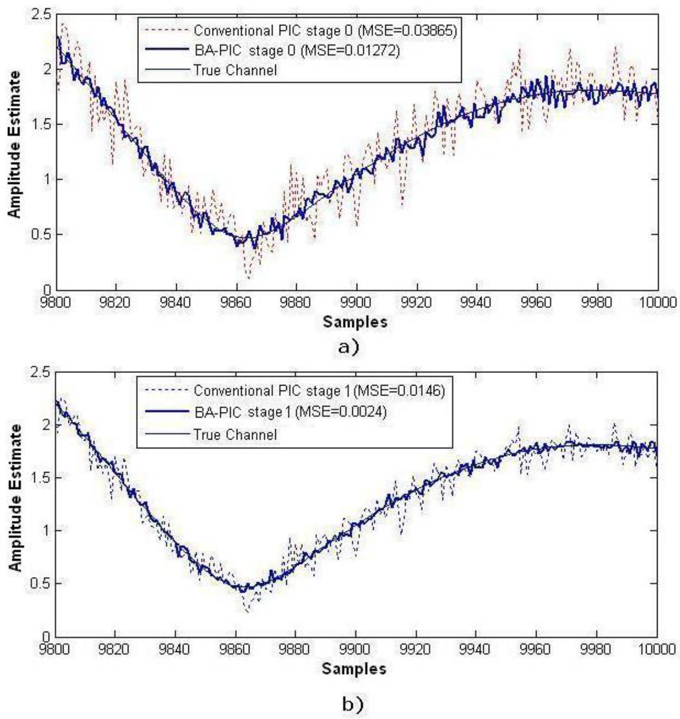


Figure 3.18: Amplitude estimates at the output of stage 0 and 1 of BA-PIC in flat Rayleigh fading channel for K=20 with $E_b/N_0 = 20$ dB, and Gold sequences of N=31

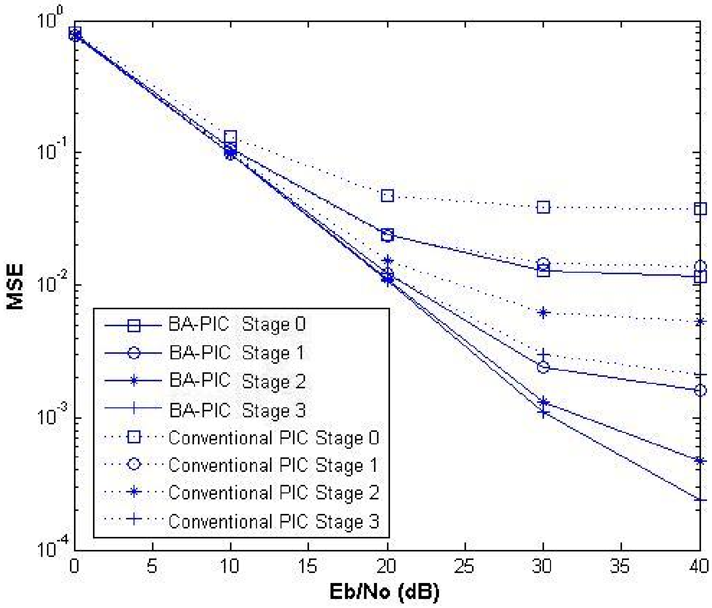


Figure 3.19: MSE of amplitude estimation vs. E_b/N_0 at the output of various stages of BA-PIC in flat Rayleigh fading channel for K=20 and Gold sequences of N=31

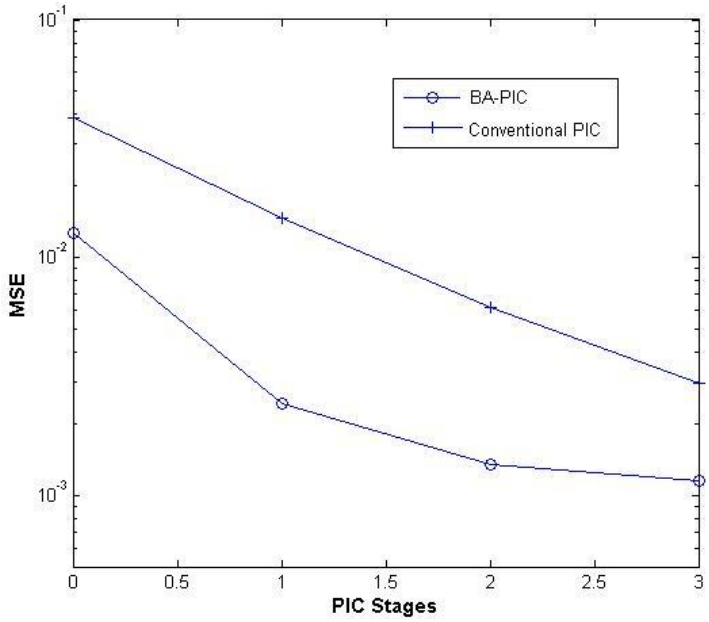


Figure 3.20: MSE of amplitude estimation at the output of various stages of BA-PIC in flat Rayleigh fading channel for K=20 with $E_b/N_0 = 30$ dB and Gold sequences of N=31

In Figure 3.21, the performance of the proposed PIC is shown at different stages for MAI cancellation while the variance of noise approaching towards zero. More precisely, all the users' signals have equal average $E_b/N_0 = 30dB$. BER performances of single user BPSK and conventional MF receiver under the same fading conditions are also shown for reference. It is observed that the initial stage of the proposed BA-PIC removes considerable amount of MAI under all user loading conditions. As predicted, most of the performance gain of the PIC comes from the first stage of interference cancellation and the results show near single user performance for large number of users. This indicates that our PIC approach indeed performs very accurately and removes significant amount of MAI from the system. The error performance is further improved by adding more stages as expected.

In Figure 3.22, the BER performance of BA-PIC in Rayleigh fading and equal power users with $K = 10$ is compared with PIC with partial cancellation factors [52] or the recently proposed Weighted Linear PIC (WLPIC) as in [62] where the cancellation weight is varied from 0 (which also corresponds to a Conventional MF) to 2.6. It can be clearly seen that the proposed BA-PIC achieves the best performance among the schemes considered. The optimum weight for the considered system using a WLPIC is shown to be 1.4 under which BER performances of the BA-PIC and the WLPIC are identical.

Figure 3.23 shows the performance of the proposed BA-PIC and conventional PIC with 20 users in fading channels and under different nearfar ratio $\Omega = \max\{P_i\}/P_k = 0{-}20dB, \forall i, i \neq k$

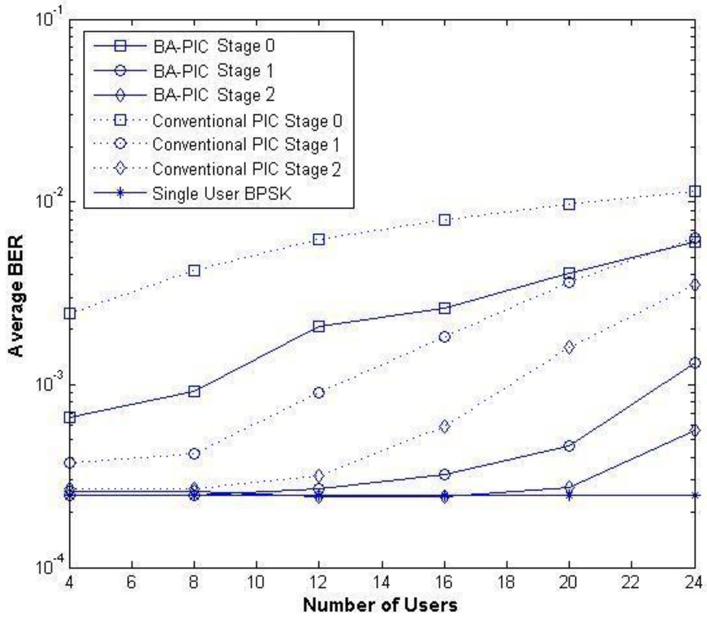


Figure 3.21: Performance of BA-PIC in flat Rayleigh fading channels with $E_b/N_0 = 30$ dB and Gold sequences of N=31

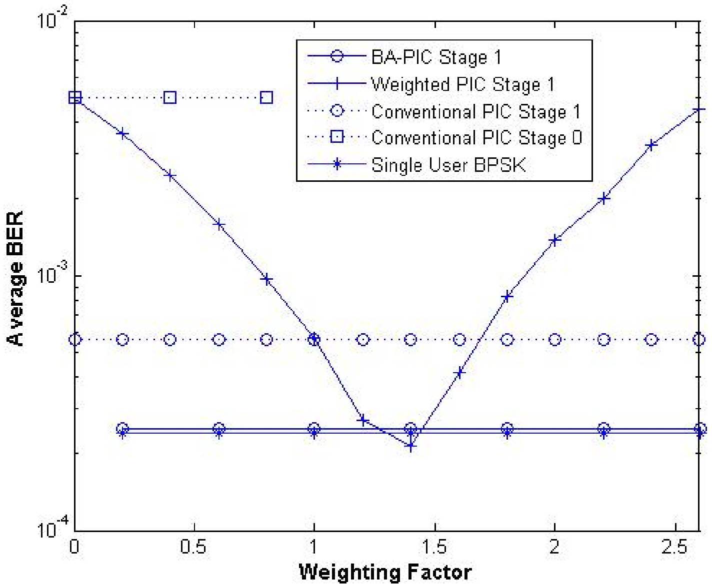


Figure 3.22: Performance of BA-PIC in flat Rayleigh fading channels for K=10, $E_b/N_0 = 30$ dB and Gold sequences of N=31

conditions. The received signal power of desired user k is assumed unity $0dB$ with $E_b/N_0 = 20dB$. The power of all the interferers i are assumed uniformly distributed between $0dB$ to ΩdB. Here, the step size $\mu_k = 0.0001/\Omega$ is assumed in the CMA algorithm in all stages. It can be clearly seen from the figure that the BA-PIC shows more robust performance as compared to conventional PIC under the whole range of nearfar ratio conditions. It is noted that stage 1 of the proposed PIC can offer near single user performance under the nearfar ratio as high as 15 dB. With conventional PIC, even with stage 2, the performance rapidly degrades as the nearfar ratio increases, performing far worse than the proposed BA-PIC at the stage 1. This significant result of the proposed PIC can be attributed to the adaptive design approach used. At each stage, the cancellation of MAI contributions takes into account the power variations of desired and interfering users' signals on individual basis and corresponding amplitude estimates are selected to reflect on their strength.

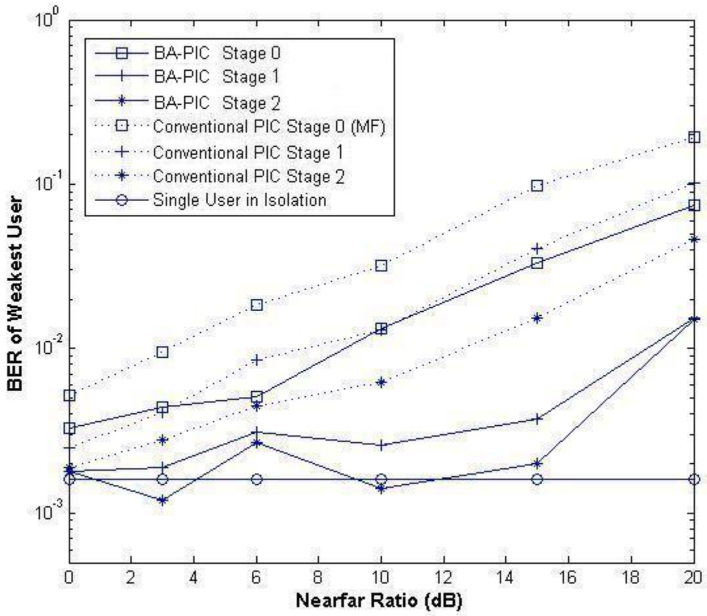


Figure 3.23: Performance of BA-PIC in flat Rayleigh fading channels and near far conditions for K=20 with E_b/N_0 of the weakest user=20dB and Gold sequences of N=31

The user capacity performance of BA-PIC is also presented in channel environments with AWGN only, but with severe nearfar problems. Such environments may occur when the users are slowly moving and are under direct line of sight from the base-station. The received signal power of desired user is assumed to be unity $0dB$. The signal power of all the other users are assumed to be uniformly distributed between 0-20dB. In Figure 3.24, BER performance of BA-PIC under the above described scenario of users' signal power is shown. The effect of nearfar conditions caused by the presence of interfering high power users to the desired weak user with effective $E_b/N_0 = 7dB$ is observed. The number of users in the system is varied from 5 to 25.

The BA-PIC shows improved performance compared to conventional PIC at the stage 0. At the stage 1, the BA-PIC shows rapid performance improvement as we expected. The additional stages improve the performance even further; but as can be seen, the gain is not as high as in the stage 1. The conventional PIC shows good performance under low system load of up to 12 users. As the load increases the performance degrades dramatically and approaches that of conventional MF receivers.

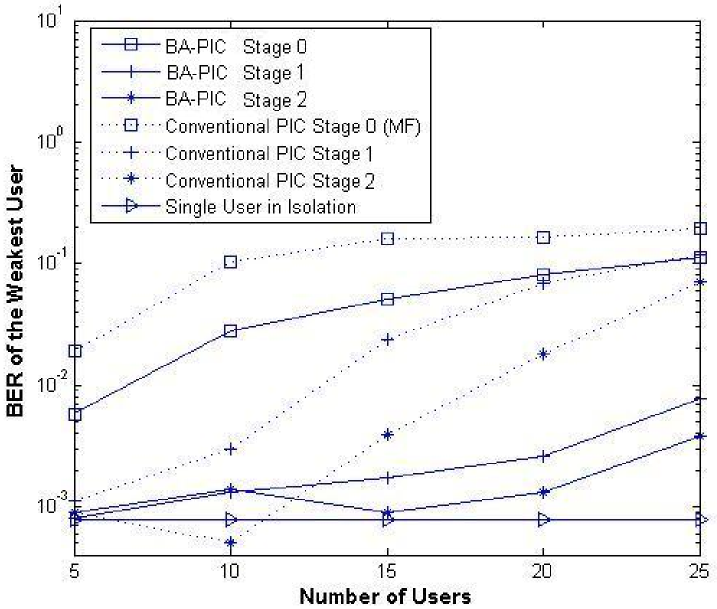


Figure 3.24: Performance of BA-PIC in AWGN channel with near far ratio of 20 dB with E_b/N_0 of the weakest user=7dB and Gold sequences of N=31

Figure 3.25 shows the performance of the proposed BA-PIC and conventional PIC with 20 users in AWGN channel environment and under different nearfar ratio $\Omega = \max\{P_i\}/P_k = \in [0, 20]dB, \forall i, i \neq k$ conditions. The received signal power of the desired k^{th} user is assumed unity $0dB$ and the signals power of interferers i are assumed uniformly distributed between $0dB$ to ΩdB. Here, the step size $\mu_k = \frac{0.0001}{10^{\Omega/10}}$ is assumed in the CMA algorithm in all stages similar to the case of fading channels. The signal to noise ratio of desired (weakest) user is $E_b/N_0 = 7dB$. It can be clearly seen from the figure that BA-PIC offers more robust performance compared to the conventional PIC under all range of nearfar ratio conditions. It is noted that the proposed PIC at stage 1 can offer near single user performance under all the nearfar ratio conditions. This is a more improved performance of the BA-PIC compared to that in fading channels as shown in Figure 3.23. Because the AWGN only environment provides enough time duration for the convergence of weights of CMA algorithm to optimal values $\mathbf{w}_k^{opt}$. As expected, it is observed that amplitude estimates generated are very accurate and thus the interference cancellation process of the BA-PIC shows resilience under all nearfar ratio conditions considered. With conventional PIC, even at the output of stage 2, the performance rapidly degrades as the nearfar ratio increases and performs far worse than the proposed BA-PIC with stage 1.

The performance of BA-PIC as well as conventional PIC operating in a two path frequency selective channels with average power of 0.8 and 0.2, respectively and AWGN is shown in Figure 3.26. The receivers perform interference cancellation on each path and RAKE combining is used at each stage to obtain decision variables. As can be seen from the figure, the BA-PIC shows improved BER. The BER improvement with additional PIC stage is found to be moderate here; this may be caused due to imperfect interference estimation within PIC processes.

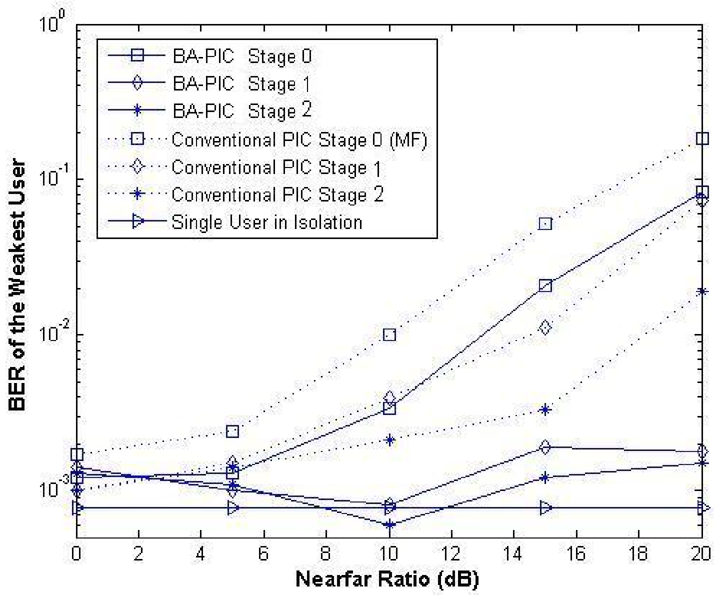


Figure 3.25: Performance of BA-PIC in AWGN channel with near far conditions for K=20 with E_b/N_0 of the weakest user=7dB and Gold sequences of N=31

The spectral efficiency of BA-PIC for multiple stages under Rayleigh fading channels and $K = 20$ users are shown in Figure 3.27. It can be clearly seen from the figure that the BA-PIC offers notable increase in achievable spectral efficiency compared with conventional PIC. Actually, even with a single stage of cancellation BA-PIC shows performance comparable with three stage of cancellation with the conventional PIC. Also, the achievable spectral efficiency comparison of two receivers (CMA-SIC and BA-PIC) proposed in this thesis are presented in Figure 3.28. It can be seen from the figure that BA-PIC offers much higher spectral efficiency compared with CMA-SIC as the number of cancellation stage increases. Even with single stage of cancellation BA-PIC shows much higher information rate than that of CMA-SIC.

The effect of different step-size $\mu = 0.00001 - 0.001$ on the BER of BA-PIC in Rayleigh flat fading channels is investigated in Figure 3.29. The system loading of $K = 10$ and 20 with equal power $P_k = 1, \forall k$ are used and operating of $E_b/N_0 = 20$ dB is selected. It can be seen that the

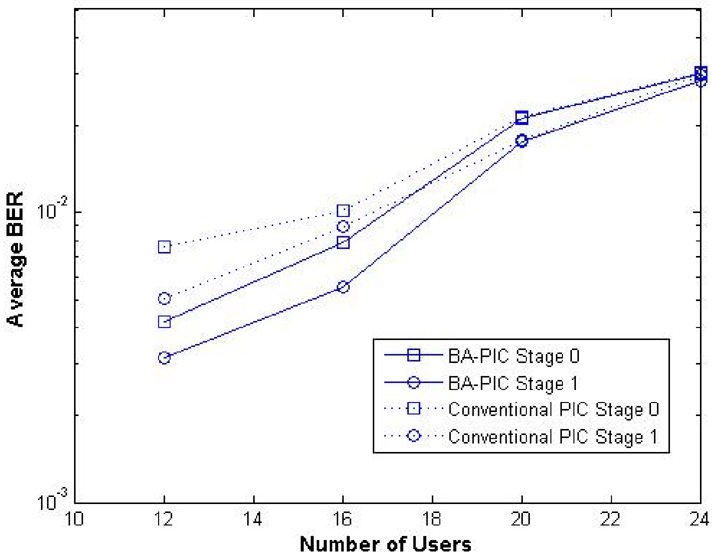


Figure 3.26: Performance of BA-PIC in frequency selective multipath channels with AWGN, $E_b/N_0 = 15$ dB and Gold sequences of N=31

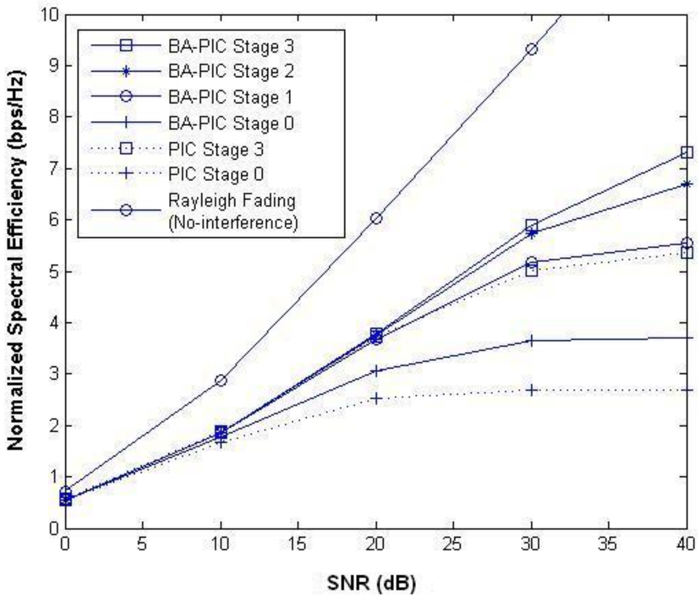


Figure 3.27: Spectral efficiency of BA-PIC in Rayleigh flat fading channels for K=20 and Gold sequences of N=31

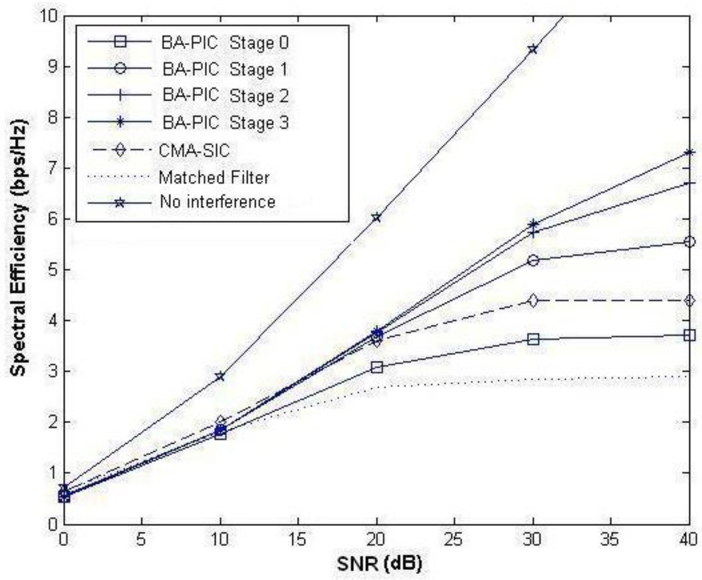


Figure 3.28: Spectral efficiency comparison of BA-PIC and CMA-SIC in Rayleigh flat fading channels for K=20 and Gold sequences of N=31

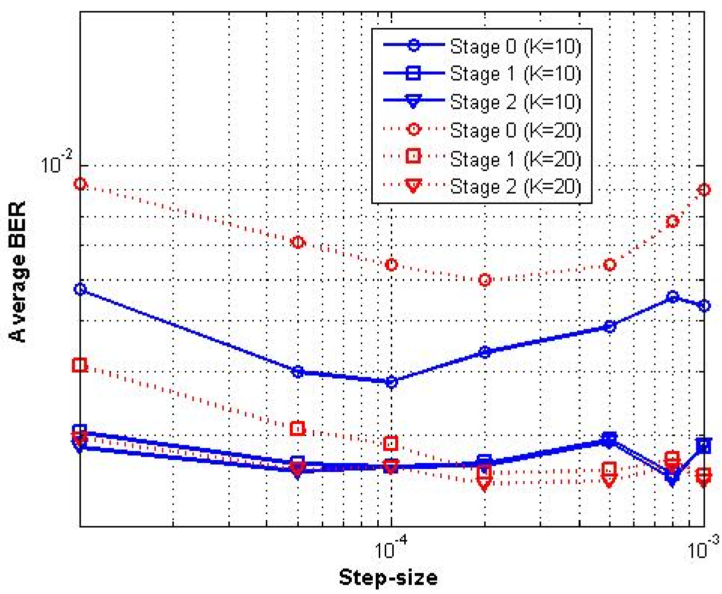


Figure 3.29: Effect of step-size on the performance of BA-PIC in flat Rayleigh fading channel for K=10 and 20, and Gold sequences of N=31

performance of BA-PIC is effected by the step-size used at different PIC stages. Also noted that the $\mu = 0.0001$ used in previous simulations is not the optimum step-size and there is still room for further performance improvement. The gap in performance due to system loading of $K = 10$ and 20 at the initial stage (Stage 0) becomes narrower with added PIC stages.

From the analysis and range of simulation results, it can be concluded that the blind adaptive approach of the BA-PIC is a very effective way of improving the performance of uplink CDMA. Besides the improved performance, it also greatly relaxes the requirements on accurate power control.

3.4 Blind Adaptive Subcarrier Combining Technique for MC-CDMA Receiver in Mobile Rayleigh Channel

3.4.1 Literature Review of MC-CDMA Techniques

The CDMA transmissions often undergo multipath fading due to the wideband nature of users' signals. Conventional approach with single carrier CDMA transmission is to use RAKE receiver to favorably combine multipath signals to improve the frequency diversity. The multipath signals can however create more interference than the diversity gain, if they are not exploited thoroughly. Alternatively, one can use some form of equalization scheme to mitigate the interference effect due to such channels. There are different non-blind and blind equalization techniques available in the literature such as MMSE, ZF [5] and CMA [66]. However the single carrier based approaches have many shortcomings such as severe loss in SINR as the number of multipath increases, high complexity of equalization algorithm etc..

MC-CDMA is an effective multiple access scheme for the multipath fading and frequency selective channels and is currently being considered for the future mobile wireless systems. The main idea behind the MC-CDMA is to allow the transmissions to be carried out over a large number of subcarriers such that each of them experiences frequency flat fading to avoid the multipath effects and [69], [70], [71]. MC-CDMA compared with its single carrier based counterpart, achieves higher performance with simple techniques such EGC and MRC combining receivers. However they are severely effected by MAI, leading to poor detection performance as system loading increases. More advanced techniques incorporating multiuser detection, or for example, multistage RLS subcarrier-combining scheme recently proposed in [98] could be used to improve the performance. However, they are more computationally complex. A different approach is proposed in [99], that uses transmit power adaptations in time and frequency domain for MRC to reduce the degrading effects caused by weak subcarriers. However, it requires closed loop operation and very accurate channel estimation. In [100], MMSE filtering based scheme is proposed for operation in mobile Rayleigh channel for Multicarrier DS-CDMA. Although it shows much improved performance, however it requires tracking of the users' fading processes for inversion of signal correlation matrix resulting in a more complex system. In [101], a blind adaptive technique for Multicarrier DS-CDMA is proposed using a constrained constant modulus inverse QR decomposition (IQRD)-RLS algorithm for static channel conditions. The scheme is shown to outperform MRC and LMS type schemes, however, at the cost of significant increase in the complexity. In this Letter, a low complexity subcarrier combining technique is proposed for frequency-domain spread MC-CDMA that also uses the constant modulus algorithm as in [101], which is a well known algorithm in the literature. However, it is designed here in a different way that suppresses MAI with additional channel tracking capability inherent in the detection process for operation in mobile Rayleigh fading channel. More specifically, we propose an improved alternative to conventional EGC and MRC methods with similar order of complexity and without need for closed loop power adaptation strategies as in [99].

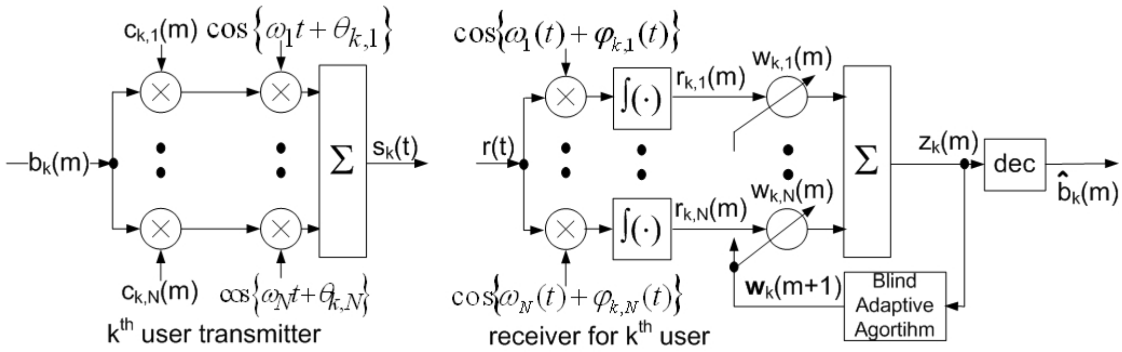


Figure 3.30: Proposed MC-CDMA system model

3.4.2 System Model

An uplink synchronous MC-CDMA system of K users is considered. The transmitted signal of k^{th} user $s_k(t)$ as shown in Figure 3.30, can be written as:

$$s_k(t)=\sqrt{P_k}\sum_{m=-\infty}^{\infty}\sum_{n=1}^{N} b_k(m)p(t-mT_b)c_{k,n}(m)\cos\{\omega_n t+\theta_{k,n}(t)\}, \tag{3.66}$$

where P_k is the signal power, $b_k(t)$ is the data signal, where $b_k(m)$ is a binary sequence taking values $[-1,+1]$ with equal probabilities, $p(t)$ is rectangular pulse with period T_b. The spreading sequence is denoted as $c_{k,n}$ of rectangular pulse shaping function $p(t)$ with period T_c and with normalized power over a symbol period equal to unity $|c_{k,n}(t)|^2 = 1$. The spreading factor is $N = T_b/T_c$. The spreading chip elements $c_{k,n}(m)$ are modulated using separate subcarriers with frequency ω_n radians per second and satisfying the orthogonality as described in [72], $\theta_{k,n}$ is the phase uniformly distributed between $[0, 2\pi]$.

It is assumed that transmit channels of each user undergo flat fading on each subcarrier, given by

$$\beta_{k,n}(t) = \alpha_{k,n}(t)e^{j\pi\phi_{k,n}(t)}, \tag{3.67}$$

which is a zero mean and unit variance complex Gaussian random variable with $\alpha_{k,n}(t)$ is the amplitude and $\phi_{k,n}(t)$ is the phase component. For the purpose of simplicity in analysis, it is assumed here that fading on each subcarrier is independent and identically distributed although these assumptions are very difficult to be fulfilled in practice [72]. The correlated of channels does not however significantly impact our system design approach and the benefit and can naturally be extended to operate in such condition.

The received signal is the sum of K users' transmitted signals given by:

$$r(t)=\sum_{m=-\infty}^{\infty}\sum_{k=1}^{K}\sqrt{P_k}\sum_{n=1}^{N}\alpha_{k,n}(t)b_k(m)p(t-mT_b)c_{k,n}(m)\cos\{\omega_n t+\psi_{k,n}(t)\}, \tag{3.68}$$

where $\psi_{k,n}(t) = \theta_{k,n}(t) + \phi_{k,n}(t)$ is the received phase for k^{th} user signal, $\eta(t)$ is the AWGN with two sided power spectral density $N_0/2$

Assuming coherent detection and perfect phase knowledge of users' subcarrier and fading channel phases, the received signal is first subcarrier demodulated followed by chip matched filtering and integrating over the m^{th} symbol period to give an output variable $r_{k,n}(m)$ as follows:

$$r_{k,n}(m) = \int_{(m-1)T_b}^{mT_b} r(t)\cos\{\omega_n t+\varphi_{k,n}(t)\}. \tag{3.69}$$

The signal $r_{k,n}(m)$ is then weighted by the proposed algorithm to give an output variable $z_{k,n}(m)$ as:

$$z_{k,n}(m) = r_{k,n}(m)w_{k,n}(m), \tag{3.70}$$

where $w_{k,n}(m)$ is a weighting factor that is dependent on different method of subcarrier combining for obtaining the final decision variable $z_k(m)$ for estimating the user's transmitted symbols.

3.4.3 Conventional Subcarrier Combining Techniques for MC-CDMA

There are two main methods for subcarrier combining in an uplink MC-CDMA receiver. The technical issues of both methods and comparisons their performance are studied extensively in [69], [70], [71].

Equal Gain Combining This combining method uses the same weight i.e. 1 for all subcarriers. This is obtained simply by

$$w_{k,n}(m) = 1c_{k,n}(m). \tag{3.71}$$

This method does not require users' amplitude estimation and hence is a simple and practical method for a MC-CDMA receiver.

Maximum Ratio Combining With MRC, the combining of subcarrier is done based on the amplitude estimate of each subcarrier signal i.e. $\alpha_{k,n}$. This method assigns weight for each subcarrier for the combining as follows:

$$w_{k,n}(m) = \hat{\alpha}_{k,n}(m)c_{k,n}(m). \tag{3.72}$$

The MRC method is known to maximize the SNR under single user transmission. Since multiuser signals on each subcarrier of MC-CDMA inevitably comes with MAI, this method does not retain the same SNR performance and performs even worse than EGC as will be shown later in the performance results Section.

Finally, the estimate of the user's data is obtained by performing decision (for BPSK user data, this is simply the sign operation) on the combined output signal as follows

$$\hat{b}_k(m) = dec\left\{ \sum_{n=1}^{N} z_{k,n}(m) \right\}. \tag{3.73}$$

3.4.4 Proposed Blind Adaptive Subcarrier Combining (BASC) Technique

The EGC and MRC are linear combining methods which do not utilize the correlation between the magnitudes of current and previous output decision variables for the calculation of combining weights. Furthermore, they do not exploit the knowledge of fixed modulus property of transmitted data signals of users. The technique proposed here utilizes these information and follows a blind and adaptive approach for the subcarrier combining in MC-CDMA receiver by attempting to achieve the set minimum error cost function.

The proposed receiver structure is shown in Figure 3.30. The receiver uses a blind adaptive algorithm to maintain the magnitude of output signal $z_k(m)$ equal to the expected magnitude which is calculated as $\gamma = E\left\{ \sum_{n=1}^{N} |b_k c_{k,n}|\alpha_{k,n} \right\}$ and generates an error when their values differ. Since the channels vary insignificantly over a symbol period, it utilizes $z_k(m)$ and the error signal to correct weight for each subcarrier for the next symbol period, hence, attempts instantaneously to reduce the probability of generating wrong decision variable.

The BASC algorithm: In the first symbol period $m = 1$, the weights $w_{k,n}(m)$ are is initialised with elements of user's spreading sequence $w_{k,n}(m) = c_{k,n}(m)$, $\forall k$, $\forall n$. Next, a step-size μ of the adaptive algorithm is selected. For $m = 1, 2, 3, ..$ the weighting of each subcarrier signal $r_{k,n}(m)$ is carried out to obtain output variables $z_{k,n}(m)$ as follows:

$$z_{k,n}(m) = r_{k,n}(m) w_{k,n}(m). \tag{3.74}$$

The decision variable is obtained by combining the output variables $z_k(m) = \sum_{n=1}^{N} z_{k,n}(m)$ and decision is made to obtain user data $\hat{b}_k(m) = dec\{z_k(m)\}$. Using $z_k(m)$ and γ, the minimum error cost function is evaluated:

$$\min \left[e_k(m)^2 = \left\{ z_k(m) \{ z_k(m)^2 - \gamma \} \right\}^2 \right]. \tag{3.75}$$

It can be observed that when $e_k(m)^2 \longrightarrow 0$, our scheme approaches to the MMSE performance. Note that explicit channel tracking as in [100] is not required here. Using $e_k(m)$ and $r_{k,n}(m)$, error gradient signals at the m^{th} symbol period are calculated:

$$\nabla_{k,n}(m) = e_k(m) r_{k,n}(m). \tag{3.76}$$

Subcarrier weights for the next symbol $w_{k,n}(m+1)$ are updated using $\nabla_{k,n}(m)$ as follows

$$w_{k,n}(m+1) = w_{k,n}(m) - \mu\nabla_{k,n}(m). \tag{3.77}$$

where $\mu > 0$ is the step-size. It has to be noted that, unlike in [101] where weights are used for MAI suppression only, while assuming the channels to be static, here each weight $w_{k,n}(m)$ in (3.77) is updated with dual purpose of suppressing MAI as well as tracking of instantaneous channel amplitude $\alpha_{k,n}(m)$. The accuracy of tracking and detection performance is dependent on the appropriate selection of μ value and can be optimized according to the channel Doppler rate and system loading. The above processes are repeated for all K users. The conventional EGC method can be considered as the special case with $w_{k,n}(m) = c_{k,n}(m), \forall m$. As can be seen, the proposed algorithm has complexity in order of $\mathcal{O}(N)$ comparable to EGC and MRC. This approach is motivated by two simple ideas:

1. Using the knowledge of i.i.d. distribution of channel magnitude of each subcarrier, the LLN theory predicting the long term stability in convergence of combined signal to the expected sum magnitude can be utilized. This ensures combining weights are updated more and more reliably at every symbol period to minimize the error between the expected magnitude and sum of weight updated signals

2. Since the channels of users on each subcarrier experience slow and flat fading, the time correlation to the channels can be exploited to select an appropriate step size in the blind adaptive algorithm for updating the subcarrier weights for optimum performance.

In the Table 3.3 below, the BASC algorithm is summarized.

Table 3.3: The proposed BASC algorithm steps

	a) At $m = 1$, initialize $w_{k,n}(m) = c_{k,n}(m)$, $\forall k$, $\forall n$ b) Set the step size μ, $\gamma = E\Big\{ \sum_{n=1}^{N} \lvert b_k c_{k,n} \rvert \alpha_{k,n} \Big\}$ c) For $m = 1, 2, 3, ..$, perform steps $1 - 7$
1	Perform weighting, $z_{k,n}(m) = r_{k,n}(m)w_{k,n}(m)$
2	Combine signals, $z_k(m) = \sum_{n=1}^{N} z_{k,n}(m)$
3	Obtain data estimate, $\hat{b}_k(m) = dec\big\{z_k(m)\big\}$
4	Evaluate, min $\Big[e_k(m)^2 = \big\{z_k(m)\{z_k(m)^2 - \gamma\}\big\}^2\Big]$
5	Calculate gradients, $\nabla_{k,n}(m) = e_k(m)r_{k,n}(m)$
6	Update weights, $w_{k,n}(m+1) = w_{k,n}(m) - \mu\nabla_{k,n}(m)$
7	Stop after the detection of all K user

3.4.5 Performance Results

An uplink synchronous MC-CDMA system with K users with equal number of subcarriers and spreading length of $N = 32$ is considered. Note however that the number of subcarriers need not be equal to N and the equality is assumed here for simplicity in performance evaluation only. The system is simulated in MATLAB using 10000 BPSK data symbols for each user with i.i.d. Rayleigh flat fading on each subcarrier with normalised Doppler rates of $f_d T_b = 0.003$ and 0.01, respectively. For the evaluation of MRC receivers, perfect channel estimation is assumed (although unrealistic this is assumed here for comparison purpose only). Two types of spreading sequences with $N = 32$ are used: orthogonal Walsh sequences and Gold sequences [89]. The Gold sequences are obtained from two fifth order polynomials with octal values 45 and 75, giving the sequence length of 31 and single random bits are appended to make the sequence length $N = 32$ for the fair performance comparison with Walsh sequences.

BER vs. E_b/N_0

The BER performance of the BASC technique is compared with EGC and MRC under the same system loading of $K = 10$ users is shown in Figure 3.31 for Walsh sequences. BASC with $\mu = 0.003$ is selected for reasonable adaptivity. As expected, it shows significantly improved BER compared with EGC and MRC. MRC shows degraded BER than EGC under Walsh sequences and approaches EGC in the case of Gold sequences. The MRC receiver shows BER degradation compared with EGC [71] and the proposed technique. It can be noted that the BASC receiver restores some of the orthogonality of Walsh sequences used and hence achieves improved BER. Similarly BER performance using Gold sequences are shown in Figure 3.32. BASC continue to show improved BER compared with EGC and MRC while the performance of EGC approaches that of MRC. All schemes show less performance with Gold sequences due to higher cross-correlation. In higher mobility scenario with $f_d T_b = 0.01$, BASC shows slightly less performance gain, due to insufficient channel tracking. In Figure 3.33, the BER of three techniques are compared under the same system load of $K = 16$ using Gold sequences and compared with Walsh sequences and $f_d T_b = 0.003$. The proposed receiver shows BER much improved compared with EGC and MRC. For example at the $E_b/N_0 = 15dB$, the BER of 0.0065 is achieved compared with 0.023 and 0.031 for the other receivers. The use of Gold sequences shows slight degradation compared with Walsh sequences at higher SNR region i.e. $E_b/N_0 > 15dB$. This result provides some insights into operation of the BASC under asynchronous environments where non-orthogonal sequences such as Gold or PN sequences are normally used.

User Capacity

The user capacity gain of the proposed receiver using Walsh sequences is presented in Fig-

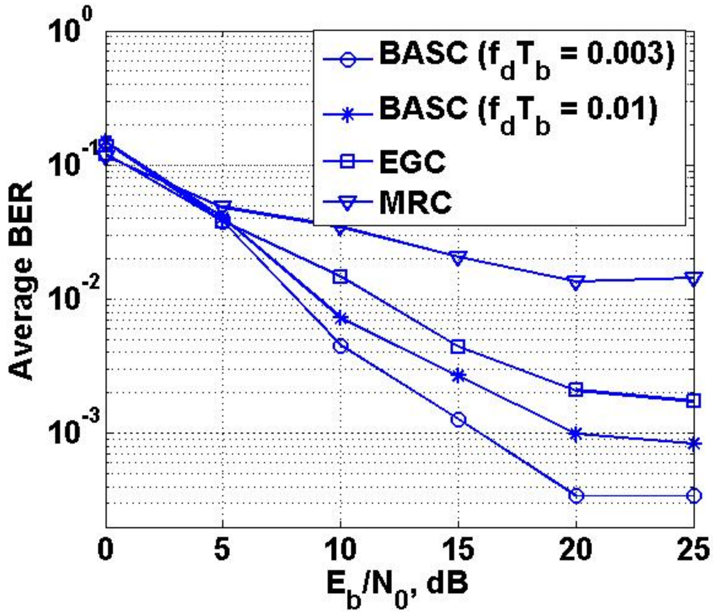


Figure 3.31: BER vs. E_b/N_0 performance of Blind Adaptive Subcarrier Combining for uplink MC-CDMA with K=10, $\mu = 0.003$, and Walsh sequences of N=32

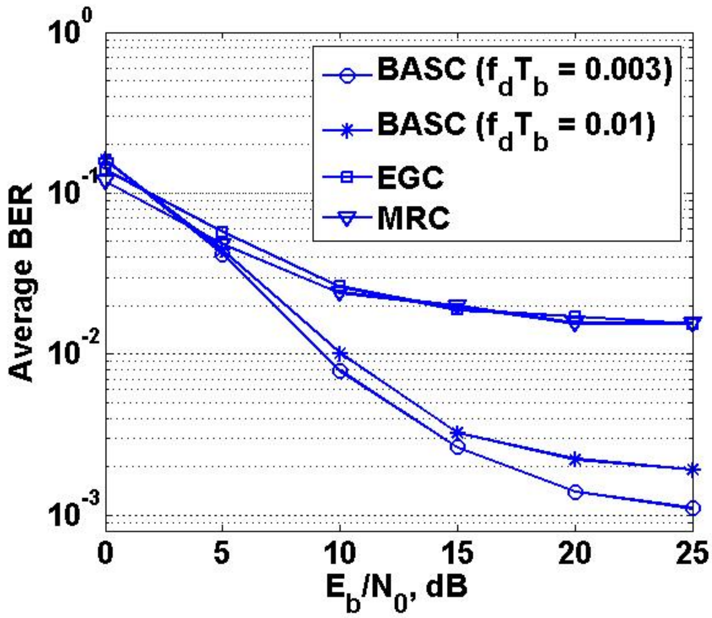


Figure 3.32: BER vs. E_b/N_0 performance of Blind Adaptive Subcarrier Combining for uplink MC-CDMA with K=10, $\mu = 0.003$, and Gold sequences of N=32

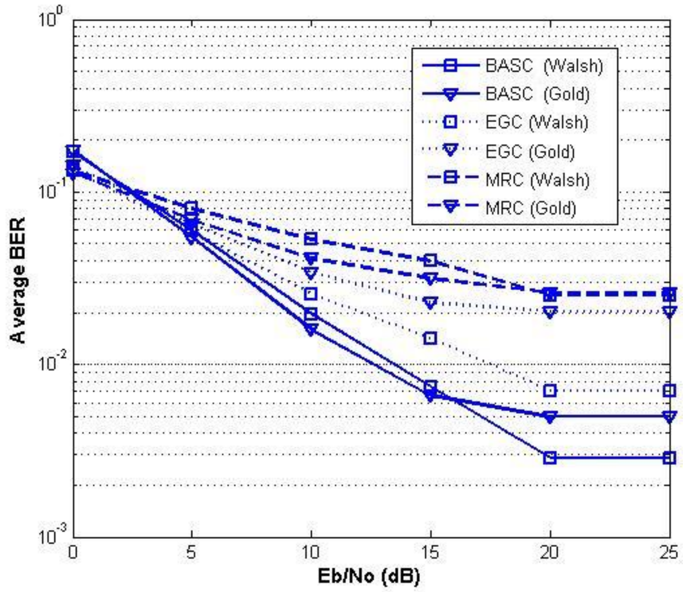


Figure 3.33: BER vs. E_b/N_0 performance of Blind Adaptive Subcarrier Combining technique for uplink MC-CDMA for K=16 , $\mu = 0.003$; Walsh (squares) and Gold (triangles) sequences of N=32

ure 3.34. It can be clearly seen from the figure that the BASC receiver shows much higher user capacity compared with other techniques. For example, at the same BER of 0.01 the proposed technique can support 19 users compared with 15 of EGC and less than 8 users with MRC. Similarly the techniques are compared with the use of Gold sequences under the same settings in Figure 3.35. The BASC receiver shows significantly higher user capacity under this system settings. For example, at the same BER of 0.01 the BASC receiver can support 17 users compared with 4 of EGC and 7 users with MRC, confirming the significant gain of the proposed blind adaptive design approach. Interestingly MRC approach shows improved performance under low system loading compared with EGC and gradually degrades as loading increases ($K > 8$). Again, in higher mobility scenario with $f_d T_b = 0.01$, BASC supports slightly less users, due to insufficient channel tracking.

The effects of step-size

More interesting aspects of the proposed technique is now investigated. The effect of step-size on the performance of the proposed blind adaptive receiver under system loading of $K = 10$ and 16 under the same $E_b/N_0 = 20$ dB are presented. In Figure 3.36 the BER results of the proposed blind adaptive receiver is investigated under the range of step-sizes $\mu = 0.0003 - 0.04$ using Walsh sequences. It can be clearly seen from the figure that the proposed technique provides significant BER improvement under all step-size used. The vertical line with stars presents BER under the step size $\mu = 0.003$ used in earlier simulations. Interestingly, the optimum step-size for $K = 10$ and 16 are shown to be $\mu = 0.025$ and 0.02, respectively. With the use of optimum step-size for

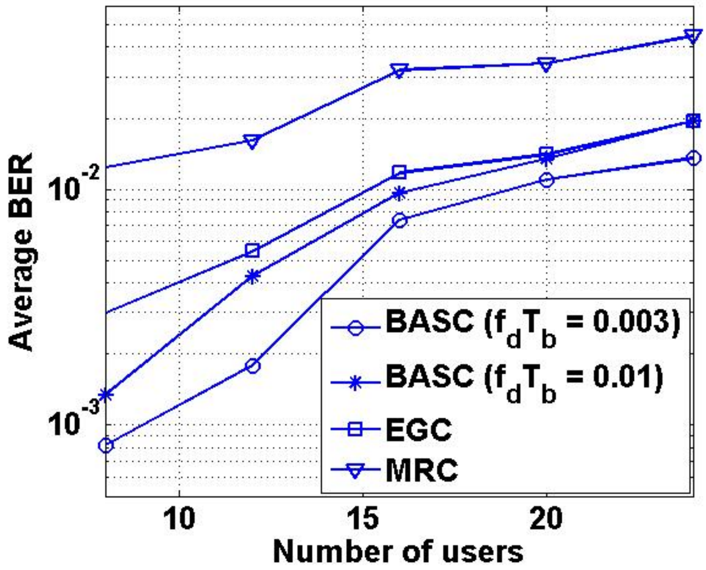


Figure 3.34: BER vs. number of users performance of Blind Adaptive Subcarrier Combining for uplink MC-CDMA, $E_b/N_0 = 15$dB, $\mu = 0.003$, and Walsh sequences of N=32

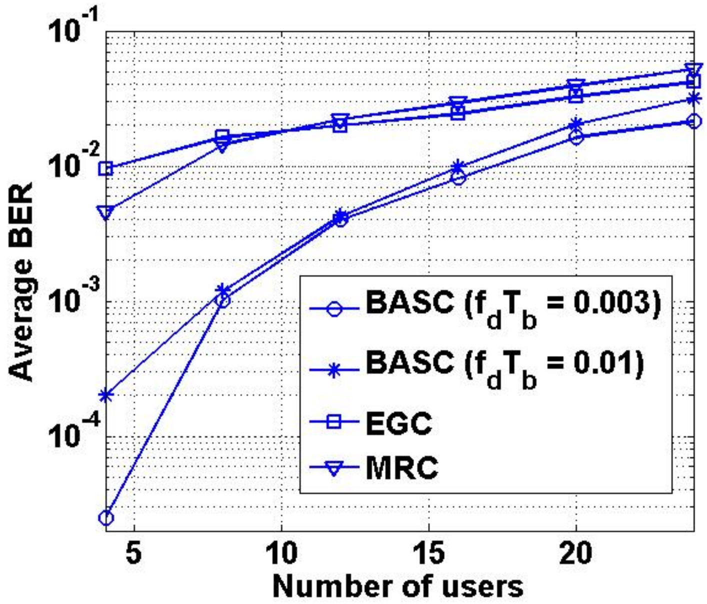


Figure 3.35: BER vs. number of users performance of Blind Adaptive Subcarrier Combining technique for uplink MC-CDMA, $E_b/N_0 = 15$dB, $\mu = 0.003$, and Gold sequences of N=32

$K = 10$, significant improvement in BER is shown to be achievable with the proposed receiver compared with EGC and MRC. Similar BER performance improvement is also seen for the case of $K = 16$.

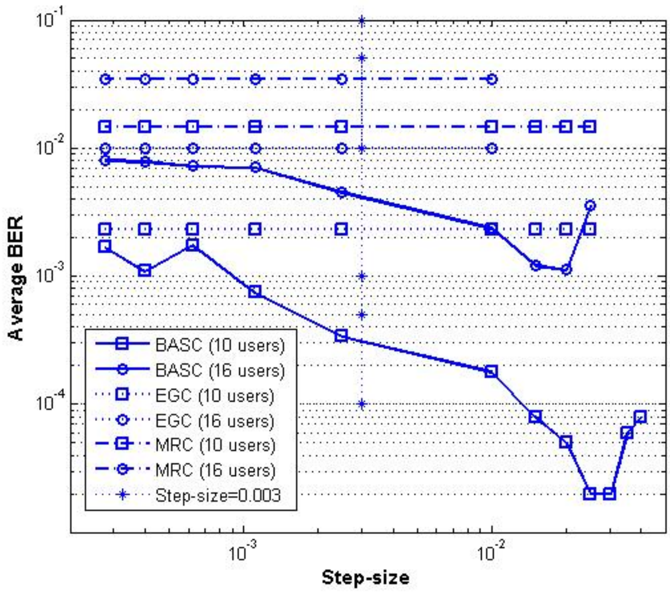


Figure 3.36: Effect of step-size μ on the BER performance of Blind Adaptive Subcarrier Combining for uplink MC-CDMA with K=10 and 16, $E_b/N_0 = 20$dB, and Walsh sequences of N=32

In Figure 3.37 the BER results of the proposed blind adaptive receiver with Gold sequences is investigated under the range of step-sizes $\mu = 0.0003 - 0.04$. The proposed technique provides significant BER improvement with most step-sizes used. The vertical line with stars presents BER under the step size $\mu = 0.003$ used in earlier simulations. The optimum step-size for $K = 10$ and 16 are shown to be $\mu = 0.015$ and 0.01, respectively. For example, with the use of optimum step-size for $K = 10$ it achieves significant BER improvement compared with EGC and MRC giving 0.0004. Similar gains are also shown for $K = 16$.

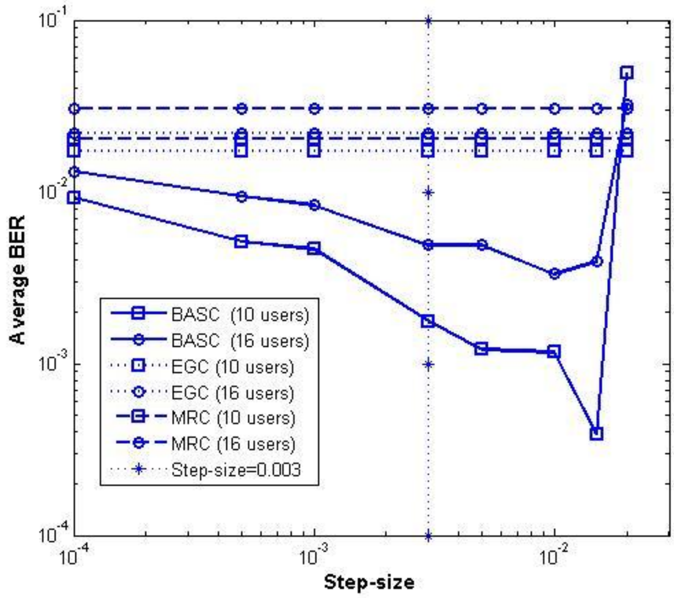


Figure 3.37: Effect of step-size μ on the BER performance of Blind Adaptive Subcarrier Combining for uplink MC-CDMA with K=10, $E_b/N_0 = 20$dB, and Gold sequences of N=32

3.5 Chapter Summary

This Chapter is focused on maximizing the number of users in uplink CDMA by designing more advanced receiver techniques and algorithms. In Section 3.2, an improved CMA-SIC receiver is proposed for DS-CDMA, using CM algorithm embedded in each SIC stage to perform the user's amplitude estimation for the detection and cancellation. It showed good performance improvement in fading and near far conditions by combining MAI suppression capabilities of the CMA and SIC. At reasonable system loading of 10 users, the proposed design structure gave the same performance of a single user. Much improved user capacity compared with conventional SIC and CMA and MF receivers are shown to be achieved. The effect of step-size on the BER is also investigated.

A novel blind multistage PIC receiver referred to as BA-PIC for uplink DS-CDMA is proposed in 3.3, exploiting the constant modulus property of users' signals. An effective algorithm for blind interference suppression for detection and MAI amplitude estimation for the cancellation is also proposed to mitigate the error propagation and fading channel estimation problems of conventional PIC. Besides being blind and low complexity PIC, it showed significantly improved performance compared to conventional PIC in fading and AWGN environments also with severe near far problem. For examples, with 2 stage of interference cancellation, the proposed BA-PIC showed near single user performance for the weakest user under high system load of $K/N \approx 0.65$ and power of interferers as high as $15dB$.

The MAI estimation performance and achievable spectral efficiency of the proposed blind adaptive PIC and SIC techniques in Rayleigh fading environments is also investigated and compared with other receivers. The spectral efficiency in bps/Hz is calculated based on output SINR, which is also the function of MSE of MAI estimation conditioned on users' data. Particularly, the blind PIC is shown to give highest ≈ 7.3 bps/Hz compared with ≈ 5.3 for conventional PIC, ≈ 4.5 for CMA-SIC, ≈ 3.0 for SIC, and ≈ 2.7 for MF receivers.

In Section 3.4, a low complexity subcarrier combining technique for MC-CDMA receiver is presented in mobile Rayleigh channel. It is shown that, by exploiting the structure formed by users' repeating sequences into an adaptive algorithm with simple CM cost function and judicious selection of step-size, simultaneous supersession of MAI and implicit channel tracking is achieved without any knowledge of the channel. Improved BER and user capacity compared with MRC and EGC are shown. For example, at the BER of 0.01, it can support 17 users compared with 7 and 4 users for MRC and EGC, respectively. Optimum step-size are also investigated using simulations to give more improved performance.

To sum up, this Chapter has made an effort to apply a blind adaptive approach to tackle the MAI and imperfect channel estimation problems in the uplink of CDMA. The next Chapter is focused to address the problems in achieving spatial diversity for the uplink of CCMA and CDMA. We propose to solve the problems of deep fade as well as the MAI in a joint manner by combining user collaboration and multiuser detection.

Chapter 4

Collaborative Space Diversity Schemes

4.1 Introduction

The work presented in this Chapter are aimed at finding new ways of providing spatial diversity in uplink of wireless systems employing multiple access schemes such as CCMA and CDMA. These investigations are inspired by popular techniques such as space-time coding [73, 2], cooperative diversity [8, 7], which are used extensively for single user systems to improve the outage capacity and to counter the effects of slow fading, shadowing etc. There has been relatively very little work on applying these techniques to solve the problem of degraded BER performance under multiuser settings. The main reasons behind the lack of research in this subject area are:

- The uplink of multiuser systems naturally comes with problem of reduced orthogonality (in the case of CDMA) among the users' transmissions and hence the loss in achievable rate due to MAI
- The interference problem arising in such environments requires more advanced MUD receivers, which may complicate the performance analysis significantly.

The contributions presented in this Chapter attempts to fill this gap to some extent. It is however acknowledged that there is a need for more research work to make full use of the user cooperation diversity technique in practical multiuser wireless systems. The term cooperative diversity is usually used to denote the spatial diversity that is achieved via cooperation of users/nodes [7, 8, 79, 75, 85], which is also referred to here as collaborative diversity. Although there are differences where the word Collaboration is used, the two terms are treated equivalent in this Chapter and hence used interchangeably.

The first two work are proposed for uplink of CCMA to provide transmitter diversity gain. In the first scheme presented in Section 4.2, diversity gain is achieved by simply allowing each user node to transmit from one antenna at a time. The latter scheme described in Section 4.3, is focused on providing the diversity gain from user cooperation and hence presents a practical alternative to the former. The cooperative diversity technique is also investigated for uplink of CDMA with MAI. Two schemes are presented for CDMA that address the issues of providing transmitter diversity gains while minimizing the effects of the MAI. More specifically, new schemes consisting of user cooperation diversity and successive interference cancellation technique are proposed and evaluated in Section 4.4. Finally, in Section 4.5, a new bandwidth efficient collaborative diversity (BECD) scheme is proposed and evaluated in uplink of CDMA.

4.2 Simple Transmitter Diversity Scheme for CCMA

In CCMA, the multiple access function is achieved by employing collaborative codes, allowing multiple users to transmit independently without subdivision in time, frequency or orthogonal codes. The CCMA technique potentially offers higher transmission rate than other multiple access techniques leading to higher capacity and more efficient system [9, 10, 11, 12, 13], [14], [46]. However, one of the main problems of this technique in wireless systems is the practical combination of signals to provide the unique decodability of the composite signals at the receiver. In [15, 46] effective and practical approach to address this problem is proposed by the use of complex valued CCMA (CV-CCMA) and joint channel estimation and detection.

It is well understood that the use of space, time and other forms of diversity schemes significantly improve the error performance of both single user and multi-user communication systems [31, 102, 33, 34, 103, 104, 105]. The receiver diversity is implemented simply by having multiple receivers chain and this provides the corresponding diversity order gains. The transmitter diversity is an interesting approach to achieve the spatial diversity and has been mainly investigated for single user systems, e.g. [73]. The implementation of the transmitter diversity for multi-user communication is more complicated compared to the single user systems by the fact that the presence of co-channel signals from multiple users' and their antennas degrades the detection performance due to loss of orthogonality among the signals [2]. Therefore, very little work could be found in the literature that address the issue of obtaining transmit diversity in a multiple access system environments. In [105], design and performance evaluation of signalling schemes are proposed for a synchronous uplink system with users employing multiple transmit antennas and a base-station receiver with single or multiple receiver antennas using so-called Interference Resistant Modulation (IRM). The scheme is based on assigning unique phase rotation matrix to the users and reception is a generalization of ML joint multiuser detection with a processing layer of interference avoidance using the unique phase matrix. It is to be noted that the IRM scheme transforms BPSK modulated constellation to multilevel signals before transmission. CCMA technique proposes a solution to the detection problem under multiuser environment with unique decodability property of superposition of user codewords and can also include some error correction capability [10], [14]. Performance gain with the use of receive diversity for CCMA is investigated in [15, 46].

In this Section, a transmitter based scheme is proposed for achieving the diversity gain in the uplink of multiuser system employing CCMA technique. Here, the users equipped with multiple antennas transmit their codewords to the base-station using collaborative coding to achieve spatial diversity [1]. The communication protocol is designed as such to transmit the coded signals over two codeword periods from the two antennas of the users to allow for providing two copies of the user signals at the receiver from different channels. The receiver uses the received signal from two codeword periods and using the channel knowledge of the users' signals, performs search for the codewords that minimizes the sum of distance metric to the received signals. The estimates of transmitted user data are then obtained using the look-up table decoding of the codewords to data as used in the transmitters. Full system design is explained next.

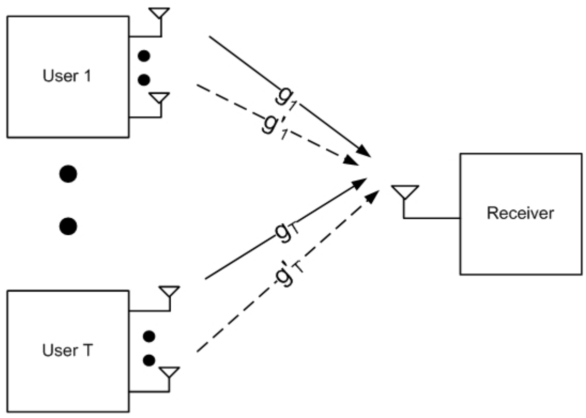


Figure 4.1: Transmit diversity system model for CCMA

4.2.1 Transmission and Detection Scheme

The multiuser communication system with T users and a base-station receiver is considered and is shown in Figure 4.1. The data from i^{th} user b_i is first encoded with collaborative codes $\mathbf{C}_i = \{C_{i1}, C_{i2}, ..C_{il}.., C_{iN_i}\}$, where C_{il} is the l^{th} codeword of i^{th} user, of length n bits and N_i is the number of codewords for the i^{th} user assumed equal for all users. The transmitted signal elements of each codeword is denoted by $C_{il} = \{s_{i1}, s_{i2}, ..s_{ij}.., s_{in}\}$ that use BPSK mapping. The channels are assumed flat Rayleigh fading and remain constant over the duration of a codeword period, modeled here as samples of complex Gaussian random variables with variances $\sigma_i^2 = 1$. We assume that system is fully synchronized before the transmissions and reception are carried out.

Under the proposed transmitter diversity scheme, a codeword signal of i^{th} user C_{il} is trans-

[1]It is assumed that the antennas are placed at least $\frac{\lambda}{2}$ apart to ensure independence of fading on each antenna subchannel [30]

mitted from two antennas during the first and second codeword period and is shown in Figure 4.2. The same signalling scheme applies to all other consecutive time periods. The received signals at

	Antenna 1	Antenna 2
Time 1	C_{il}	–
Time 2	–	C_{il}

Figure 4.2: Signalling scheme for i^{th} CCMA user

the base-station from all users during the first and second period r_j, r'_j respectively are given by

$$
\begin{aligned}
r_j &= \sum_{i=1}^{T} s_{ij} g_i + w, \\
r'_j &= \sum_{i=1}^{T} s'_{ij} g'_i + w, \\
& 1 \leq j \leq n
\end{aligned}
\tag{4.1}
$$

where g_i, g'_i are the transmit channels of the i^{th} user to the base-station from first and second antenna respectively and w is AWGN component with two sided power spectral densities $N_0/2$.

In a T-user CCMA transmission, there are L allowable codeword combinations $\mathbf{A}_k = \{a_{k1}, a_{k2}, .., a_{kn}\}$, consisting of the users' codewords over their corresponding complex channels. Each element a_{kj} is the j^{th} symbol of k^{th} allowed codeword combination and is given by

$$
\begin{aligned}
a_{kj} &= \sum_{i=1}^{T} s_{ij} g_i, \\
1 \leq k \leq L&, 1 \leq j \leq n.
\end{aligned}
\tag{4.2}
$$

Similarly codeword combinations $\mathbf{A}'_k = \{a'_{k1}, a'_{k2}, .., a'_{kn}\}$ are defined for the second period and each element is given by

$$
\begin{aligned}
a'_{kj} &= \sum_{i=1}^{T} s_{ij} g'_i, \\
1 \leq k \leq L&, 1 \leq j \leq n.
\end{aligned}
\tag{4.3}
$$

The receiver performs joint detection and ML decoding of users' codewords and provides the estimates of transmitted data signals of all users. The distance metrics from the received signals for each combination of codewords at two consecutive periods are denoted as d_k and d'_k, respectively. The distance metrics are calculated by utilizing the estimates of users' corresponding channels

$\hat{g}_i, \hat{g}'_i$ for each combination of codewords as follows

$$\begin{aligned} d_k &= \sum_{j=1}^{n} \left| r_j - \sum_{i=1}^{T} s_{ij} \hat{g}_i \right|^2 \\ d'_k &= \sum_{j=1}^{n} \left| r'_j - \sum_{i=1}^{T} s_{ij} \hat{g}'_i \right|^2 \\ &\forall k, 1 \leq k \leq L. \end{aligned} \tag{4.4}$$

The calculated distance metrics d_k and d'_k for each combination of codewords from the two periods are used to perform ML decoding and the combination of allowable codewords that minimizes the sum of distance metrics are selected as the transmitted codewords of the users

$$\{\hat{C}_1, \hat{C}_2\} = \arg \min_{C_{1l}, C_{2l} \in \mathbf{W}} \left\{ d_k + d'_k \right\}, \tag{4.5}$$

where **W** is the set of codewords of all users. Finally, the data symbols of the users are obtained by using a lookup table decoding of the codewords to the symbols as used in the transmitters. The complexity of this ML search is reasonably low since the codewords are usually very short $n = 2, 3$. Note that the scheme achieves second order diversity performance at the cost of increased bandwidth requirement since the users transmit signals from two antennas in orthogonal time slots.

4.2.2 BER Performance Analysis

In order to assess the performance of CCMA under different diversity schemes, a simple technique based on alternative representation of Q-function for linearly modulated digital signals in fading channels [106] is presented. For this purpose, error metric associated with each codeword combination of the transmitted signals are derived. Under the transmit antenna diversity scheme, the signal power of a user is equally distributed among antennas thus reducing the received SNR from each antenna by the factor of half. The absolute magnitude of squared distance between different codeword combinations are calculated and normalized with n to give an error metric z_m as follows

$$\begin{aligned} z_m = \frac{\left| \left\{ \sum_{i=1}^{T} C_i \right\}_x - \left\{ \sum_{i=1}^{T} C_i \right\}_y \right|^2}{n}; \\ 1 < m < M; 1 < x \neq y < L, \end{aligned} \tag{4.6}$$

where $\left\{ \sum_{i=1}^{T} C_i \right\}_x$ and $\left\{ \sum_{i=1}^{T} C_i \right\}_y$ are any two possible codeword combinations of the users, $M = \sum_{m=1}^{L-1} m$ is the total number of possible distances between codeword combinations. Assuming that all the codewords are equally likely to be transmitted, it is desirable to find the average error metric so that the tools designed to evaluate the performance of a single user transmission in fading channels can be used. Averaging z_m over all M possible distances between the codeword combinations, the average error metric z is obtained as follows

$$z = \frac{\sum_{m=1}^{M} z_m}{M}. \tag{4.7}$$

The average BER performance of the system conditioned on the instantaneous SNR of channels of users' transmit antennas can also be written as

$$P(z|_{\gamma,\gamma'}) = Q\left(\sqrt{z\Big(\gamma + \gamma'\Big)}\right), \tag{4.8}$$

where $Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt$, γ and γ' are the instantaneous SNRs of the transmit channels of a user from first and second antenna, respectively. It is assumed that all users have independent fading distributions with same average SNR. Therefore $P(z)$ conditioned on instantaneous fading SNR of a single user well approximates the average BER of a system. In the case of unequal user transmit power, $P(z)$ will be different for each user and (4.8) can be modified accordingly. To obtain the exact error probability $P(z)$ in fading channels, (4.8) has to be calculated over all the fading events of the users' channels [106] and is given by

$$P(z) = \int_0^\infty \int_0^\infty P(z|_{\gamma\gamma'}) p(\gamma) p(\gamma') d\gamma d\gamma', \tag{4.9}$$

where $p(\gamma)$ and $p(\gamma')$ are the PDF of fading distributions of a user from the first and second antenna, respectively. Using the alternative approach for Q-function proposed in [107], $P(z)$ can then be written as

$$P(z) = \frac{1}{\pi} \int_0^{\pi/2} \left(1 + \frac{z\Gamma}{sin^2\theta}\right)^{-1} \left(1 + \frac{z\Gamma'}{sin^2\theta}\right)^{-1} d\theta, \tag{4.10}$$

where $\Gamma = \Gamma' = \Gamma_0/2$, are the ensemble average SNR of fading distributions of a user's antenna and Γ_0 is the total average SNR for a user's signal. An upper bound on the $P(z)$ is obtained by knowing the fact that the integrands in (4.10) are maximized when $sin^2\theta = 1$. Since average distance measure of composite codewords is used in (4.10), we are able to obtain an approximation of probability of error. Which for CCMA signals with two antenna transmit diversity can be shown as

$$P(z) \approx \frac{1}{2}\left(\frac{1}{1 + z\Gamma_0/2}\right)\left(\frac{1}{1 + z\Gamma_0/2}\right). \tag{4.11}$$

4.2.3 Simulation Results and Comparisons

A simple two user CCMA system with two antennas each using BPSK mapping and two codewords per user each of length n=3 is used. The modulated codewords of user 1 and user 2 are $C_1 = \{1, 1, 1\}, \{1, -1, 1\}$ and $C_2 = \{-1, -1, 1\}, \{-1, 1, -1\}$, respectively and as shown in Figure 4.3 their combinations are unique. These codes are taken for consistency with results of

	{ 1, 1, 1 }	{ 1, -1, 1 }
{ -1, -1, 1 }	{ 0, 0, 2 }	{ 0, -2, 2 }
{ -1, 1, -1 }	{ 0, 2, 0 }	{ 0, 0, 0 }

Figure 4.3: Two-user collaborative codeword combinations

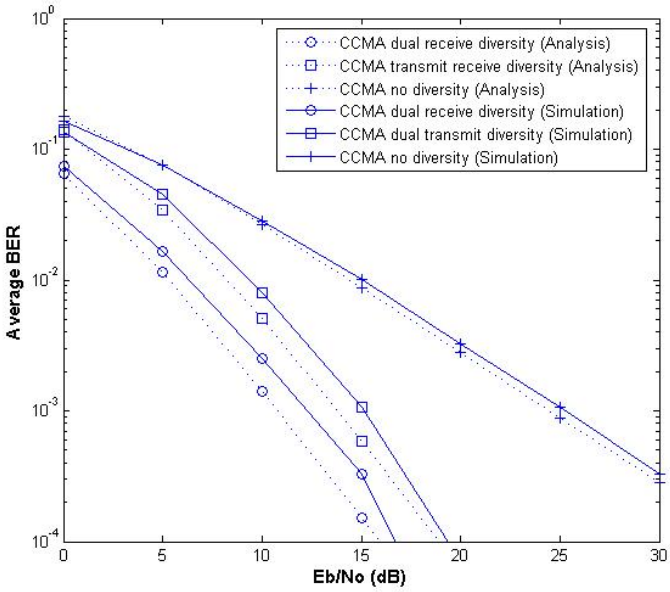


Figure 4.4: Performance of transmit diversity in flat Rayleigh fading channels for a 2-user CCMA

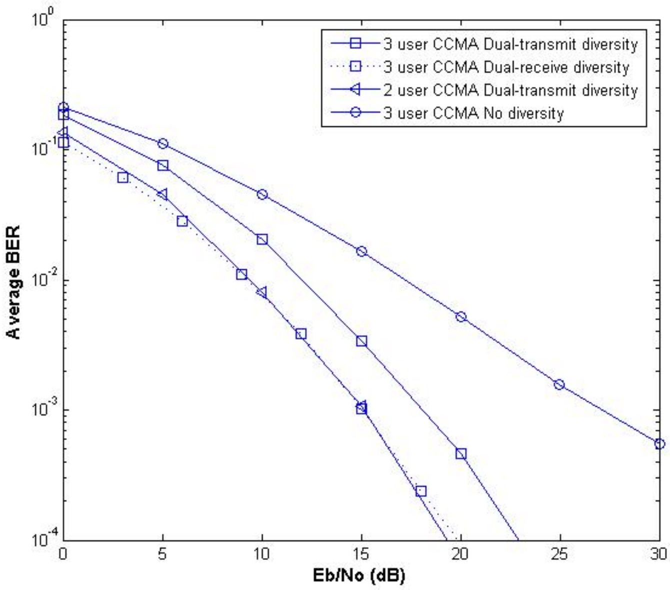


Figure 4.5: Performance of transmit diversity in flat Rayleigh fading channels for a 3-user CCMA

earlier schemes employing CCMA as in [15, 46]. It is assumed that the base-station receiver has perfect knowledge of channels of the users from all antennas. In Figure 4.4, the derived BER performance bounds of the CCMA is plotted using these codewords as given in equations (4.11) for the case of CCMA with transmit diversity and using (4.27) and (4.28) for the receive diversity and no diversity. The BER performance results of the users from the simulations are averaged and shown in the plot. As expected, the transmitter diversity scheme is shown to offer significant performance gain compared to no diversity case. The performance loss of 3dB compared to CCMA with dual receive diversity is incurred. This is due to the need for splitting the power to allocate to different antennas at the transmitters for ensuring that the total power is the same as without using multiple antennas. It is noted that the simulation results match closely with that obtained from the analysis based on average distances of composite codewords.

In Figure 4.5, the performance of dual transmitter diversity scheme for a 3 user system is presented. The results confirm that transmit diversity gain is achieved regardless of number of users. As expected, the transmitter diversity scheme shows 3dB penalty in BER compared that the system with a dual receive diversity It is also noted that the performance of a 3 user system with dual receive diversity has comparable performance to 2 user system with dual transmit diversity indicating approximately 3dB of loss for each additional user in the system. To sum up, we have shown a simple transmitter based diversity scheme for CCMA and evaluated and verified its performance. How we can achieve the similar diversity performances without using multiple antennas is the subject of study in the next section.

4.3 User Collaborative Diversity for CCMA

Recently, cooperative communication has emerged as an interesting approach to improve the link performance of wireless networks by sharing the antennas and other resources among the mobile nodes (users) [83, 75, 8, 7]. It becomes more useful particularly for the mobile users, which can not due to their size and power limitations, employ more than one antenna to communicate with other users or base-station. The cooperation concept is used much earlier in the case of improving the performance of node to node communication by using additional relay nodes e.g.[78, 77]. The paper by Sendonaris et.al. [7] has considered user cooperation in multiuser communications and particularly for CDMA. In this Section, a collaborative scheme is proposed for achieving the diversity gain in the uplink of a multiuser system employing CCMA [12], [14], [46]. Here, the users are proposed to work in pairs to transmit their own and each other's information to the base-station with the use of collaborative codes. The reception is based on the joint detection of codewords similar to CV-CCMA. The communication protocol is designed as such to transmit the coded signals over two codeword periods from the two cooperating users to allow for the exchange of users' data and provide two copies of the signals at the receiver from different channels.

The Section is organized as follows. In subsection 4.3.1, the cooperative CCMA system model is presented. The proposed cooperative CCMA implementation is described in subsection 4.3.2. The bit error performance analysis and simulation results are presented in subsection 4.3.3 and 4.3.4, respectively.

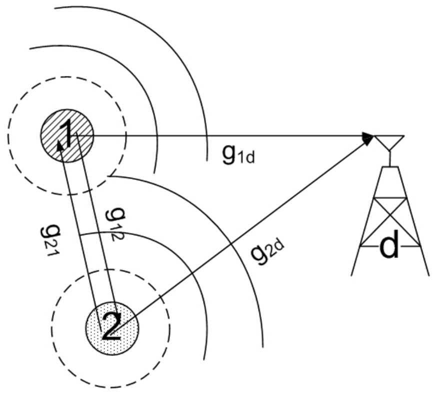


Figure 4.6: Two user cooperation scenario for CCMA

4.3.1 System Model

A multiuser communication system of an uplink CCMA with T users and a base-station receiver $\{d\}$ is considered. The cooperation scenario between two users is shown in Figure 4.6, where the users communicate between themselves and the base-station receiver. The cooperating users are denoted as user i with its partner u and vice versa. The data from i^{th} user b_i is first encoded with collaborative codes $\mathbf{C}_i = \{C_{i1}, C_{i2}, ..C_{il}.., C_{iN_i}\}$, where C_{il} is the l^{th} codeword of i^{th} user of length n bits and N_i is the number of codewords for the i^{th} user assumed equal for all users. The transmitted signal elements of each codeword is denoted by $C_{il} = \{s_{i1}, s_{i2}, ..s_{ij}.., s_{in}\}$ that use BPSK mapping. The channels are assumed flat Rayleigh fading and remain constant over the duration of the cooperating periods. Which are obtained from samples of complex Gaussian random variable with variances $\sigma^2_{iu}, \{i \neq u\}, \sigma^2_{id}$. We assume that the system is fully synchronized before the transmissions and detection methods are carried out at different nodes.

For $T = 2$, the cooperation is performed over two codeword periods, to exchange users' information and retransmit to the base-station. The received signals during the exchange period at the first and second user are given by

$$\begin{aligned} r_{1j} &= s_{2j} g_{21} + w_1, \\ r_{2j} &= s_{1j} g_{12} + w_2, \\ &1 \leq j \leq n \end{aligned} \tag{4.12}$$

where r_{1j} is the j^{th} received coded symbol at the 1^{st} user, s_{2j} is j^{th} transmitted coded symbol of the 2^{nd} user, g_{12}, g_{21} are the inter-user channels between the cooperating users with variances σ_{12}^2, σ_{21}^2, respectively and w_1 is AWGN with two sided power spectral densities $N_0/2$ for 1^{st} user.

The received signals at the base-station d from both users during the first and second period r_{dj}, r'_{dj} , respectively are given by

$$\begin{aligned} r_{dj} &= s_{1j}g_{1d} + s_{2j}g_{2d} + w_d, \\ r'_{dj} &= s'_{1j}g_{2d} + s'_{2j}g_{1d} + w_d, \\ & 1 \leq j \leq n \end{aligned} \tag{4.13}$$

where $s_{1,j}$ and $s'_{1,j}$ are the transmitted signals of user 1 from its own and partner's node, g_{1d} and g_{2d} are the transmit channels of the users to the base-station with variances σ_{1d}^2 and σ_{2d}^2, respectively.

In a 2-user cooperative CCMA transmissions, there are L allowable codeword combinations $\mathbf{A}_k = \{a_{k1}, a_{k2}, .., a_{kn}\}, 1 \leq k \leq L$, consisting of the two users' composite codewords over their corresponding complex channels. Each element a_{kj} is the j^{th} symbol of k^{th} allowed composite codeword is given by

$$\begin{aligned} a_{kj} &= s_{1j}g_{1d} + s_{2j}g_{2d}, \\ 1 \leq k &\leq L, 1 \leq j \leq n. \end{aligned} \tag{4.14}$$

Similarly codeword combinations $\mathbf{A}'_k = \{a'_{k1}, a'_{k2}, .., a'_{kn}\}$ are defined for the second period and each element is given by

$$\begin{aligned} a'_{kj} &= s'_{1j}g_{2d} + s'_{2j}g_{1d}, \\ 1 \leq k &\leq L, 1 \leq j \leq n. \end{aligned} \tag{4.15}$$

For a cooperative scheme to perform satisfactorily, the inter-user channel gains are desired to be higher or at least equal to that of the respective transmit channels of the users to the destination (base-station in this work) [78]. Assuming equal average signal power and noise variances of all users' and base-station receivers, the relative signal to noise ratio (SNR) gain $\{\beta_1\}$ and $\{\beta_2\}$ of inter-user channels compared to the respective transmit channels of the users to the base-station can be expressed as

$$\beta_1 = \frac{\sigma_{12}^2}{\sigma_{1d}^2}, \quad \beta_2 = \frac{\sigma_{21}^2}{\sigma_{2d}^2}. \tag{4.16}$$

4.3.2 Protocol and Detection Scheme

Based on the system model, a signalling method of the proposed scheme with two users spanned over two consecutive codeword periods is shown in Table 4.1. A single cycle of cooperative

At Transmitter

	t_1	t_2
User 1	C_{1l}	C'_{2l}
User 2	C_{2l}	C'_{1l}

At Receivers

	t_1	t_2
User 1	C_{2l}	-
User 2	C_{1l}	-
Base Station	C_{1l}+C_{2l}	C'_{2l}+C'_{1l}

Table 4.1: Two user cooperation protocol for CCMA

codewords transmission can also be written in a matrix form as

$$\mathbf{S} = \begin{pmatrix} C_{1l} & C_{2l}^{'} \\ C_{2l} & C_{1l}^{'} \end{pmatrix} \qquad (4.17)$$
$$1 \le l \le N_i,$$

where the columns indicate the codeword time periods (t_1, t_2) and the rows indicate users' codewords transmission. The signals in first column represent users' own codewords , whereas the signals in the second column are the estimated codewords at the partners' nodes.

Phase I: Codewords exchange

In the first period, the users' transmit their signals as shown in (4.17). Due to the broadcast nature of the channels, the signals are simultaneously received both at the cooperating users and at the base-station receiver. At the same time, the received signals are independently processed at the users' receivers. For handling these operations, full duplex capabilities are assumed to be available [7, 78]. The received signals at this period are given in (4.12). The detection of signals at each other user node is achieved by first obtaining the soft estimates of the signals, for example, the first user this is given by

$$\tilde{s}_{1j} = r_{2j} g_{12}^{*}; 1 < j < n, \qquad (4.18)$$

where * denotes complex conjugation operation. Then, by taking the sign of the real part of the signal $\tilde{s}_{1j}$ over the codeword length n, the estimate of the first user's transmitted signal s_{1j}' is obtained for transmission in second period

$$s_{1j}' = sgn\Big\{Re\big\{\tilde{s}_{1j}\big\}\Big\}; 1 < j < n, \qquad (4.19)$$

where $sgn\{.\}$ and $Re\{.\}$ denote the signum function and the real part of a complex number, respectively.

Phase II: Retransmission of detected codewords

During this phase, the cooperating users simply forward the detected codewords of the partners to the base-station receiver C'_{il}. It should be noted that this codeword may not be identical to the codewords transmitted by source users in the first phase due to detection errors.

Joint detection and decoding at base-station receiver:

The receiver performs joint detection and maximum likelihood (ML) decoding of users' codewords and provides the estimates of transmitted data signals of both users. The distance metrics from the received signals for each combination of codewords at two consecutive periods are denoted as d_k and d'_k respectively. The distance metrics are calculated by utilizing the estimates of users' corresponding channels for each combination of codewords as follows

$$\begin{aligned} d_k &= \sum_{j=1}^{n} \left| r_{dj} - \{s_{1j}\hat{g}_{1d} + s_{2j}\hat{g}_{2d}\} \right|^2 \\ d'_k &= \sum_{j=1}^{n} \left| r'_{dj} - \{s_{1j}\hat{g}_{2d} + s_{2j}\hat{g}_{1d}\} \right|^2 \\ & \qquad \forall k, 1 \leq k \leq L, \end{aligned} \tag{4.20}$$

where $\hat{g}_{1d}$ and $\hat{g}_{2d}$ are the channel estimates of the 1^{st} and 2^{nd} user respectively. The sum of calculated distance metrics d_k and d'_k for each combination of codewords from the two phases are used to perform ML decoding such that the combination of codewords that minimizes the sum of distance metrics are selected as the transmitted codewords of the users

$$\{\hat{C}_1, \hat{C}_2\} = \arg \min_{C_{1l}, C_{2l} \in \mathbf{W}} \left\{ d_k + d'_k \right\}, \tag{4.21}$$

where **W** is the set of codewords of all users. Finally, the data symbols of the users are obtained by using a lookup table decoding of the codewords to the symbols as used at the transmitters.

4.3.3 BER Analysis

The probability of bit error of the collaborative coded multiuser signals using BPSK mapping in flat slowly fading channels can be derived by doing some simple modifications to the tools developed for single user signals in [107, 108]. For this purpose, error metric associated with each codeword combination of the transmitted signals is derived. The absolute magnitude of distance between different composite codeword combinations are calculated and normalized with n to give an error metric z_m as follows

$$\begin{aligned} z_m &= \frac{\left| \{C_{1x} + C_{2x}\} - \{C_{1y} + C_{2y}\} \right|^2}{n} \\ & \qquad 1 < m < M; 1 < x \neq y < L, \end{aligned} \tag{4.22}$$

where $\{C_{1x}, C_{2x}\}$ and $\{C_{1y}, C_{2y}\}$ are any possible two composite codeword combinations of user 1 and 2, $M = \sum_{m=1}^{L-1} m$ is the total number of possible distances between codeword combinations.

Assuming that all the codewords are equally likely to be transmitted, it is appropriate to find the average error metric so that the tools developed for single user signals can be used. Averaging z_m over all M possible distances between the codeword combinations, the average error metric z , that is used for the bit error performance approximation of the CCMA scheme, is obtained as

$$z = \frac{\sum_{m=1}^{M} z_m}{M}. \tag{4.23}$$

Using z, the probability of bit error of existing (non-cooperative) CCMA over fading channels is now calculated, both with and without diversity. For the ideal cooperation case, i.e. where each user perfectly decodes its partner's codewords, the performance of the proposed scheme becomes identical to that of CCMA scheme with dual space diversity reception. The BER performance conditioned on the transmit channels of the cooperating users can also be written as

$$P(z|_{\gamma_{1d},\gamma_{2d}}) = Q\left(\sqrt{z\Big(\gamma_{1d} + \gamma_{2d}\Big)}\right), \tag{4.24}$$

where $Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} e^{-t^2/2} dt$ is the well known Gaussian error function , γ_{1d} and γ_{2d} are the instantaneous SNRs of the transmit channels of user 1 and 2, respectively. To obtain the average error probability $P(z)$ in fading channels, calculation has to be done (4.24) over all the fading events of the users [108] and is given by

$$P(z) = \int_0^{\infty} \int_0^{\infty} P(z|_{\gamma_{1d}\gamma_{2d}}) p(\gamma_{1d}) p(\gamma_{2d}) d\gamma_{1d} d\gamma_{2d}, \tag{4.25}$$

where $p(\gamma_{1d})$ and $p(\gamma_{2d})$ are the PDF of fading distributions of user 1 and 2 respectively. In [108] a unified approach to calculate the probability of error of linearly modulated single user digital signals in arbitrary fading channels is proposed. This approach, using moment generating function (MGF) and alternate representation of Q function, is originally proposed by Craig in [107]. It allows the expressions within indefinite integrals of fading events in (4.25) to be accurately approximated using set of definite integrals. Using this approach $P(z)$ can then be written as

$$P(z) = \frac{1}{\pi} \int_0^{\pi/2} \left(1 + \frac{z\Gamma_{1d}}{sin^2\theta}\right)^{-1} \left(1 + \frac{z\Gamma_{2d}}{sin^2\theta}\right)^{-1} d\theta, \tag{4.26}$$

where Γ_{1d} and Γ_{2d} , are the ensemble average SNR of fading distributions of the user 1 and 2 with instantaneous SNR of γ_{1d} and γ_{2d}, respectively. The upper bound on the $P(z)$ is obtained by knowing the fact that the integrands in (4.26) are maximized when $sin^2\theta = 1$. Since average distance of codewords is used for our analysis, an approximation of probability of error of CCMA signals with dual receive diversity can be shown as

$$P(z) \approx \frac{1}{2}\left(\frac{1}{1 + z\Gamma_{1d}}\right)\left(\frac{1}{1 + z\Gamma_{2d}}\right). \tag{4.27}$$

For the case of the CCMA without diversity, using the error metric of (4.23), the the probability of error $P(z)$ can be derived as follows:

$$P(z) \approx \frac{1}{2}\left(\frac{1}{1+z\Gamma_d}\right), \tag{4.28}$$

where $\Gamma_d = 1/2(\Gamma_{1d} + \Gamma_{2d})$ is the averaged SNR of users' channels to the base-station receiver.

	{ 1, 1, 1 }	{ 1, -1, 1 }
{ 1, -1, 1 }	{2, 0, 2 }	{2, -2, 2 }
{-1, 1, 1 }	{0, 2, 2 }	{0, 0, 2 }

Figure 4.7: Two user collaborative codeword combinations

4.3.4 Performance Results and Comparisons

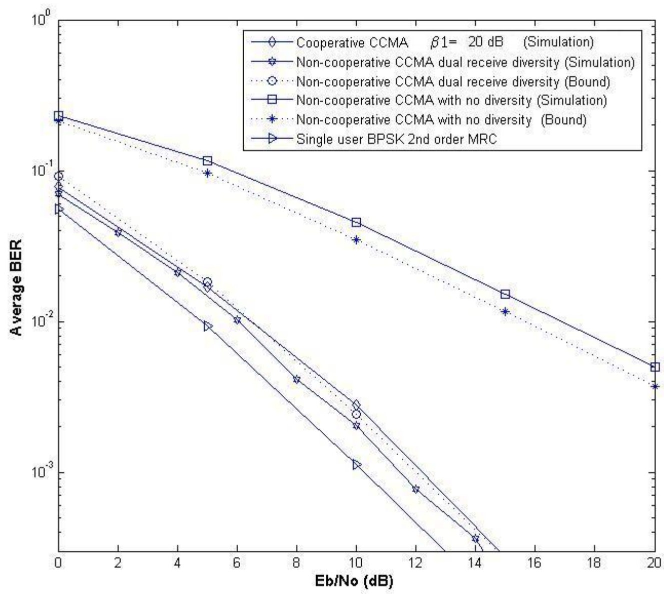


Figure 4.8: BER performance in flat Rayleigh fading channels for cooperative and non-cooperative 2-user CCMA schemes

In this subsection, the performance bounds and simulation results of the cooperative and non-cooperative (with and without receive diversity) CCMA schemes are presented under different channel conditions. A simple 2-user CCMA system with BPSK mapping and two codewords per user each of length 3 is used. The modulated codewords of user 1 and user 2 are $C_1 = \{1, 1, 1\}, \{1, -1, 1\}$ and $C_2 = \{1, -1, 1\}, \{-1, 1, 1\}$, respectively and shown in Figure 4.7. It is assumed that all the users' and the base-station receivers have perfect knowledge of their received channels.

In Figure 4.8, the derived BER performance bounds of the non-cooperative CCMA using the codewords as given in equations (4.27) and (4.28)are shown. For the purpose of verification, the simulation results are also obtained and shown in Figure 4.8. It is noted that the derived BER bounds become tighter with the increase of diversity order. The performance of the cooperative CCMA with inter-user channel SNR gain of ($\beta_1 = \beta_2 = 20dB$) is shown as expected to be within the range of the dual diversity and no diversity bounds. Also, the BER performance of a single user BPSK with dual diversity using maximum ratio combining (MRC) is shown for comparison. Figure 4.9 shows the BER simulation results of the proposed 2-user cooperative and the non-

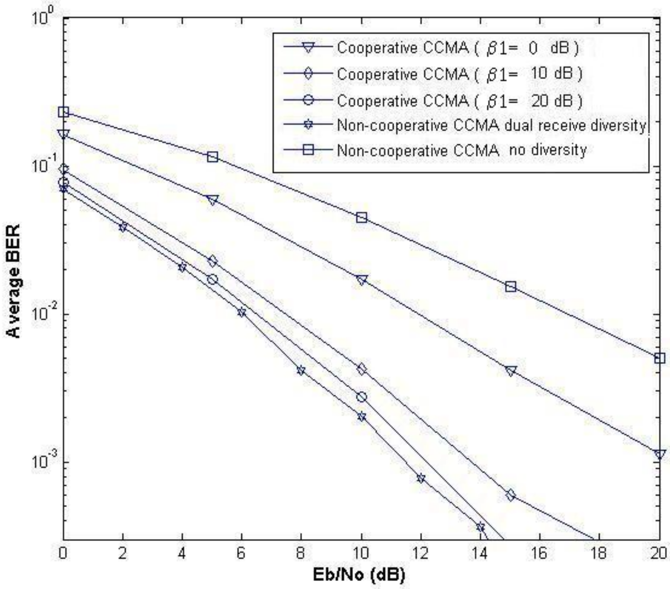


Figure 4.9: BER performance in flat Rayleigh fading channels for the 2-user cooperative CCMA under different inter-user channel SNR gains

cooperative CCMA with same channel settings. The ratios β_1 and β_2 described in (4.16) are used to quantify the degree of cooperation of the proposed scheme. The variances of all users transmit channels to the base-station are assumed to be equal to one ($\sigma_{1d}^2 = \sigma_{2d}^2 = 1$) and $\beta_1 = \beta_2$. Then as expected, as β_1 increases, the degree of cooperation increases and the proposed scheme shows rapid improvement in BER performance. Also it is noted from the figure that even when all the channels have average equal variances i.e. $\beta_1 = 0dB$, the scheme still offers several dB of E_b/N_0 gain compared to the non-cooperative scheme. When $\beta_1 = 20dB$, at the BER of 10^{-3}, the performance of cooperative CCMA is only around $0.5dB$ worse to that of CCMA with dual receive diversity. Figure 4.10 shows the BER performance of CCMA schemes for the case of non-identical average variances of each cooperating users' transmit channels to the base-station

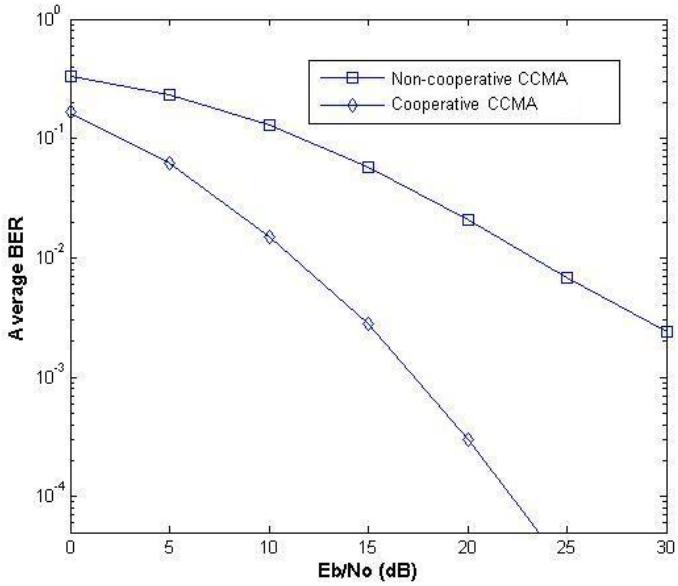


Figure 4.10: BER performance in flat Rayleigh fading channels for the 2-user cooperative CCMA and non-cooperative CCMA with channel asymmetry condition of $\beta_d = 10dB$ and inter-user SNR gains $\beta_1 = 0dB$, $\beta_2 = 10dB$

receiver. The condition, also termed as 'channel asymmetry', is defined by the ratio given below

$$\beta_d = \frac{\sigma_{1d}^2}{\sigma_{2d}^2} \tag{4.29}$$

For the channel asymmetry ratio of $\beta_d = 10dB$, when the inter-user channels have relative SNR gain of $\beta_1 = 0dB$, $\beta_2 = 10dB$, the cooperative CCMA scheme offers significant improvement in error performance for both users compared to that with non-cooperative CCMA without diversity. This result is very beneficial for the mobile users, and particularly for the weaker users as their performances are now significantly less sensitive to the fading than in the non-cooperative case.

The transmit antenna diversity and user collaborative diversity techniques for CCMA have shown to be very effective for operation in fading environments. Unlike CCMA where all users are detected jointly, in CDMA each user is detected separately while other users contribute to MAI. Hence achieving full diversity in this case becomes more challenging, the work presented in next section addresses this problem.

4.4 Improved User Collaborative Diversity with Interference Cancellation for CDMA

4.4.1 Introduction

The need for increased data rates, reliable communication and improved Quality of Service (QoS) has necessitated for high capacity and reliable uplink CDMA systems. It is well known that the use of antenna diversity provides significant capacity improvement of wireless communications [31]. While employing multiple antenna scheme at the base-station is easier to perform than that in mobile handsets, in many cases, alternative solutions to the scheme may be desirable. Cooperative communication has emerged as an interesting approach to address this problem and improve the link performance of wireless networks by sharing the antennas and other resources among the users [7, 8, 79, 75, 85, 20].

Cooperative diversity for uplink of CDMA has been the subject of study in many recent work e. g. [7, 85, 80, 20, 76, 109, 84]. The paper by Sendonaris et. al. [7] has considered user cooperation for CDMA system. However, the main practical problem of multiple access interference (MAI) is not considered. In practice MAI may have profound effects the performance in cooperative CDMA systems. The performance of a single user with relay assisted diversity for uplink of CDMA under different propagation environments is investigated in [84]. It is shown that the conventional matched filter detector fails to attain full diversity gain. And hence an improved receiver is proposed that suppresses MAI from relays. However, the cooperation and reception techniques for CDMA under realistic multiuser environment has not been considered. To address the MAI, more complex multiuser detection (MUD) techniques such as decorrelation or MMSE combined with user cooperation are also investigated in [85, 76, 109]. It is well known that SIC is a very effective detection technique for CDMA with complexity comparable to that of simple MF receivers. The combined user cooperation and successive interference cancellation is proposed here to address the MAI and complexity issue in uplink of CDMA. Our arguments for the new scheme are also supported by number of simulation results and also an asymptotic analysis of capacity gains.

Two techniques employing user collaboration and SIC are proposed in this Section. The first scheme is referred to as Collaborative SIC (C-SIC) that uses MF outputs for MAI estimation and cancellation. The estimation of such SIC, however, is known to exhibit error propagation to later users' stages and thus may perform poorly when system load increases [65]. This problem can be alleviated great extent by employing the blind adaptive SIC (BA-SIC) [16] that takes into account multiuser interference while despreading and estimation. The work employing the BA-SIC with the user cooperation also referred to as C-BASIC. As will be shown later, the C-BASIC improves the system performance considerably.

The Section is organized as follows. In subsection 4.4.2, the system model is presented. The collaborative transmission protocol is described in subsection 4.4.3 and the operation of C-SIC receiver is described in subsection 4.4.4. The achievable rate analysis of the C-SIC and comparisons are presented in subsection 4.4.5. The C-BASIC receiver is then described in subsection 4.4.6. Subsection 4.4.7 shows the BER simulation results and comparisons.

4.4.2 System Model

A typical multiuser communication scenario of an uplink synchronous CDMA with a pair of cooperating users $\{1, 2\}$ and the base-station receiver $\{d\}$ system employing the proposed collaborative scheme is shown in Figure 4.11. A common multiple access channel (MAC) with equal power BPSK modulated user signals with fading and AWGN is assumed. To gain clear insight into the impact of cooperation on multiuser SIC reception under MAI conditions, the notes on few assumptions made in this work as follows:

- The cooperating pair of users are chip synchronized before they start to cooperate and transmit each others' data. Extension to asynchronous case should be possible with some modification to the scheme.

- The amount of interference from non paired user nodes to the cooperating pair of users is small and can be treated as background noise. This may be well justified due to uniform distribution of users within the coverage of a typical cellular system, the user nodes see much less interference from other users compared to the base-station receiver (which is usually placed in the center of a cell to be able to transmit and receive to/from all users more efficiently).

- The pairing users are assumed to know the phases of their transmit channels such that while transmission their signals are multiplied with appropriate phase offsets for coherent combining at the base-station receiver [7].

In the collaborative uplink CDMA, during the first symbol period, the users transmit their own data with their originally assigned spreading sequences. At the same time, the users that are pairing with transmitting users receive and decode the transmitted signals. During the second period, the pairing users forward the decoded data using the spreading sequences of their partners. The signals received at the cooperating pairs $\{k, i\}$ at the first period can be written as:

$$\begin{aligned} r_i(t) &= \sqrt{P_k} g_{ki}(t) b_k(t) c_k(t) + v_i(t), \\ r_k(t) &= \sqrt{P_i} g_{ik}(t) b_i(t) c_i(t) + v_k(t), \end{aligned} \tag{4.30}$$

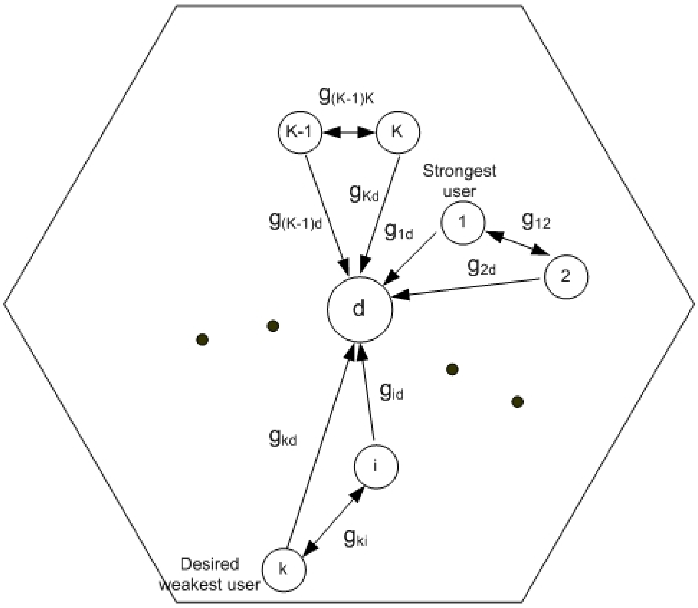


Figure 4.11: Cooperation scenario between pairs of CDMA users

where P is the signal power, $g(t) = \alpha(t)e^{-j\pi\phi(t)}$ is the complex fading channel of the users with amplitude $\alpha(t)$ and phase $\phi(t)$ components with variance σ^2, $b_k(t) = \sum_{m=-\infty}^{\infty} b_k(m)p(t - mT_b)$ is the data signal, where $b_k(m)$ is a binary sequence taking values $[-1, +1]$ with equal probabilities, $p(t)$ is rectangular pulse with period T_b. The spreading sequence is denoted as $c_k(t) = \sum_{n=-\infty}^{\infty} c_k(n)p(t-nT_c)$ with antipodal chips $c_k(n)$ of rectangular pulse shaping function $p(t)$ with period T_c and with normalized power over a symbol period equal to unity $\int_0^{T_b} c_k(t)^2 dt = 1$. The spreading factor is $N = T_b/T_c$ and $v(t)$ is the AWGN with two sided power spectral density $N_0/2$.

The received composite signal at the base-station receiver d from all users' transmissions during the first period $r_d(t)$ and from that of the partnering users' in the second period $r'_d(t)$ can be written as:

$$\begin{aligned} r_d(t) &= \sum_{k=1}^{K} g_{kd}(t)s_k(t) + v(t), \\ r'_d(t) &= \sum_{i=1, i\neq k}^{K} g_{id}(t)s_i(t) + v(t), \end{aligned} \tag{4.31}$$

where $s_k(t) = \sqrt{P_k}b_k(t)c_k(t)$ is the transmitted signal of k^{th} user. The model of signal $s_i(t)$ transmitted during the second period is exactly the same as above but they are originated from the $i, i \neq k$ user with k^{th} user's estimated data $b'_k(t)$ with their own channels $g_{id}(t) = \alpha_i(t)e^{-j\pi\phi_{id}(t)}$. The signal model described above applies to all pairs of cooperating users with appropriate modifications.

The cooperative scheme performs satisfactorily when the inter-user channel gains are higher or at least equal to that of the respective transmit channels of the users to the destination (base-station in this work) [78]. Assuming average noise variance of the users and the base-station receivers are equal, the relative signal to noise ratio (SNR) gain in dB of inter-user channels β_k, β_i compared to the respective transmit channels of the users to the base-station can be shown as

$$\beta_k = \frac{\sigma_{ki}^2}{\sigma_{kd}^2}, \beta_i = \frac{\sigma_{ik}^2}{\sigma_{id}^2} \tag{4.32}$$

where, σ_{ki}^2, σ_{ik}^2 and σ_{kd}^2, σ_{id}^2 are the variances of inter-user channels g_{ki}, g_{ik} and the users' channels to the base-station receiver g_{kd}, g_{id}, respectively. Symmetry of inter-user channels i.e. $g_{ki} = g_{ik}, \forall k$ assumed here is reasonable as in [7]. In the near far condition, $\sigma_{kd}^2 \neq \sigma_{id}^2$ and hence $\beta_k \neq \beta_i$ with nearfar ratio being defined as $\Omega = \max\{\sigma_{jd}^2\}/\sigma_{kd}^2, j \in \{1, 2, ., i, ., K\}, j \neq k$.

4.4.3 Collaborative Transmission Protocol

Based on the system model, a signalling structure which applied to both C-SIC and C-BASIC, with two users spanned over two consecutive symbol periods is shown in (4.33). The same signalling structure applies to all other consecutive periods. When appropriate, the signals are presented in vector form and indices denoting time dependance are dropped. A single cycle of collaborative transmission scheme can also be written as

$$\begin{aligned} \mathbf{s}_k &= \underbrace{\sqrt{P_k} g_k b_k \mathbf{c}_k}_{\text{first period}}, \underbrace{\sqrt{P_k} g_k b_i' \mathbf{c}_i}_{\text{second period}} \\ \mathbf{s}_i &= \underbrace{\sqrt{P_i} g_i b_i \mathbf{c}_i}_{\text{first period}}, \underbrace{\sqrt{P_i} g_i b_k' \mathbf{c}_k)}_{\text{second period.}} \end{aligned} \tag{4.33}$$

In the first period, the users transmit their signals as shown in equation (4.33). Due to the broadcast nature of the channels, the signals are simultaneously received both at the cooperating users and at the base-station receiver. At the same time, the received signals are independently processed at the users' receivers. Full duplex capability [78] is assumed available. The use of relays as in [80] can be considered likewise. The received signals at the cooperating pairs at this period are given in (4.30). The detection of signals at each other user node is performed by first obtaining the soft estimates of the signals by despreading the received signal with the known spreading sequence, for example the k^{th} user this is given by

$$z_k = \frac{1}{T_b} \int_0^{T_b} \mathbf{r}_k \mathbf{c}_k, \forall k. \tag{4.34}$$

Then, by performing channel phase correction and taking the sign of the real part of the signal z_k , the estimate of the k^{th} user's transmitted signal b_k' is obtained

$$b_k' = sgn\Big\{\Re\{z_k g_{ik}^*\}\Big\}, \tag{4.35}$$

where $sgn\{.\}$, $\Re\{.\}$ and * denote sign and real and complex conjugation operation, respectively.

During the second period, the cooperating users simply forward the detected data the partners b'_k to the base-station receiver using the their partners spreading sequences $\mathbf{c}_k$. It should be noted that the estimated data may not be identical to the transmitted data of the by originating users due to the detection errors in (4.34) and (4.35). The accuracy of detection and thus error performance improvement of the system due to the collaboration depends on the relative SNR gains of the inter-user channels g_{ki} and g_{ik} to their respective transmit channels to the base-station receiver g_{kd} and g_{id}. The processes (4.34) and (4.35) are performed at all the mobile nodes each acting both as a user and a partner for transmitting their data.

4.4.4 Collaborative SIC (C-SIC) Receiver

During both the first and second period of the proposed collaborative scheme, the base-station receiver processes the received signals to perform detection of users' data. In our proposed system, the C-SIC receiver performs the detection of user signals based on order of their estimated strength using the principle of a SIC, e.g. [65, 63, 53, 16]. Note that the collaborative matched filter (C-MF) scheme is obtained when there is only despreading and data detection is used i.e. no cancellation is performed. In the first period the signal estimation of the strongest among the users is carried out, followed by the cancellation of its MAI contribution from the remaining composite received signal. The relative power estimates of the users are generated at the output of the corresponding users' matched filters and the one with maximum is selected at a time given by

$$z_{\max} = \max\left\{\frac{1}{T_b}\int_0^{T_b} \mathbf{r}_d \mathbf{c}_k\right\}, \forall k. \tag{4.36}$$

In the second period the estimate for the strongest user denoted by index $\{\max\}$ from the first period is carried out from the out of bank of MF as follows:

$$z'_{\max} = \frac{1}{T_b}\int_0^{T_b} \mathbf{r}'_d \mathbf{c}_{max}. \tag{4.37}$$

The estimated signals of the user from the two periods is maximum ratio (MRC) combined to form a final decision statistic $Z_{\max}$. Note that other combining methods such as EGC, MMSEC [2] are equally applicable for the diversity combining. For MRC with coherent detection , the combined signal can be shown as follows:

$$Z_{\max} = z_{\max}\hat{\alpha}_{\max} + z'_{\max}\{\hat{\alpha}\}'_{\max}, \tag{4.38}$$

where $\hat{\alpha}_{\max}$ and $\{\hat{\alpha}\}'_{\max}$ are amplitudes of estimated strongest user in first period. Finally, the hard decision of data of the user with index $k = \max$ is performed from the SIC stage as follows:

$$\hat{b}_k = sgn\Big\{\Re\{Z_{\max}\}\Big\}. \tag{4.39}$$

The cancellation process now has to be applied to both the received signals from the first and second periods for improving the estimation of weaker users signal that follows the same processes (4.36) - (4.39). The estimates of the user's signal in the first and second period $z_{\max}$ and $z'_{\max}$ are separately spread using the spreading sequence of the detected user with index $k = \max$ and subtracted from the respective received signals $r_d(t)$, $r'_d(t)$ to obtained less interfered received signals as follows

$$\begin{aligned} \mathbf{r}_d &= \mathbf{r}_d - z_{max}\mathbf{c}_k \\ \mathbf{r}'_d &= \mathbf{r}'_d - z'_{max}\mathbf{c}_k. \end{aligned} \tag{4.40}$$

The processes (4.36) - (4.40) are then carried out until all users' data signals are detected.

4.4.5 Achievable Rate of the Collaborative SIC

The achievable rate analysis presented here for the proposed system is developed based on some well-known theories of capacity of single user channel [110], that of relay channels [78] and also the Gaussian Approximation method [36] used extensively in CDMA systems. The cooperation between two users is considered for simplicity, however the method applies to more than two users with appropriate modifications. The achievable rates for the cooperating users R_k and R_i can be shown as [7]:

$$\begin{aligned} R_k &= R_{kd} + R_{ki} \\ R_i &= R_{id} + R_{ik} \end{aligned} \tag{4.41}$$

where, R_{kd} and R_{id} are the rates using the k^{th} and i^{th} users' own channels to the transmitter. R_{ki} and R_{ik} are the additional rates that come from employing the cooperation scheme using the inter-user channels. Since, the CDMA systems often operate in an interference limited environment (i.e. assuming high SNR region with $P_k/T_bN_0 >> 1$), the capacity calculation has to be done not on the given SNR basis but on the signal to noise plus interference ratio (SINR) at the decision point of each user. The SINR expression Γ_{kd} for the k^{th} user's decision variable conditioned on its data b_k can be shown as follows:

$$\Gamma_{kd} = \frac{E^2\{z_k|b_k\}}{var\{z_k|b_k\}} = \frac{\alpha_{kd}^2}{\sum_{j\neq k}^{K} \rho_{jk}^2 \alpha_{jd}^2 + N_0} \tag{4.42}$$

where, $E\{.\}$ and $var\{.\}$ denote the expectation and variance of a random variable z_k and ρ_{jk}^2 is the power of cross correlation or MAI between users' sequences. We make a simplified assumption that the SIC perfectly cancels the interference at each stage to obtain the expected SINR $E\{\Gamma_{kd}\}$ at the detection point for k^{th} user. The SINR at the output of k^{th} stage (equi-correlated spreading sequences are assumed for simplicity with $\rho_{jk}^2 = \rho^2, \forall j, \forall k$) can be given by

$$\begin{aligned}\Gamma_{kd} &= E\left[\frac{N_0}{\alpha_{kd}^2} + \sum_{j=k+1}^{K} \rho^2 \alpha_{jd}^2\right]^{-1} \\ \Gamma_{id} &= E\left[\frac{N_0}{\alpha_{id}^2} + \sum_{j=i+1}^{K} \rho^2 \alpha_{jd}^2\right]^{-1}\end{aligned} \tag{4.43}$$

where, $\Gamma_0 = E\{\frac{\alpha_{kd}^2}{N_0}\}$ is the SNR for a given transmit power under single user condition and assumed equal for all users unless otherwise stated. Here, the Gaussian approximation can be used for MAI signal distributions [36] and when random spreading sequences are used, this allows us to approximate the variance of a user's MAI contribution to be $\rho^2 = 1/N$ for each user. To further simplify the analysis and to obtain the average SINR for k^{th} under the SIC, it is assumed that $k = K/2$ i.e., in average k^{th} user is detected after $(K-1)/2$ users are canceled from the order statistic used [53] and this leads to the SINR of k^{th} user using random sequence with SIC as

$$\Gamma_{kd} = \frac{\alpha_{kd}^2}{K\alpha_{jd}^2/2N + N_0} > 2 \quad for \quad K \longrightarrow N \tag{4.44}$$

Although the SINR calculation given above may not well characterize the true SIC performance, this approach may be useful to gain an insight into the upper and lower bounds on rates a practical SIC can give. A very useful study on achievable rate analysis of a non cooperative SIC with random sequences under different conditions such as different power ordering as well as power control, the effect of channel fading distributions are carried out in [88]. It was also noted that the well known Rayleigh distributed fading of channels have near optimum distribution for SIC with power ordering. The value of Γ_{kd} is maximized when distribution of interfering users' channel gains α_{jd} is such that

$$\begin{gathered}\Gamma_{kd} \Longrightarrow \Gamma_0 \\ if \sum_{j=k+1}^{K} \rho_{jk}^2 \alpha_{jd}^2 << \alpha_{kd}^2\end{gathered} \tag{4.45}$$

Also, it can be clearly seen from (4.45) that the value of Γ_{kd} is also effected by the type of sequences used i.e. cross correlation values ρ_{jk}. Thorough analysis of such system is very much involved and hence is not carried out here. Assuming interference free inter-user channel, the upper bound on the achievable rate of k^{th} user with the proposed cooperation scheme can be given by

$$R_k < E\left[C\Big\{\Gamma_{kd} + \min\{\Gamma_{ki}, \Gamma_{id}\}\Big\}\right] \tag{4.46}$$

where, $C(X) = \log_2(1 + X)$ is the well-known expression from the AWGN channel capacity theorem [110]. From the rate R_k achieved for k^{th} user as in (4.41), it can be justified that under the most practical cooperation conditions where $\Gamma_{ki} \geq \Gamma_{id}$ each user's achievable rate under the proposed Cooperative SIC scheme R_{c-sic}, is strictly higher than the rate that can be achieved from cooperation using MF (without interference cancellation) R_{c-mf} or SIC detection without cooperation R_{sic}. The expression for achievable rates for these schemes dependent on system loading, cross correlation and perfect MAI cancellation in the case of using SIC can also be shown as:

$$\begin{aligned} &R_{c-sic} > R_{c-mf} \Longrightarrow \\ &E\left[C\left[E\left\{\frac{N_0}{\alpha_{kd}^2} + \sum_{j=k+1}^{K} \rho^2 \alpha_{jd}^2\right\}^{-1}\right]\right] + \min\left[C\left[E\left\{\frac{\alpha_{ki}^2}{N_0}\right\}, E\left\{\frac{N_0}{\alpha_{id}^2} + \sum_{j=i+1}^{K} \rho^2 \alpha_{jd}^2\right\}^{-1}\right]\right] \\ &> E\left[C\left[E\left\{\frac{N_0}{\alpha_{kd}^2} + \sum_{j=1, j\neq k}^{K} \rho^2 \alpha_{jd}^2\right\}^{-1}\right]\right] + \min\left[C\left[E\left\{\frac{\alpha_{ki}^2}{N_0}\right\}, E\left\{\frac{N_0}{\alpha_{id}^2} + \sum_{j=1, j\neq i}^{K} \rho^2 \alpha_{jd}^2\right\}^{-1}\right]\right] \end{aligned} \tag{4.47}$$

$$\begin{aligned} &R_{c-sic} > R_{sic} \Longrightarrow \\ &E\left[C\left[E\left\{\frac{N_0}{\alpha_{kd}^2} + \sum_{j=k+1}^{K} \rho^2 \alpha_{jd}^2\right\}^{-1}\right]\right] + \min\left[C\left[E\left\{\frac{\alpha_{ki}^2}{N_0}\right\}, E\left\{\frac{N_0}{\alpha_{id}^2} + \sum_{j=i+1}^{K} \rho^2 \alpha_{jd}^2\right\}^{-1}\right]\right] \\ &> E\left[C\left[E\left\{\frac{N_0}{\alpha_{kd}^2} + \sum_{j=k+1}^{K} \rho^2 \alpha_{jd}^2\right\}^{-1}\right]\right] \end{aligned} \tag{4.48}$$

From above, it becomes clear that the rate increase due to cooperation can be maximized if technique of interference cancellation is designed as such that maximizes the SINR for each user. In reality, the SIC always generates residual error due to imperfect cancellation therefore reducing the achievable rates of the user cooperation scheme. The numerical results on achievable rates of the scheme employing different sequences will be presented in future work.

It is well known that the conventional SIC using correlators suffers from imperfect MAI estimation problem as system loading increases. This causes error propagation to later users' detection stages and the SIC may perform even worse than without cancellation [65]. The CMA based detection e.g. in [86, 58, 66, 59], is known to approach the performance of MMSE receiver under static channels conditions upon full convergence. The fading wireless channels brings many challenges and hence interference cancellation technique can be employed to significantly improve the performance in such conditions. It is shown in [16], that the problem of conventional SIC can be alleviated to great extent by improved SIC design that uses constant modulus (CM) property of user transmitted signals to blindly suppress MAI while estimating desired user's signal at the despreader output. Furthermore it uses a simple gradient descent based adaptive algorithm to update the estimates blindly within the SIC process and referred to here as BA-SIC. The scheme proposed here employs the BA-SIC within the cooperation diversity system framework of uplink CDMA.

As will be shown later, the performance limitations of collaborative scheme with conventional SIC can be significantly improved by employing blind adaptive (BA)-SIC and is described next.

4.4.6 Collaborative Blind Adaptive SIC (C-BASIC) Receiver

The principles and detailed algorithm of the BA-SIC technique is presented in Section 3.2 [16]. Here, the BA-SIC algorithm embedded within the collaborative diversity system is proposed. At the first symbol period, the weights of the CMA are initialized with user's spreading sequence $\mathbf{w}_k(1) = \mathbf{c}_k$ and $\mathbf{w}'_k(1) = \mathbf{c}_k$, respectively. Without loss of generality, it is assumed the first user (strongest among K users) to be detected is user 1. Similarly next strongest user is assigned an index as user 2 and so on. At the first stage, the received signal can be expressed as $\mathbf{r}_d(m) = \mathbf{r}_1(m)$ and $\mathbf{r'}_d(m) = \mathbf{r}_1(m)$, respectively. The remaining composite signal after cancellation at k stage for the detection of a user's and its cooperating pair signals are expressed as $\mathbf{r}^{k+1}(m)$ and $\mathbf{r'}^{k+1}(m)$, respectively. Below, the diversity combining, detection and interference cancellation procedures for k^{th} stage is described.

At stage k^{th}, the decision statistic $z_k(m)$ is obtained by multiplying chips of $\mathbf{r}_k(m)$ with the vector of weights $\mathbf{w}_k(m)$ and summed over the symbol period given by

$$\begin{aligned} z_k(m) &= \mathbf{w}_k^T(m)\mathbf{r}_k(m) \\ z'_k(m) &= \mathbf{w'}_k^T(m)\mathbf{r'}_k(m). \end{aligned} \tag{4.49}$$

The CMA criterion J_{CM} can be written as minimization of the following cost function

$$J_{CM} = E\{z_k(m)^2 - \gamma\}^2, \tag{4.50}$$

where $E\{.\}$ is the expectation operator, γ is the dispersion constant, which is equal to unity for binary phase shift keying (BPSK) signals. The instantaneous error signal $e_k(m)$ is calculated as

$$\begin{aligned} e_k(m) &= z_k(m)\{z_k(m)^2 - \gamma\} \\ e'_k(m) &= z_k(m)\{z'_k(m)^2 - \gamma\}. \end{aligned} \tag{4.51}$$

The estimated gradient vector of the error signal is then calculated by

$$\begin{aligned} \nabla_k(m) &= \mathbf{r}_k(m)e_k(m) \\ \nabla'_k(m) &= \mathbf{r'}_k(m)e_k(m). \end{aligned} \tag{4.52}$$

Using the gradient of (4.52), the weight vector at next symbol $\mathbf{w}_k(m+1)$ is updated as follows

$$\begin{aligned} \mathbf{w}_k(m+1) &= \mathbf{w}_k(m) - \mu\nabla_k(m) \\ \mathbf{w}'_k(m+1) &= \mathbf{w}'_k(m) - \mu\nabla'_k(m), \end{aligned} \tag{4.53}$$

where μ is the step-size that adapts the elements of the weight vector to minimize the cost function (4.50). The output signals $z_k(m)$ and $z'_k(m)$ are combined using MRC using combining weights i.e. the amplitude estimates of $\hat{\alpha}_k(m)$ and $\hat{\alpha}_i(m)$ and is delivered to the decision making process to perform hard decision

$$\hat{b}_k(m) = sgn\Big\{\Re\{z_k(m)\hat{\alpha}_k(m) + z'_k(m)\hat{\alpha}_i(m)\}\Big\}. \tag{4.54}$$

The cancellation process also requires amplitude estimate of the detected user signal and spreading. The estimates are obtained, using the weights of the CMA algorithm and the known spreading sequence as follows

$$\begin{aligned}\tilde{\alpha}_k(m) &= \frac{\breve{c}_k(m)}{\breve{w}_k(m)}\\ \tilde{\alpha'}_k(m) &= \frac{\breve{c}_k(m)}{\breve{w}'_k(m)},\end{aligned} \tag{4.55}$$

where $\breve{c}_k(m) = 1/N\sum|c_k(mN+n)|$ and $\breve{w}_k(m) = 1/N\sum|w_k(mN+n)|$, $n = 1, 2, ..N$, are the mean amplitude of user's spreading sequence chips and the mean of the weight vector updated by the CMA, respectively. The estimated symbol is then scaled with its new amplitude estimate $\tilde{\alpha}_k(m)$ and spread to generate the cancellation term as follows

$$\begin{aligned}\mathbf{x}_k(m) &= \tilde{\alpha}_k(m)z_k(m)\mathbf{c}_k(m)\\ \mathbf{x'}_k(m) &= \tilde{\alpha'}_k(m)z'_k(m)\mathbf{c}_k(m).\end{aligned} \tag{4.56}$$

The remaining composite signal after the interference cancellation is

$$\begin{aligned}\mathbf{r}_{k+1}(m) &= \mathbf{r}_k(m) - \mathbf{x}_k(m)\\ \mathbf{r'}_{k+1}(m) &= \mathbf{r'}_k(m) - \mathbf{x'}_k(m).\end{aligned} \tag{4.57}$$

The processes (4.49)-(4.57) are repeated for each stage until the weakest user is detected. The sum rate analysis of C-BASIC can be carried in a similar manner given subsection in 4.4.5 and will be studied in details in future work.

4.4.7 Simulation Results and Comparisons

A model of K user synchronous uplink DS-CDMA system employing BPSK and short binary Gold sequences of length N = 31 is used. The channel used is Rayleigh flat fading channel with Doppler shift of 185Hz corresponding to $f_dT_b = 0.003$.

First, in Figure 4.12, the BER performance of a single users under two cases namely, No Collaboration and Perfect Collaboration are presented. The No Collaboration is the conventional method of transmitting data where users do not relay each other's data. The Perfect Collaboration is the case when the user's decoded and forwarded data from its partner is identical to the previous transmitted data from the user due to very high SNR gains of inter-user channels. These results will be used as the references in the sequel to compare the performance of various collaborative SIC receivers.

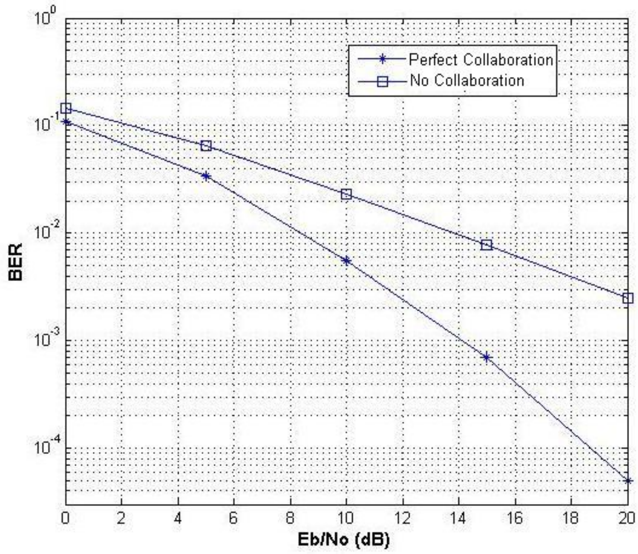


Figure 4.12: Performance of relay assisted collaborative diversity in flat Rayleigh fading channels under different cooperation conditions

4.4.7.1 C-SIC

Figure 4.13 shows the BER performance of C-SIC under the system load of $K = 10$ users. The performance of collaborative matched filter (C-MF) is also shown for the comparison. As expected, the performance of C-SIC is improved significantly compared with C-MF. It can be seen from the figure, the use of interference cancellation provides significant BER gain and approaches the performance of a single user system with Perfect Collaboration under higher ratios of inter-user channel SNR conditions as described earlier.

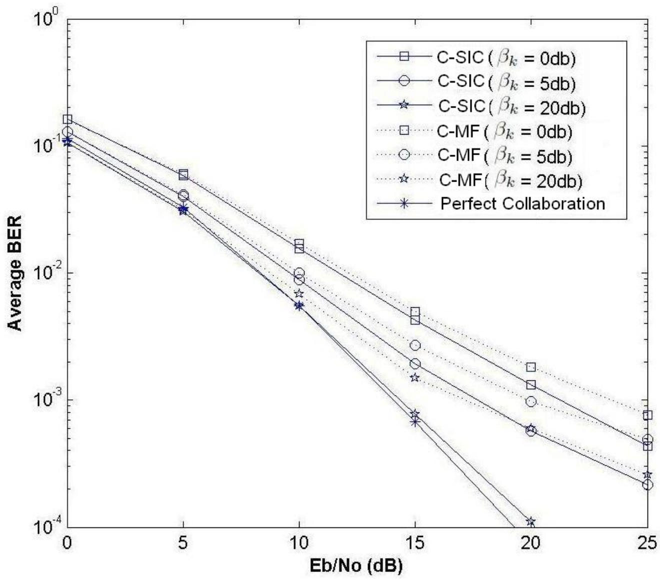


Figure 4.13: Performance of C-SIC in flat Rayleigh fading channels for K=10 and Gold sequences of N=31

The BER performance of C-SIC under higher system load of $K = 20$ users is shown in Figure

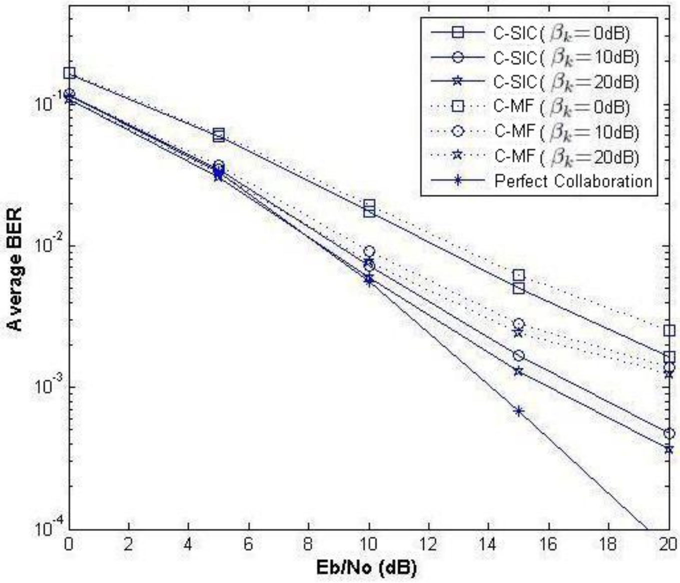


Figure 4.14: Performance of C-SIC in flat Rayleigh fading channels for K=20 and Gold sequences of N=31

4.14. The BER of C-MF is shown for comparison. The performance C-SIC is significantly better compared with the C-MF as expected. The effect of system loading on degradation in BER for both systems is clearly seen from the figure. The performance gaps between the C-MF and C-SIC are still considerable, however it is noted both systems suffer for residual MAI and hence BER curves are not steep at high SNR region.

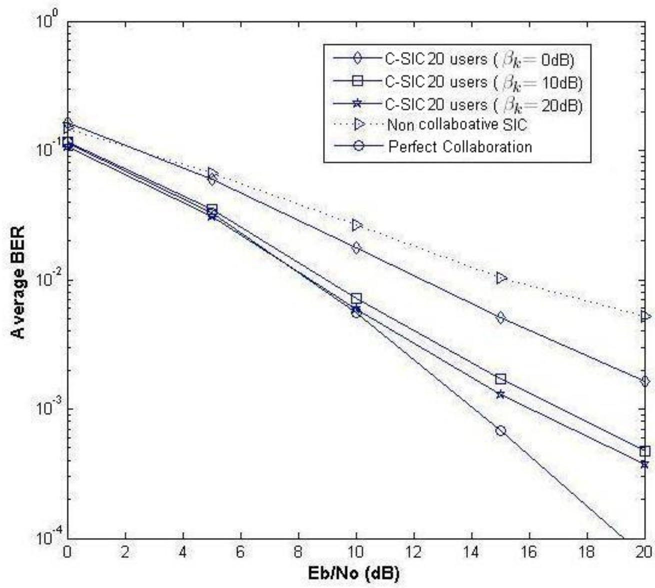


Figure 4.15: BER vs. number of users of C-SIC in flat Rayleigh fading channels with $E_b/N_0 = 20$dB, and Gold sequences of N=31

The BER performance of C-SIC is shown in Figure 4.15 and compared with SIC without collaboration (Non Collaborative SIC) for 20 users under various degrees of user collaboration. Which are quantified by the relative SNR gains of inter-user channels to the transmit channels of

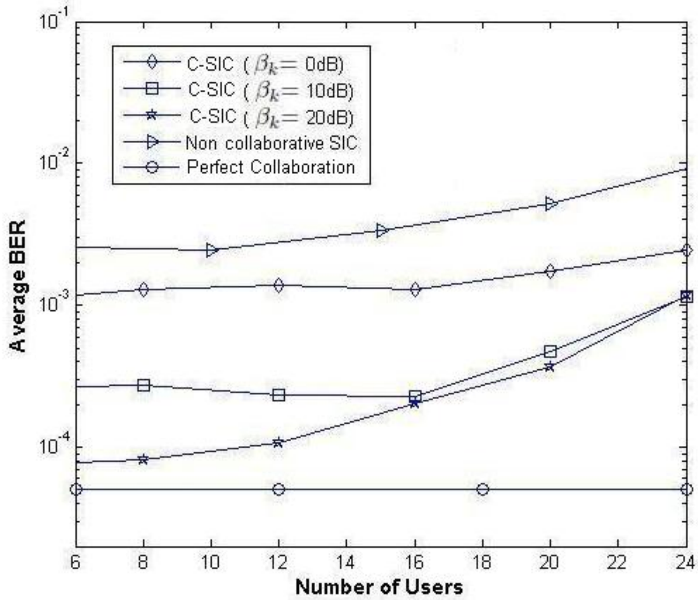


Figure 4.16: BER vs. number of users of C-SIC in flat Rayleigh fading channels with $E_b/N_0 = 20$dB and Gold sequences of N=31

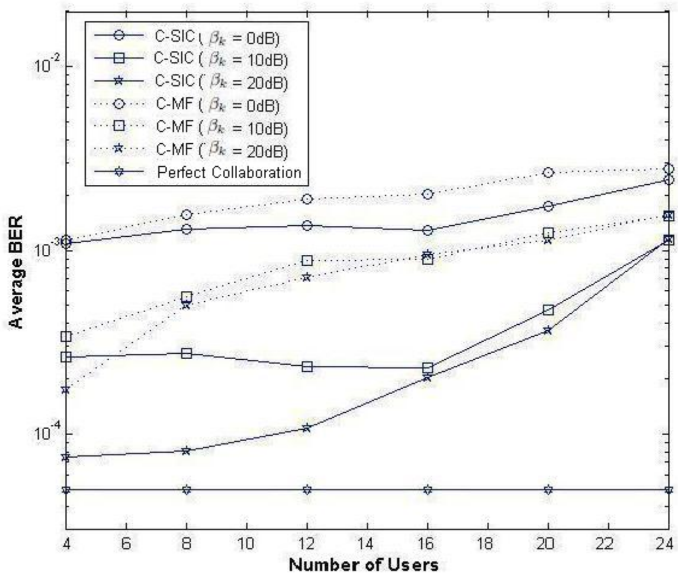


Figure 4.17: Performance of C-SIC and C-MF in flat Rayleigh fading channels with $E_b/N_0 = 20$ dB and Gold sequences of N=31

the respective users. The C-SIC showed a superior BER performance giving several dB of gain for a given target BER. For example, with relative SNR gain of inter-user channels of only 0 dB, it shows gain of about 4 dB for the same target BER of 10^{-3}. As the quality of inter-user channels improve (higher values of β_k), the error performance of C-SIC improves rapidly. It is noted that the relative inter-user channel gain of 10dB is sufficient to provide the near asymptotic gain from the user cooperation and the performance is very near to that of the system with inter-user SNR gain of 20 dB. The gap in BER performances between the Perfect Collaboration and C-SIC is due to imperfect interference generation and cancellation.

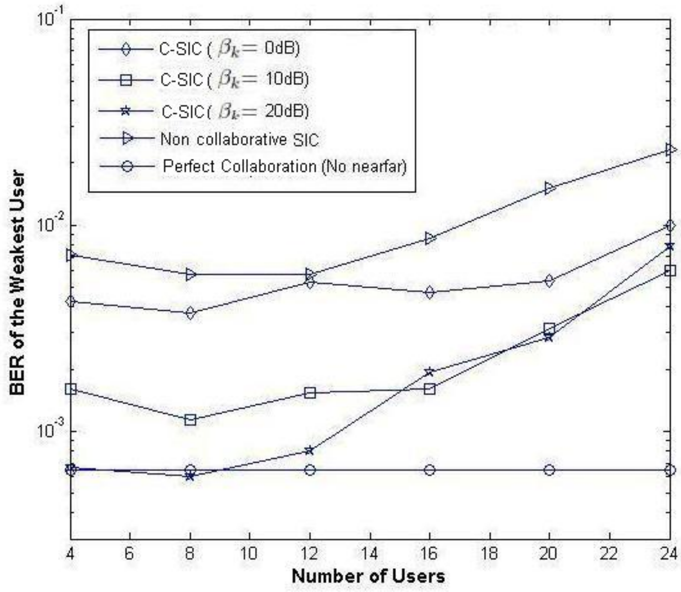


Figure 4.18: Performance of C-SIC in flat Rayleigh fading channels with near far ratio of 10 dB with E_b/N_0 of the weakest user=15dB and Gold sequences of N=31

In Figure 4.16 BER simulation results of C-SIC is plotted under different system loads in fading channel environments. As expected, it shows significant improvement in the error performance compared to the Non collaborative SIC as the inter-user channel gains β_k increase. Also, the simulation performance results of C-SIC show that the BER performance under different user loading conditions does not degrade significantly to unacceptable level as the number of users increase. The performance of C-MF as shown in Figure 4.17 , is much worse than the C-SIC. Also, it is noted that the BER performance of C-MF do not improve much with higher SNR gains of inter-user channels under higher user loading conditions.

Figure 4.18 shows the performance of the C-SIC in fading channels and nearfar ratio of 10 dB. The desired user (weakest user) has unity power corresponding to the expected SNR at the receiver of 15 dB, while all other users have been assigned powers uniformly distributed between 0 and 10dB. It can be clearly seen from the figure that as β_k increases, the BER performance of the desired user with the C-SIC is much improved compared to that of SIC without collaboration.

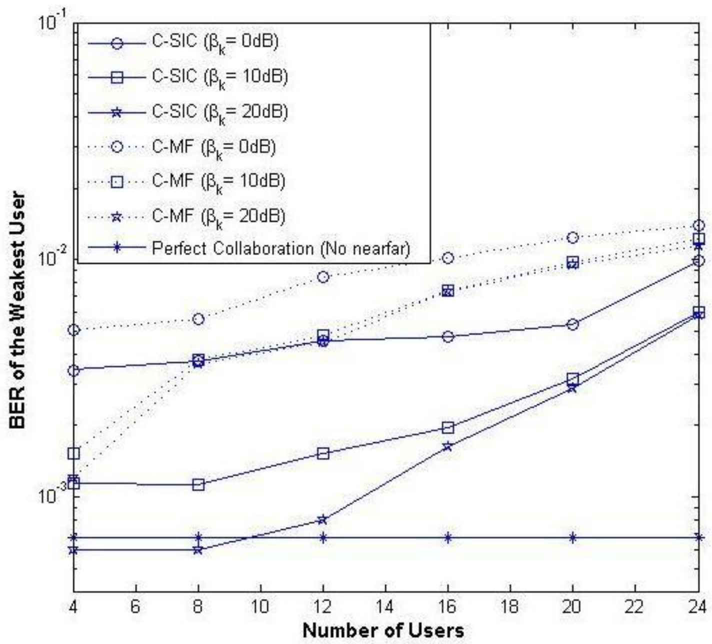


Figure 4.19: Performance of C-SIC in in flat Rayleigh fading channels with nearfar ratio of 10dB with E_b/N_0 of the weakest user=15dB and Gold sequences of N=31

As the number of users increases, the system BER performance is degraded as usual. Due to the collaboration, it is expected that the desired weak user benefits from the strong channel of its partner. From Figure 4.18, it can be noted that the performance of the desired user with C-SIC is indeed improved compared to the case of equal power users case as in Figure 4.16. This improved result can also be verified using the analytical approach and is left as the future study.

The comparison of collaborative schemes, C-SIC and C-MF under 10 dB nearfar conditions is shown in Figure 4.19. It is observed that use of C-MF does not improve BER with nearfar conditions as it did with C-SIC. The performance of C-MF based scheme is shown to be worse than in equal power case as in Figure 4.17. The main reason behind this is that, in nearfar conditions the estimates of users' data becomes highly unreliable and hence diversity combining does not assist much to improve the performance of final decision.

4.4.7.2 C-BASIC

In Figure 4.20, BER simulation results of the proposed Cooperative BA-SIC (C-BASIC) system is plotted under system loading of $K = 20$ users. The BER of Non cooperative SIC and C-SIC under the same system conditions are also shown. The C-BASIC that uses blind adaptive approach for suppressing and canceling MAI, improves the amplitude and data estimates of users' and that of cooperative signals, much improved BER at high system load is expected. As expected, the proposed technique shows noticeable improvement in the error performance compared to the Co-operative SIC (C-SIC) receivers under the same system settings as the β_k increases: for example, a SNR gains of ≈ 2 dB for a BER of 10^{-3} can be seen here.

The BER performance of the proposed C-BASIC is shown in Figure 4.21 and compared with C-SIC for system loading of $K = 6 - 24$ users under fixed $E_b/N_0 = 20$ dB. The degree of cooperation is quantified by the relative SNR gains of inter-user channels to the transmit channels of the respective users i.e. β_k. The C-BASIC shows a superior BER performance under high system loading conditions, which is expected due to improved estimation of users' data signals and amplitudes of the blind adaptive approach to MAI cancellation.

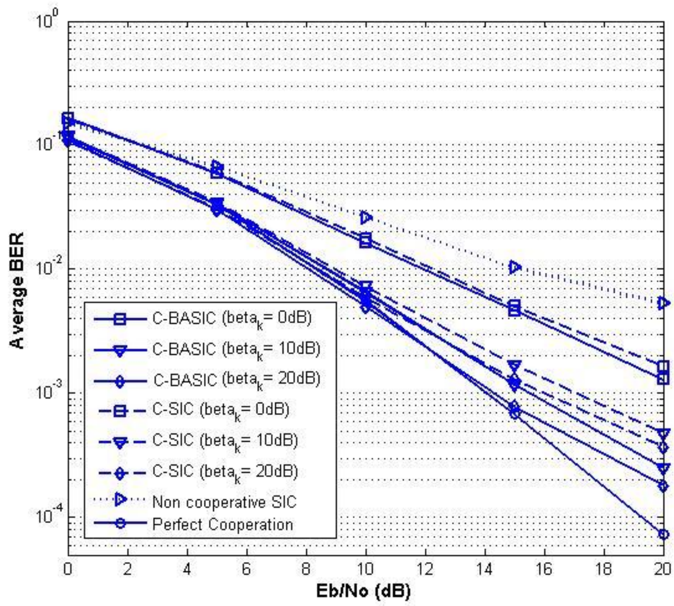


Figure 4.20: Performance of C-BASIC in flat Rayleigh fading channels for K=20 and Gold sequences of N=31

The impact of nearfar conditions on the BER of C-BASIC is shown in Figure 4.22. System loading of $K = 6 - 24$ is considered, where the desired weakest user has unity power while all other users' signals are received at power level uniformly distributed between $0 - 10$ i.e. $\Omega = 10$ dB. The E_b/N_0 of the desired weakest user is assumed to be 20 dB. The step-size of adaptive algorithm is set as $\mu = \frac{0.0001}{\Omega}$. The BER of the user is plotted against different system loading i.e. the number of users and also under different ratio of inter-user channel SNR gains β_k. It can be observed that the performance of C-BASIC is robust in nearfar conditions and high system loading of up to 18 users. The performance of C-SIC shows similar gain, however the degradation in BER starts with much lower system loading of 12 users compared with C-BASIC. Also it is noted that C-SIC does not benefit much from strong channel SNR of the partners' compared with C-BASIC. The reason for this is the conventional SIC while removes part of MAI also suffers from unreliable estimation of weaker users' signals and hence only the fraction of diversity gain achieved. The results for C-BASIC and C-SIC on the other hand are in strong contrast with Cooperative MF (C-

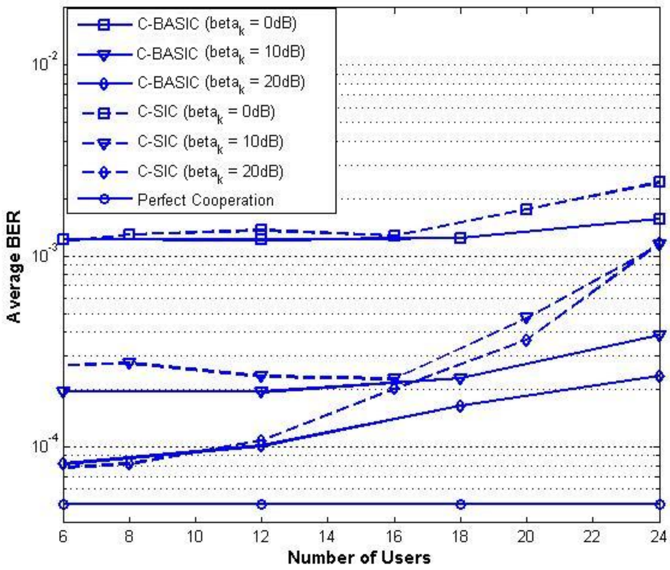


Figure 4.21: BER vs. number of users of C-BASIC in flat Rayleigh fading channels with $E_b/N_0 = 20dB$ and Gold sequences of N=31

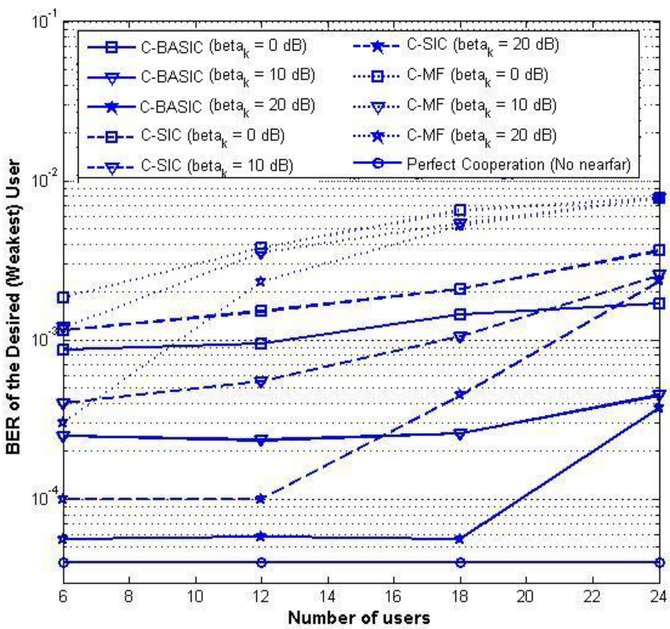


Figure 4.22: BER vs. number of users of C-BASIC in flat Rayleigh fading channel with nearfar ratio of 10dB, E_b/N_0 of the weakest user=20dB and Gold sequences of N=31

MF) detection technique, where the BER of weakest user degrades rapidly even under very small number of users.

It is noted that collaborative diversity and SIC can be used jointly to achieve high performance in uplink CDMA. An important problem of collaborative diversity is the loss of bandwidth efficiency due to requirement of orthogonal time/frequency slots for forwarding partners' data from cooperating nodes. Can we achieve full rate transmission also with full diversity from the user collaborative scheme? This question is the main driver for the work presented in next section.

4.5 Near-Unity-Rate Collaborative Diversity Scheme for Multiple Access Fading Channels

4.5.1 Introduction

The cooperative or collaborative communication in multi terminal wireless networks is a well investigated topic [7, 8, 75, 79, 84, 18]. The cooperation scheme exploits the inherent spatial diversity of distributed nodes to form so called virtual antenna array and hence improves the outage capacity and throughput significantly. Most of the work on cooperative communication [79, 84, 83, 75, 8] have focused on single user point to point link with relay nodes assisting the source by retransmitting its data to the destination node. The seminal work by Sendonaris et. al. on user cooperation diversity considered the scheme in the context of multiple access channels and in particular to CDMA. The objective of user cooperation in such scenario is different from node to node cooperative relaying [79, 84, 75, 8], because the users who are transmitting their data, must also serve as relay for their partners. Hence the main problem is, the full duplex operation is necessary and this requires some form of echo cancellation [78, 7] to remove unwanted replica of the user's own transmitted signal to the incoming signal. Despite the use of this technique, the cooperation diversity in [7] is achieved at a loss in bandwidth efficiency.

There are some interesting work tackling this issue of providing bandwidth (BW) efficient collaborative diversity in multiple access channels such as [111, 80, 76]. The opportunistic multipath scheme using multiple idle users (relays) proposed by Rebiero attempts to capture the benefit of spread spectrum signals in resolving the multipath and hence significant part of diversity gain is shown to be achieved with this method. The multicode version of the work with single relay for each user is also presented. The scheme achieves the aforementioned gain when there are two idle relays around each user and statistically orthogonal PN sequences [38] are employed. A scheme using quadrature signalling within TDMA framework for cooperative communication is proposed in [111]. The scheme essentially makes use of two signalling dimensions available at the cooperating nodes to compensate for the loss in bandwidth efficiency of the cooperation and hence may potentially loose the maximum theoretical spectral efficiency by half. The scheme proposed by Verdhe [76] is another method of achieving bandwidth efficient multiple access in CDMA systems by exploiting the multicode transmission at cooperating nodes and linear preprocessing and multiuser detection at the base-station. A superposition coding assisted cooperative multiple access (CMA) scheme is proposed in [112], where the relaying user superimposes its own data along with source users data. It assumes full knowledge of channels and their precoding matrix at all user nodes and destination. This requires closed loop operation with feedback channels, leading to a more complex system and some loss in throughput. Space-Division Relay (SDR) is proposed in [113] that allows two users to exchange their data in first and second periods, and in the third period, space division multiplexing rather than time division is used for simultaneous relaying of users' data. It shows improved rate of $\lambda = 2/3$ compared with $\lambda = 1/2$ e. g. in [7].

In this section, a full rate $\lambda = 1$ cooperative diversity scheme referred to as bandwidth efficient collaborative diversity (BECD) is proposed and evaluated for the uplink of CDMA system. Our main idea is the use of the collaborative transmission approach presented in [46] so as to permit the pairs of users and their relays to communicate simultaneously using the same available resources. However, unlike in [46] no form of collaborative codes is used here. Instead, the differences in fading channel magnitudes of the co-channel users and their relays are exploited to jointly detect the data using a simple maximum-likelihood (ML) algorithm based on calculating the Euclidian distance between the received signal and all possible data combinations.

The Section is organized as follows. Two possible system approaches for the implementation of the proposed BECD scheme are discussed in 4.5.2. The proposed scheme without relays is described in subsection 4.5.3 and with relays in subsection 4.5.4. The performance results are presented in subsection 4.5.5.

4.5.2 System Approaches

Below, the two possible implementation approaches of the proposed BECD scheme are discussed briefly:

BECD without Relays: This approach assumes that there are no relay users available in the system and hence the users themselves act both as the sources and relays as assumed by Sendonaris et.al. in [7]. Therefore it is assumed that the users are able to transmit and receive simultaneously (full duplex mode).

BECD with Relays: This assumes that for each user there is a relay node. This setting allows for more practical implementation of the proposed scheme since the users take turn while transmitting and receiving (half duplex mode) [8, 114, 80].

To clarify the main differences in the description of signals for both approaches and possible performance results, following notes are made here:

1. In the first approach, single inter-user channel g_k is used and is assumed symmetric for simplicity. For the second approach, there exist user-relay channels for each user denoted as $g_k^l, l \in \{1, 2\}$. The statistics of the inter-user and user-relay channels for the both cases can be considered same, i.e. these channels are modeled as complex Gaussian random variables $\mathcal{CN}\{0, 1\}$ with Rayleigh distributed amplitudes. Therefore, regardless of these signal notational differences, the final BER performance of both schemes are expected to be

very similar.

2. In the latter scheme employing relays, each relay node will receive the transmitted signals from both collaborators and hence perform joint detection the users' transmitted signals. The error performance of joint ML detection scheme for co-channel users is about 2 dB inferior to that of the single user detection [82]. Despite this slight degradation and additional processing, the final BER performance of the collaborative scheme with relays remains very similar to that of the scheme without relays.

3. In the former, each user is assumed to be able to transmit and receive as the same time in both periods. With the latter scheme, in the first period, the users transmit and relays receive; in the second period, the users do not transmit but the relays forward the detected signals from its paired user in the previous period.

4.5.3 BECD without Relays

A multiuser communication scenario of an uplink synchronous CDMA with a pair of collaborating BPSK modulated users $kl, l \in \{1, 2\}$ within k^{th} group sharing the common spreading sequence $\mathbf{c}_k$ and the base-station receiver system employing the former scheme is shown in Figure 4.23.

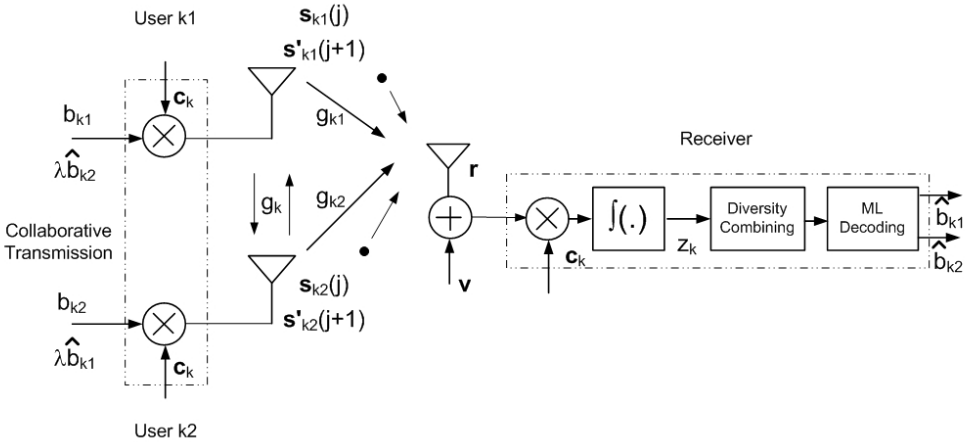


Figure 4.23: System model of BECD without Relays for uplink of CDMA

During the first period, the users transmit signals $\mathbf{s}_{kl}(j)$ using the common sequence $\mathbf{c}_k$ using their own data b_{kl}. At the same time, the pairing users $i \neq l \in \{1, 2\}$ receive and decode the transmitted signals. The received signals $\mathbf{r}_{ki}(j)$ at the first period j can be written as:

$$\mathbf{r}_{ki}(j) = \sqrt{P_{kl}/L} g_k \mathbf{s}_{kl}(j) + \sum_{u=1, u\neq k}^{G} \sum_{l=1}^{L} \sqrt{P_{ul}/L} g_{ul}^{k} \mathbf{s}_{ul}(j) + \mathbf{v}(j), \tag{4.58}$$

where P_{kl} is the power and L the number of partners per group is used for normalization, $s_{kl}(j) = b_{kl}\mathbf{c}_k$ is the transmitted signal of kl^{th} user, b_{kl} is data signal taking values $[-1, +1]$ with equal probabilities and with period T_b. The sequence $\mathbf{c}_k$ consists of antipodal chips each with period T_c and normalized power over the symbol period $\int_0^{T_b} c_k(t)^2 dt = 1$. The spreading factor is $N = T_b/T_c$, g_k and g_{ul}^k are Rayleigh fading inter-user channel (the two way inter-user channels are assumed symmetric [7]) and channels from users within $1 \leq u \leq G, u \neq k$ groups, respectively which remain constant over the two symbol period), and **v** is the AWGN with two sided power spectral density $N_0/2$. The second term in (4.65) can also be seen as multiuser interference term with non-zero variance when non-orthogonal sequences are used [5].

The detection of partner's data b_{kl} at each pairing user node $ki, i \in \{1, 2\}$ is performed by first obtaining the soft estimates of the signals by despreading the received signal with sequence $\mathbf{c}_k$, given by $z_{kl}(j) = \int_0^{Tb} \mathbf{r}_{ki}(j)\mathbf{c}_k^T; 1 \leq k \leq G$; where $\mathbf{c}^T$ denotes a transpose of vector **c**. Full duplex capability is assumed available at user terminals as in [7] for simplicity, however the use of additional relays as in [80] can also be used easily for a more practical half duplex operation. The data estimate $\hat{b}_{kl}$ is obtained as $\hat{b}_{kl} = sgn\Big[\Re\big\{z_{kl}(j)g_k^*\big\}\Big]$; where $sgn[.]$, $\Re\{.\}$ and * denote sign, real and complex conjugation operation, respectively. The estimated data $\hat{b}_{kl}$ is then spread using the group spreading sequence $\mathbf{c}_k$ to form transmitted signals in the second period as follows:

$$s'_{ki}(j+1) = \lambda \hat{b}_{kl}\mathbf{c}_k, \tag{4.59}$$

where $\lambda \in \{1, -1\}$ is a factor that is dependent on the specific transmission scheme. The received composite signal at the base-station receiver from all users' transmissions during the first period $\mathbf{r}(j)$ and from that of the partnering users' in the second period $\mathbf{r}(j+1)$ can be written as:

$$\begin{aligned} \mathbf{r}(j) &= \sum_{k=1}^{G}\sum_{l=1}^{L}\sqrt{P_{kl}/L}\,g_{kl}\mathbf{s}_{kl}(j) + \mathbf{v}(j), \\ \mathbf{r}(j+1) &= \sum_{k=1}^{G}\sum_{i\neq l}^{L}\sqrt{P_{ki}/L}\,g_{ki}\mathbf{s}'_{ki}(j+1) + \mathbf{v}(j+1), \end{aligned} \tag{4.60}$$

where g_{kl} is the Rayleigh fading uplink channel of kl^{th} user and assumed constant over the two symbol period. The structure of the signal $\mathbf{s}'_{ki}(j+1)$ transmitted during the second period is the same as $\mathbf{s}_{kl}(j)$, but the data are originated from the $ki^{th}, i \neq l$ pairing user terminal with kl^{th} user's estimated data in the j^{th} period $\hat{b}_{kl}$ and transmitted in the next $(j+1)^{th}$ period via its uplink channel g_{ki}.

Assuming average noise variances of all users' and the base-station receivers are equal, the relative SNR gain in dB of inter-user channels β_k compared to the respective uplink channels can be expressed as

$$\beta_k = \frac{E\{|g_k|^2\}}{E\{|g_{kl}|^2\}} = \frac{E\{|g_k|^2\}}{E\{|g_{ki}|^2\}}, \tag{4.61}$$

where $E\{.\}$ is the expectation operator. Having introduced with basic signal and channel models, we describe the operation of collaborative transmit diversity schemes next.

Two schemes for the collaborative diversity are presented next. The first scheme is an uncoded scheme where the signals from users sharing the same sequence $\mathbf{c}_k$ are received as a simple superposition and the signals are jointly detected and decoded. The second scheme employs distributed space-time coding of users' data and corresponding detection and decoding at the receiver.

4.5.3.1 Signal Superposition Based BECD

Under this scheme, In the first period, the users transmit their own data $\{b_{kl}, b_{ki}\}$. During the second period $(j+1)$, the cooperating users simply forward the detected data the partners in the previous period $\hat{b}_{kl}$ and $\hat{b}_{ki}$ to the base-station receiver using the same spreading sequence $\mathbf{c}_k$. Based on the above system model, a signalling structure of the proposed scheme with two partnering users and spanned over two consecutive symbol period $\{j, j+1\}$ can be shown in a matrix form $\mathbf{B}_k$ below.

$$\mathbf{B}_k = \begin{bmatrix} b_{k1} & b_{k2} \\ \hat{b}_{k2} & \hat{b}_{k1} \end{bmatrix},$$

where the rows and columns indicate time periods and signals from users $\{k1, k2\}$, respectively. It should be noted that the estimated data $\{\hat{b}_{kl}, \hat{b}_{ki}\}$ may not be identical to the transmitted data of the originating users due to the decoding errors in partners' receivers.

The base-station receiver first obtains soft estimates of composite data signals for decoding the symbols. This is achieved by first performing CDMA despreading of $\{\mathbf{r}(j), \mathbf{r}(j+1)\}$ using the k^{th} group's spreading sequence $\mathbf{c}_k$ to obtain the soft data signals $\{z_k(j), z_k(j+1)\}$ shown as follows:

$$\begin{aligned} z_k(j) &= \int_0^{Tb} \mathbf{r}_k(j)\mathbf{c}_k^T; 1 \le k \le G \\ z_k(j+1) &= \int_0^{Tb} \mathbf{r}_k(j+1)\mathbf{c}_k^T. \end{aligned} \tag{4.62}$$

Since $\{z_k(j), z_k(j+1)\}$ consist copies of data signals of both collaborators $\{b_{kl}, b_{ki}\}$ via different uplink channels $\{g_{kl}, g_{ki}\}$, either joint ML detection and decoding or space-time decoding is performed to extract the final estimates of users' data. It is noted that, the latter scheme is particularly more appealing in terms of detection complexity, as the users' data signals are detected separately due to the orthogonal design of space-time codes [73]. Although, the use of error correction techniques such as CRC codes as in [80, 76] can also be used, we do not consider any form of error correction here. This is because we are more interested in the effects of the channels on the diversity gain of the scheme to gain insight in to the problem and possible incorporation of coding for future study.

After the soft despread signals $\{z_k(j), z_k(j+1)\}$ are obtained, the receiver performs joint detection and ML estimation of users' transmitted data signals. The squared Euclidian distance metrics of $\{z_k(j), z_k(j+1)\}$ for each combination of possible data vectors along with uplink channels in the first period $\{g_{k1}, g_{k2}\}$ and second period $\{g_{k2}, g_{k1}\}$ are used. The receiver obtains the final estimates of users' data $\hat{b}_{k1}, \hat{b}_{k2}$ from the list of possible data vectors $\mathbf{b}_q = \{b_{1q}, b_{2q}, .., b_{Lq}\}, 1 \leq q \leq Q$ and the one that minimizes the sum distance measure is selected as the transmitted data

$$\{\hat{b}_{k1}, \hat{b}_{k2}\} = \arg\min_{\mathbf{b}_q \in \mathbf{B}} \Big[\big|z_k(j) - \sum_{l=1}^{L} g_{kl} b_{lq}\big|^2 + \big|z_k(j+1) - \sum_{i=1}^{L} g_{ki} b_{lq}\big|^2\Big], \tag{4.63}$$

where **B** is the list of all possible $Q = 2^L$ vectors of symbols for BPSK user data signals.

4.5.3.2 Space-Time Coding Based BECD

The transmission scheme we propose here is based on the simple Alamouti space-time codes [73] and is presented in matrix form as follows:

$$\mathbf{B}_k = \begin{bmatrix} b_{k1} & b_{k2} \\ -\hat{b}_{k2} & \hat{b}_{k1} \end{bmatrix},$$

where the rows and columns indicate time periods and transmitted signals from users $\{1, 2\}$, respectively. The received signals are given in (4.67), where uplink channels of users are assumed constant over the two symbol period. The final error performance of both schemes depend on how accurate the estimates $\hat{b}_{kl}$ are, which in turn are dependent on the relative SNR gain of the inter-user channels β_k. The above described transmission and detection processes are performed at all pairing $1 \leq k \leq G$ users' terminals.

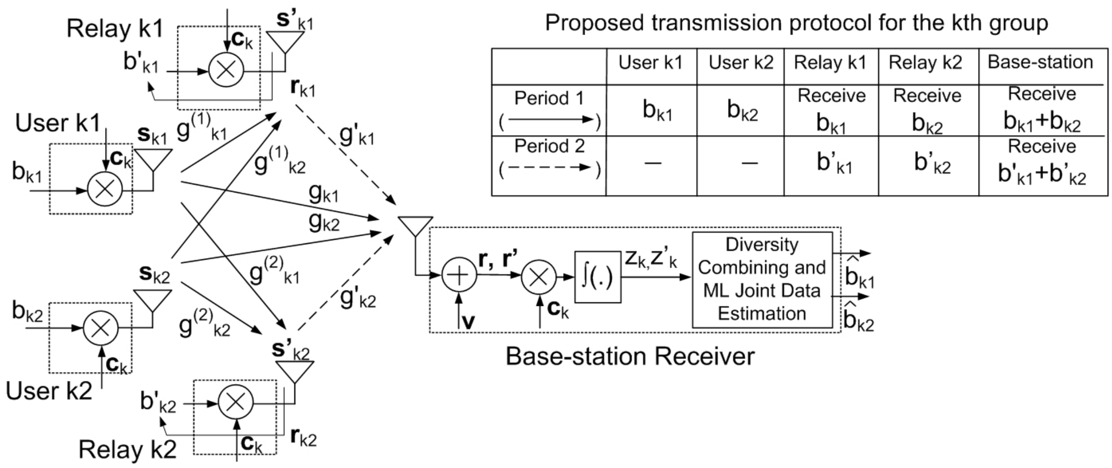
At the receiver, despreading is first performed as given earlier in (4.62). The signals $\{z_k(j), z_k(j+1)\}$ are processed here following the combining method in [73], which ensures that the data signals of users $\hat{b}_{kl}, l \in \{1, 2\}$ are detected separately and the copies are maximum ratio combined by the following process:

$$\begin{aligned} \hat{b}_{k1} &= sgn\Big[\Re\{z_k(j) g_{k1}^* + z_k^*(j+1) g_{k2}\}\Big] \\ \hat{b}_{k2} &= sgn\Big[\Re\{z_k(j) g_{k2}^* - z_k^*(j+1) g_{k1}\}\Big]. \end{aligned} \tag{4.64}$$

It is expected that under sufficiently high β_k, this scheme achieves full diversity gain and approaches that of the Alamouti scheme.

4.5.4 BECD with Relays

We consider two collaborating users i. e. $L = 2$ and two corresponding relays within each group. Each group is assigned only one spreading code $\mathbf{c}_k$ as illustrated in Figure 4.24 for the k^{th} group. The transmission protocol is also shown in the table within the Figure 4.24. In the first period, the users $k1$ and $k2$ transmit their data b_{k1} and b_{k2} using $\mathbf{c}_k$. The data are simultaneously received at the relays $k1$, $k2$ and the base-station via different fading channels denoted by



Proposed transmission protocol for the kth group

	User k1	User k2	Relay k1	Relay k2	Base-station
Period 1 (⟶)	b_{k1}	b_{k2}	Receive b_{k1}	Receive b_{k2}	Receive $b_{k1}+b_{k2}$
Period 2 (- - - -➝)	—	—	b'_{k1}	b'_{k2}	Receive $b'_{k1}+b'_{k2}$

Figure 4.24: System model of the proposed 2-user 2-relay

$\{g_{k1}^{(1)}, g_{k2}^{(1)}\}$, $\{g_{k1}^{(2)}, g_{k2}^{(2)}\}$ and $\{g_{k1}, g_{k2}\}$, respectively. At the same time, the relays independently perform CDMA despreading followed by joint detection of the co-channel signals based on the ML criterion or the minimum Euclidian distance measure over small number of data combinations to give the estimates b'_{k1} and b'_{k2}. In the second period, the relay $k1$ forward b'_{k1} via its uplink channel g'_{k1} and the relay $k2$ forward b'_{k2} using its channel g'_{k2} to the base-station also using the same code $\mathbf{c}_k$. The base-station performs despreading to give the output signals z_k and z'_k from the first and second period, respectively. The signals are then diversity combined and joint detection of the co-channel data is carried out using the ML or the minimum distance criterion over all possible data combinations to give $\hat{b}_{k1}$ and $\hat{b}_{k2}$. These processes ensures that each user's data experience two independent fading channels due to the repeated transmission from the users and their own relays. However there is a small price to be paid in terms of signal to noise ratio (SNR) compared with single user detection, owing to reduced distance due to detection in joint manner of the co-channel signals. Note that the independence of fading channels is easily satisfied in a multiuser fading environment; however the proposed scheme may also operate under fading correlation conditions though with some BER degradation. Channel estimation is required here at the receivers only unlike in [112], where it is required at both transmitter and receiver sides.

Transmission Protocol: The transmission protocol consists of two periods and processes as follows:

Period 1:- User Data Transmission: The transmitted signals in this period (shown as solid-line arrows in Figure 4.24) from the kl^{th} user $\mathbf{s}_{kl}$ is given by: $\mathbf{s}_{kl} = \sqrt{\frac{P_{kl}}{2}} b_{kl}\mathbf{c}_k$, where P_{kl} is the power, which is normalized with $L = 2$ so that the proposed scheme requires the same total power as direct transmission scheme for later comparisons, b_{kl} is BPSK data of period T_b, $\mathbf{c}_k$ is a unit norm spreading code of antipodal chips each with period T_c. The spreading factor is given by

$N = T_b/T_c$. The signal received at relays $\mathbf{r}_{kl}, l \in \{1,2\}$ is given by

$$\mathbf{r}_{kl}=\sum_{l=1}^{L=2} g_{kl}^{(l)}\mathbf{s}_{kl} + \sum_{u=1,u\neq k}^{G}\sum_{l=1}^{L=2} g_{ul}^{(l)}\mathbf{s}_{ul} + \mathbf{v} \tag{4.65}$$

where G is the number of groups, $g_{kl}^{(l)}$ is Rayleigh flat fading channel gain from the kl^{th} user to its kl^{th} relay and $\mathbf{v}$ is the AWGN noise. It should be noted here the scheme can be extended to other multiple access methods e.g. TDMA and O/FDMA by carrying out the transmission over time slots or frequencies instead of codes.

Decoding at Relays: The detection of kl^{th} user's data b_{kl} at its relay (shown just as an arrow underneath the relay in Figure 4.24 for simplicity) is performed by first CDMA despreading of the received signal $\mathbf{r}_{kl}$ with code $\mathbf{c}_k$ to give a soft signal $z_{kl} = \int_0^{Tb}\{\mathbf{r}_{kl}\}^T\mathbf{c}_k$, where $\{.\}^T$ is the transpose operation. The effects of signals from other groups in (4.65) remain zero at the despreader output due to the use of orthogonal codes between them. The estimate b'_{kl} is obtained by searching over Q possible data combinations and selecting the one that minimizes the distance measure as follows

$$b'_{kl} = \arg \min_{b_l^q\in\{\mathbf{b}_1,.,\mathbf{b}_q,.,\mathbf{b}_Q\}} \left| z_{kl} - \sum_{l=1}^{L=2} b_l^q g_{kl}^{(l)} \right|^2 \tag{4.66}$$

where, $Q = M^L$. For 2-user BPSK signals $M = 2$, gives the following four possibilities for $\mathbf{b}_q \in [\{-1-1\},\{-1+1\},\{+1-1\},\{+1+1\}]$.

Period 2:- Relay Data Forwarding: In this period, the b'_{kl} at each relay is spread using the same $\mathbf{c}_k$ to form the transmitted signals (shown as dashed-line arrows in Figure 4.24) as follows: $\mathbf{s}'_{kl} = \sqrt{\frac{P_{kl}}{2}}b'_{kl}\mathbf{c}_k$. The b'_{k1} and b'_{k2} may not be identical to the original data b_{k1} and b_{k2} due to decoding errors at the relays. This may arise due to low SNR of channels between the users and their relays. Although, error correction techniques such as CRC codes as in [80] can also be used, we do not consider error correction here for simplicity.

Decoding at the Base-station: The received signals at the base-station from the users during the first period, $\mathbf{r}$, and that from the relays during the second period, $\mathbf{r}'$, can be written as:

$$\mathbf{r}=\sum_{k=1}^{G}\sum_{l=1}^{L=2} g_{kl}\mathbf{s}_{kl} + \mathbf{v} \quad and \quad \mathbf{r}'=\sum_{k=1}^{G}\sum_{l=1}^{L=2} g'_{kl}\mathbf{s}'_{kl} + \mathbf{v}. \tag{4.67}$$

The signals are despread using a synchronised replica of $\mathbf{c}_k$ to obtain soft output signals z_k and z'_k as follows: $z_k = \int_0^{T_b}\mathbf{r}_k^T\mathbf{c}_k$ and $z'_k = \int_0^{T_b}\{\mathbf{r}'_k\}^T\mathbf{c}_k$. As can be noted z_k and z'_k consist of two copies of the users' data $b_{kl}, l \in \{1,2\}$ via independent uplink channels g_{kl} and g'_{kl}. These are diversity combined and the 2-users' data are jointly detected. For this purpose, sum of squared Euclidian distance from z_k and z'_k for each combination of data is calculated and compared with

Table 4.2: Different Collaborative diversity schemes for uplink of CDMA

Scheme	**Sequences**	$\frac{K}{N} > 1$ allowed?	**BW Efficiency**(ω)
Sendonaris et. al. Scheme [7]	Orthogonal/PN	No	0.5
Ribeiro et.al. Scheme [80]	PN only	No	≤ 1
Vardhe et.al. Scheme [76]	Orthogonal/PN	No	≤ 1
Proposed BECD Scheme	Orthogonal/PN	Yes	1

all possibilities similar to that in (4.66). The receiver makes the data decision as follows

$$\{\hat{b}_{k1}, \hat{b}_{k2}\} = \arg \min_{b_l^q \in \{\mathbf{b}_1,.,\mathbf{b}_q,.,\mathbf{b}_Q\}} \left\{ \left(\left| z_k - \sum_{l=1}^{L=2} b_l^q g_{kl} \right|^2 \right) + \left(\left| z'_k - \sum_{l=1}^{L=2} b_l^q g'_{kl} \right|^2 \right) \right\}. \tag{4.68}$$

4.5.5 Performance Results and Analysis

A baseband model of an uplink synchronous CDMA and BPSK modulated users employing Walsh Hadamard sequences with $N = 16$ are used. There are two collaborators $L = 2$ sharing a single spreading sequence and simulations are carried out for full system load conditions. To provide a fair comparison with conventional CDMA, the total transmit power of users under the proposed scheme is halved for the two consecutive transmission periods. The channel estimation is assumed perfect here, however methods for estimation with co-channel users signals can be obtained using orthogonal training sequences such as [115], [116].

Bandwidth Efficiency: The bandwidth efficiency or the rate of a cooperative CDMA is defined as $\lambda = \frac{KT_{nc}}{NT_{co}}$, where T_{co} and T_{nc} are the durations of the cooperative and the non-cooperative scheme, respectively. The proposed scheme with $G = N$ and $L = 2$, supports $K = 2N$ users, hence, even with $T_{co} = 2T_{nc}$ it achieves full rate $\lambda = 1$ compared with $\lambda = 0.5$ in [7] where $K = N$. In Table 4.2, we provide a comparison of key performance attributes of different collaborative diversity schemes and the proposed BECD scheme (applies to both superposition and space-time) for uplink CDMA. The BECD has some important advantages compared with other schemes. Unlike, the Ribeiro et.al. scheme, the scheme is not limited to the use of statistically orthogonal PN sequences for achieving diversity gain. Another advantage of our approach is that if non-orthogonal spreading sequences are used, the variance of MAI can be significantly low compared with other schemes due to sharing of same sequence. The most important of all is that, unlike all other schemes, the BECD system design can support number of users much higher than the spreading factor.

4.5.5.1 BECD without Relays

Figure 4.25 shows the average BER performance of the simple signal superposition based BECD. The BER performance of conventional CDMA with single antenna transmission (No collaboration) and that using two antenna transmit diversity using Alamouti scheme [73] is also presented. The results show that the BER performance of the proposed scheme improves significantly as the quality of inter-user channel improves. For example, with the relative SNR gain of inter-user channels compared with respective uplink channels of users $\beta_k = 10dB$, approximately 7 dB gain in terms of transmit power is achieved at BER of 10^{-3}. As the inter-user channel gain increases the BER performance further improves and approaches very near to that of Alamouti scheme. The BER of superposition based BECD scheme with perfect collaboration (no error from inter-user channels) is also shown, which is about 1dB worse than fully orthogonal (i.e. Alamouti) scheme.

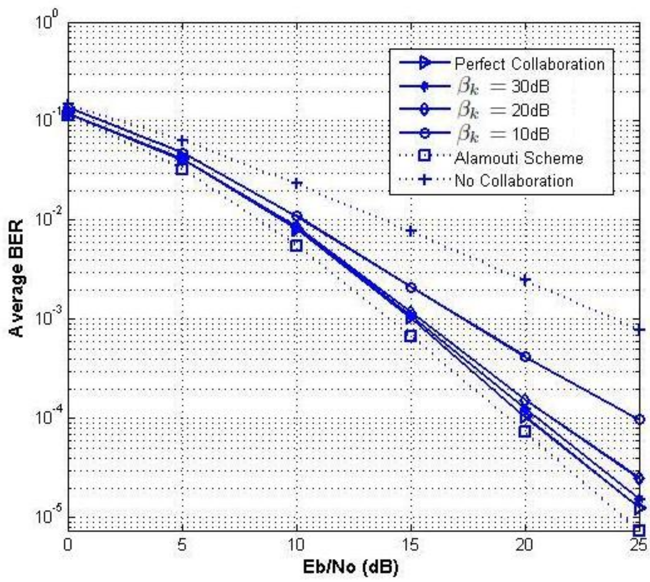


Figure 4.25: Performance of Signal Superposition based BECD scheme under different β_k, $L = 2$ and Walsh Hadamard Sequence of $N = 16$

Figure 4.26 shows the average BER performance of space-time coding based BECD. The BER performance of conventional CDMA with single antenna transmission (No collaboration) and that using two antenna transmit diversity using Alamouti scheme [73] is also presented. The BER performance of the proposed scheme is shown to improve significantly as the quality of inter-user channel improves. It is identified that the space-time coded scheme offers more improved BER than the superposition based scheme. For example, with the relative SNR gain of 10dB of inter-user channels, approximately 7.5 dB compared with 7 dB gain in terms of transmit power is achieved at BER of 10^{-3}. As the inter-user channel SNR gain becomes higher the BER performance further improves and approaches very near to that of Alamouti scheme. More specifically, a gap of approximately 0.5dB is observed with inter-user gain of 30dB, asserting the loss of decode and forward relaying compared with the space-time with an ML scheme.

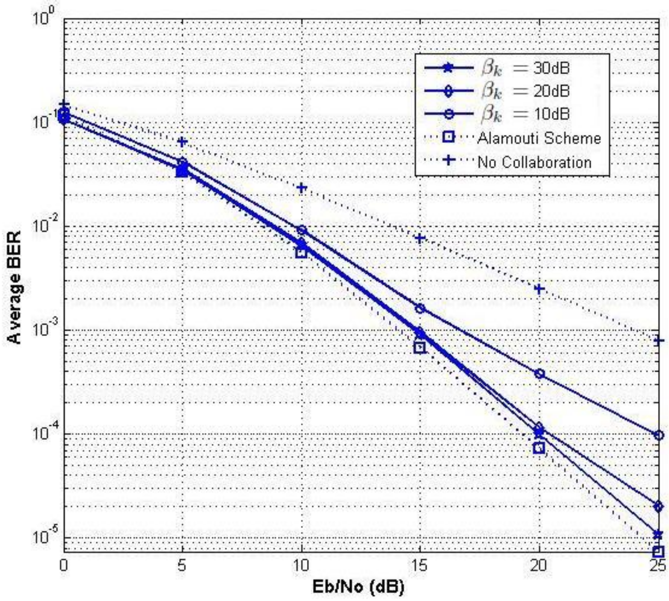


Figure 4.26: Performance of Space-time Coding based BECD scheme under different β_k, $L = 2$ and Walsh Hadamard Sequence of $N = 16$

4.5.5.2 BECD with Relays

Simulations Settings: In this mode of operation, we consider few assumptions and systems scenarios. Synchronisation is a very important practical consideration and two cases are considered here: a) perfect synchronisation and b) perfect synchronisation in the first period, but with a timing error in the second period. The timing error signal is modeled as complex Gaussian random variable with standard deviation in fraction of chip period. Perfect channel estimation is assumed for simplicity. Assuming equal average power of all the uplink channels e. g. g_{kl} and g'_{kl}, relative SNR gain of a user to relay channel g^l_{kl} compared to corresponding uplink channel g_{kl} , β_k, is given by: β_k=$E\{|g^l_{kl}|^2\}/E\{|g_{kl}|^2\}$, where $E\{.\}$ is the expectation. For ease of performance study, SNR gain of the g^l_{kl} compared to the co-channel user to relay channel $g^l_{ki}, i \neq l$ is defined as μ_k=$E\{|g^l_{kl}|^2\}/E\{|g^l_{ki}|^2\} = \beta_k$.

Diversity Performance: Figure 4.27 shows the BER under perfect synchronisation for different β_k values. BER of existing full rate schemes, single antenna 'No Cooperation' and that of space-time encoded '2 Antenna Alamouti' [73] are also shown. It is evident that the BECD achieves significantly improved BER compared with 'No Cooperation' as β_k increases. For example, with $\beta_k = 10$ dB, E_b/N_0 gain of ≈ 7 dB is achieved at the BER of 10^{-3}. This is attributed to the use of two independent fading channels of the users and their relays. As β_k increases to 30 dB, the BER further improves, approaching within 1 dB of the '2 antenna Alamouti' scheme.

Effect of Synchronisation Error: BER under the synchronisation error at different E_b/N_0 values are plotted in Figure 4.28 for $\beta_k = 30$ dB. For comparison, BER of 'No Cooperation' under perfect

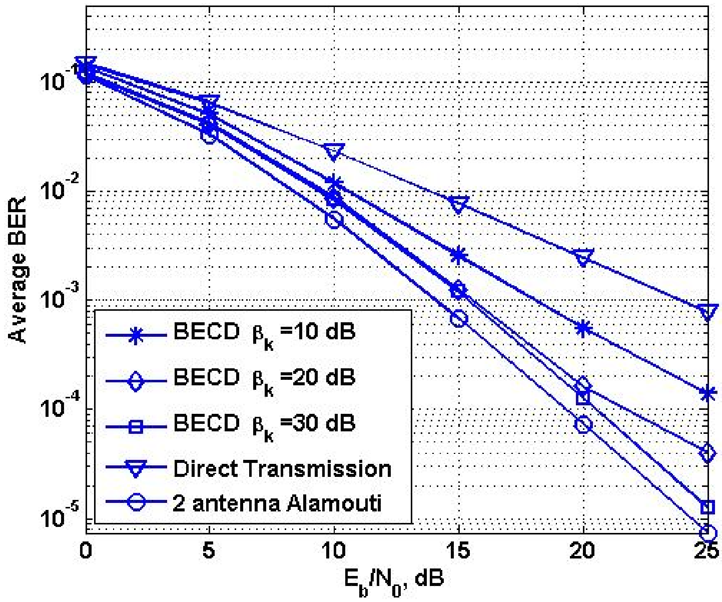


Figure 4.27: BER of the proposed BECD scheme for different β_k values

synchronisation is also plotted. As can be seen, it has significant impact on the BER, specifically at high E_b/N_0 values. The points where 'No Cooperation' line crosses the proposed scheme at respective E_b/N_0 denote the timing errors in chips required to retain its performance advantage. For example, with error of as high as 0.25 chips, it can still offer gain.

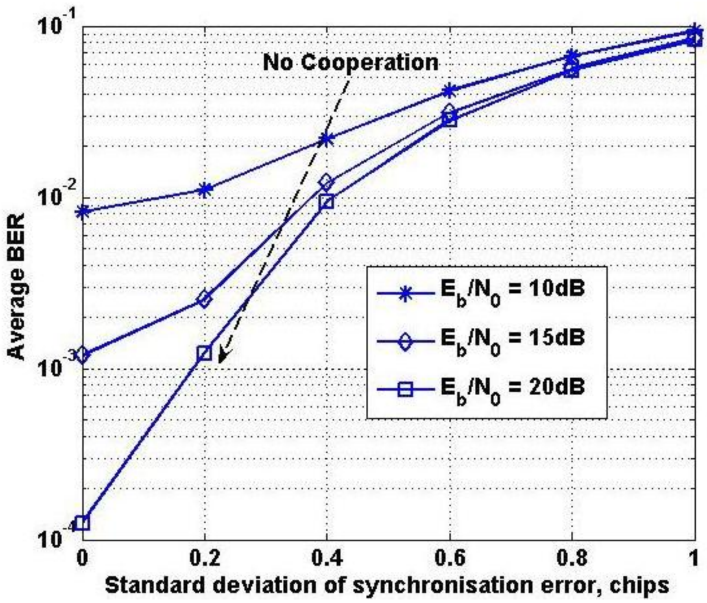


Figure 4.28: The effect of synchronisation error on the BER of the proposed BECD scheme for $\beta_k = 30$ dB

4.6 Chapter Summary

This Chapter presented new techniques for providing spatial diversity in non-orthogonal uplink multiple access (CCMA and CDMA) channels. Special attention is paid to a novel way of achieving diversity with user collaboration. First, a simple transmitter antenna diversity scheme for CCMA with multiple antennas at user nodes is proposed in Section 4.2 to improve the error performance in fading channel conditions. The scheme has shown to provide error performance gain similar to that of CCMA with receive antenna diversity. Illustrative results using both analytical and simulation techniques are presented that verify the gain attained by the scheme. Next, in Section 4.3 a new scheme is proposed to improve the performance of multi-user CCMA system by employing user cooperation diversity in fading channel conditions. The analysis of probability of error performance of CCMA and Collaborative CCMA are presented under different diversity and nearfar channel conditions. The proposed scheme is shown to provide significant gains approaching that of CCMA with dual receiver diversity as the ratio of inter-user channel SNR gain increases.

Two new contributions are presented to improve the performance of uplink CDMA by employing user collaboration diversity combined with conventional and blind adaptive SIC detection techniques in Section 4.4 and are referred to as C-SIC and C-BASIC, respectively. Using Gaussian approximation and channel capacity theorem, the analysis of sum capacity gain of the C-SIC is also carried out and compared with that of a system without cooperation and interference cancellation. Parallel with higher achievable rate, the C-SIC shows much improved BER performance as demonstrated by the range of simulation results. It is noted that the use of MF output for MAI estimation in C-SIC does not provide full diversity gain of due to imperfect MAI estimation and cancellation. The C-BASIC is shown to alleviate this problem by its inherent MAI suppression and estimation capabilities. Therefore, C-BASIC shows improve BER and robustness even under high user loading as well as nearfar conditions.

Finally, a novel near-unity rate collaborative diversity scheme referred to as BECD for multiple access fading channels is proposed in Section 4.5 and in particular evaluated for CDMA. This is shown to be achieved by pairing two users and their relays to use the same code and using joint detection of co-channel signals that exploits the differences in their fading channel magnitudes. For example, impressive SNR gain of ≈ 9 dB is achieved compared with single antenna non cooperative scheme. It is also shown that the diversity is retained under modest synchronisation error. Our future work will constitute the design of synchronisation and channel estimation techniques and also combined error correction coding to further enhance the system performance.

To sum up, it has been shown that collaborative diversity combined with MUD offers significant gain in terms of minimizing the error probability and that we are able to achieve the gain also with improved bandwidth efficiency. In these schemes, independent fading channels of users is exploited to achieve spatial diversity without requiring antenna array at both sources and the destination. In the next Chapter, another technique, termed as 'User Collaboration' will be introduced for CDMA, which can offer significant increase in user capacity compared with conventional (Non-collaborative) techniques. Questions of how and why the new technique offers the said gain without requiring extra resources such as antennas, power, and bandwidth will be answered.

Chapter 5

High User Capacity Collaborative CDMA Schemes

5.1 Introduction

Based on studies of wireless channels and various techniques, a set of new contributions under the theme of User Collaboration is presented in this Chapter. It starts with an overview of general ideas behind the new approach to achieve high capacity multiple access. The motivations behind the idea for uplink of CDMA is discussed in 5.2. Following the discussion, subsequent sections provide three important contributions to address the user capacity and error performance limitations of CDMA. A system that can accommodate more users than the processing gain is often termed as overloaded or oversaturated CDMA. The work in [25, 26] and others on overloaded CDMA will be used for comparison with collaborative CDMA proposed in this research referred to as CS-CDMA. The CS-CDMA proposes a different approach to tackle the problem of overloading in CDMA by using the concept of grouping small number of users for sharing a single spreading sequence. This reduces the number of sequences required in a system and by using CDMA detection on per group basis, the amount of interference is also reduced significantly. In depth descriptions and analysis of the proposed techniques are provided to show the performance advantages.

First, a novel technique of collaborative spreading for downlink of overloaded CDMA (CS-CDMA-DL) is proposed in Section 5.3. In Section 5.4, the idea of the collaborative spreading combined with space-time coding referred to here as collaborative space-time spreading (C-STS) is investigated to improve the performance in fading channels. In Section 5.5, the collaborative spreading approach for non-orthogonal uplink of CDMA (CS-CDMA-UL) is investigated.

5.2 User Collaboration: A New Approach for Higher User Capacity

5.2.1 Background

Diversity is one of the most well-known multiple antenna technique used in wireless fading environments. It can significantly improve the error performance when fading in each receive antenna is independent. Another benefit of having multiple antennas is the spatial data multiplexing [32]. It exploits the idea of using independent channels created by multiple antennas at the transmitter and receiver, also called Multi Input Multi Output (MIMO) channels, to increase in several fold the achievable rate than the Single Input Single Output (SISO) channels [102]. The technique assumes rich scattering environments to increase the achievable rate of the point to point communications. Using MIMO, new techniques to improve the performance of CDMA is presented in [104, 117]. However, the main practical problem is that, in a multiuser system, due to power and size limitations the mobile users cannot afford to employ more than one antenna to communicate with the basestation. Therefore the MIMO multiuser schemes [104, 117] may still be difficult to exploit under the present technology.

It is shown in [31] that, using antenna array and optimum combining method, the user capacity in the uplink of cellular system can be increased significantly. Where it is identified that the diversity gain from multiple antennas can effectively support multiple users less than number of antennas. The technique matured later to be called as SDMA, where separating multiple cochannel users is carried out by exploiting their spatial signature and nulling out the interferers by using beamforming method. The literature on adaptive antenna array and beamforming is quite rich and these techniques have also found their applications in the downlink cellular systems to suppress interference as well as to perform precoding for achieving MIMO spatial multiplexing [2]. Although use of multiple antennas is becoming more and more prevalent due to their huge capacity promise, an interesting question is: Can we use the single antennas of users and still gain the capacity of multi antenna systems? Although this may sound infeasible, due to practical constraints such as antenna spacing, limited power of terminals, it is desirable to do so with least number of antennas if possible. This question is the main driving issue for the work presented in this Chapter.

It has been noted in a discussion in Section 2.8 of Chapter 2 that the antennas of multiple users can be coordinated to achieve spatial diversity. When this technique is looked from the view of increasing the user capacity of multiple access schemes, the separate antenna of multiple users can be treated as elements of a virtual antenna array and each antenna can transmit independent data (also called spatial multiplexing), occupying the same portion of a multiple access channel. This means that independent channels of users can be exploited to reuse the single resource and the co-channel users can be jointly demodulated at the receiver. It is shown later in Section 5.5 of this Chapter that, when a group of users share the common resource, e.g. spreading sequence and the MUD scheme is effectively used, this can lead to a significant increase in the user capacity of the system for the same transmit power, signal bandwidth compared with conventional single sequence per user approach. First, simplified discussions and the basic ideas of the new collaborative transmission and detection technique is described next.

5.2.2 Basic Ideas of User Collaboration

Motivated by the need to address the aforementioned questions of feasibility of having multi antenna user nodes, a novel technique has been proposed in this Chapter to increase the user capacity for uplink of CDMA without requiring extra resources such as antennas, power or bandwidth. More specifically, a practical CDMA uplink with users employing non-orthogonal spreading sequences and use of MUD scheme [28] at the basestation receiver is considered. The total users within the given cell are first divided into groups of small number of users. The users within a group collaborate to transmit their independent data using common spreading sequence. *It has to be noted here that unlike 'Collaborative Diversity', where independent fading of users' channels are utilized to achieve spatial diversity, the 'User Collaboration' technique introduced in this Chapter, exploits independent fading channels of users to increase the sum rate of multiple access scheme by reusing the same system resource (spreading sequence) by several users.* The basestation receiver performs despreading of received signal using group specific sequences followed by a linear MUD processing on group basis and data signals of individual users belonging to a group are estimated jointly using maximum likelihood detection and decoding approach. This also means that the grouped users occupy only a fraction of signal bandwidth compared to using single spreading sequence for each user. We are relying on the fact that given the random distribution of users' channels, there is negligibly small probability that channels of collaborating $\{x, y\}$ users within say k^{th} group, $\{g_x^k, g_y^k\}$ are equal i.e. $Pr\{g_x^k = g_y^k\} \rightarrow 0$. Hence the co-spread users' data can be demodulated by exploiting their channel differences. When the users' channels are highly correlated i.e. $Pr\{g_x^k = g_y^k\} \neq 0$, the detector faces with the problem of ambiguity. This is solved by employing collaborative coding and will be briefly discussed next.

When the transmit channels of users, $\{x, y\}, x \neq y$ within k^{th} group are not independent (correlated) i.e. $Pr\{g_x^k = g_y^k\} \neq 0$ the detector may have very high error probability. Such conditions may arise when the users are collocated and are under direct line of sight from the basestation. The problem of ambiguity in detection occurs when there are two or more users sharing the same multiple access channel. This is because the same composite signal point may correspond to more than one possible combination of users' data. The use of uniquely decodable codes [12, 41, 14, 46] for each user as used in CCMA solves the ambiguity problem while also achieving sum rate higher than unity. The collaborative coding is used within the new scheme to improve the performance by resolving the problem of detection ambiguity due to correlated channels of users. The sum rate of this scheme can however be much lower than the previously described uncoded scheme (since it offer full rate transmission when channels are uncorrelated), as the former requires users to transmit codewords of $n, n > 1$ symbols to encode a single bit. Although we have focused mostly on the uplink CDMA, collaborating coding applies readily and much easily in the downlink and a new technique to achieve high user capacity referred to as Collaborative Spreading CDMA (CS-CDMA-DL) is described next.

5.3 Collaborative Spreading for CDMA Downlink (CS-CDMA-DL)

5.3.1 Introduction

It is well known that the use of orthogonal sequences maximizes the spectral efficiency of a synchronous CDMA [118], but its user capacity is limited by the spreading factor. When random (not fully orthogonal) sequences are used, the user capacity is also limited MAI arising from the non-zero cross correlation of the users' signals. However, the use of optimum multiuser receiver with maximum likelihood sequence estimation is shown to achieve both spectral and user capacity asymptotically in [6], at the cost of exponential increase in complexity with the number of users. Due to power limitations and the need for reduced implementation complexity on the downlink receivers, a simple correlator detector and orthogonal sequences are used in practice.

There are different approaches for increasing the user capacity of CDMA in the literature such as those reported in [119, 120, 121, 122, 40, 26, 23, 43, 25, 44]. It is well known that the capacity is maximized when sequences (known as Welch Bound Equality (WBE) sequences) with total squared correlation (TSC) achieving the the Welch Bound [39] are employed. However, for satisfying the optimum TSC condition, the sequences of all users need to be updated each time a user enters/leaves the system, which is not that viable in practice. A scheme to support $N + U$ users in N-dimensional global signal space without sacrificing minimum distance is proposed by Ross and Taylor in [123]. This is achieved at the cost of a complicated multiuser receiver and with a low overloading ratio of $K/N \leq 1.33$. In view of limited number of available channels of conventional CDMA with orthogonal sequences, Quasi Orthogonal Sequences (QOS) are proposed in [44] by overlaying U additional (modified) orthogonal sequences on top of the N sequences to support an overloaded system. Another scheme called Random OCDMA/OCDMA (O/O) is proposed in [26]. It uses two sets of orthogonal sequences, where the user sequences within each set are scrambled with a distinct random sequence. In this technique a multiuser interference cancellation and iterative detection is used. Improved Random OCDMA/OCDMA using time displaced sequence sets and different chip pulse bandwidth is proposed and evaluated in [40]. The spectral efficiency analysis of the O/O technique with a feedback receiver operating in AWGN and fading channel environments is presented in [23].

A group based orthogonal CDMA scheme using collaborative signal mapping for oversaturated CDMA is proposed in [124]. Superposition coding is another technique investigated in [119] for multiuser transmission using a single spreading sequence in which pairs of users decode their data from the common received signal. However it causes more interference to the weaker users and requires more aggregate transmit power. Furthermore, it requires real-time knowledge of all paired users' relative SNR differences and hence adds extra processing load at the basestation transmitter for power allocation to each user. Considering the fact that, the mobile users can not afford receivers with high complexity multiuser detection techniques, transmitter pre-processing based schemes have also been investigated. For example in [120, 122], multiple antennas and multiuser pre-processing at the transmitter are used, where a group of users is assigned a unique spreading sequence. Both require closed loop operation for updating each user's channel state information, and may result in more complex system.

In this contribution, we propose a new higher capacity and low complexity scheme by collaboratively spreading more than one user's data using a single sequence. This is inspired by the idea of collaborative coding in [14, 15, 46]. The objective is to increase the user capacity using the available orthogonal sequences while also achieving total sum rate higher than unity (assuming BPSK modulated users' data). The proposed scheme does not require channel knowledge at the transmitter and uses a simple decoding method to recover the desired data from a small set of allowable codeword combinations at the receiver. Full system design is provided and evaluated in AWGN and flat fading Rayleigh channels. The theoretical bit error rate (BER), user capacity and sensitivity to channel estimation error are also presented together with simulation analysis results. Finally, performance comparisons with other schemes such as those in [44, 26, 40] and [124, 125] are provided.

This Section is organized as follows. In subsection 5.3.2, the CS-CDMA-DL system model is described. The detection and decoding techniques are then presented in subsection 5.3.3. The BER performance analysis is carried out in subsection 5.3.4. System performance results and comparisons with different schemes are shown in subsection 5.3.6.

5.3.2 System Model

In this Section, similar to non-overloaded orthogonal CDMA we employ $\leq N$ orthogonal number of sequences for G groups of users with $G \leq N$, so that the orthogonality of signals between users' groups is retained. To support an overloaded system, each sequence is shared by a group of T users each employing a set of codewords from a collaborative code, where T is a small number (for example T = 2 or 3 is considered in the current work). In principle T could be large depending upon the availability of simple and higher rate codes and even better with inherent synchronization properties. Although the composite codewords of each group are non-orthogonal, they are uniquely decodable by the coding design, and therefore the desired user will only need to despread the received signal and identify its codeword from the composite signal to recover its own data. This method results in perfect separability of signals of all users. It should also be noted here that different collaborative codes can also be used for different groups for further system design optimization. In the case when higher throughput for a particular user is required, a multiple set of codewords and spreading sequences can be assigned as in Multi-code CDMA systems.

A generalized system model and block diagram of the proposed downlink collaborative CDMA scheme is shown in Figure 5.1 (a) and (b), respectively. The base-station transmits independent information signals to $K = GT$ users simultaneously on their respective channels $g_{kl}; 1 \leq k \leq G, 1 \leq l \leq T$. The users' data in k^{th} group are collaboratively encoded and summed before spreading and then transmitted. For practical considerations, the encoding of each user data is performed locally at the base-station and can be independent to ensure the privacy of each user data. In addition, the synchronization between all users' data is easily achieved at this single point. At the receiver of the l^{th} user within the k^{th} group, the received signal $r_{kl}(t)$ is first despread and the composite codeword is then decoded to form estimates of the kl^{th} user's original transmitted data $\hat{b}_{kl}$.

There are various coding design schemes for Gaussian multiple access and broadcast channels [12, 42, 41, 14, 9, 11, 126, 13]. A practical system design using complex valued collaborative coding technique for fading multiple access channel conditions called CV-CCMA is presented in [46]. However, in this work the downlink channel is considered, where all users' data are under the control of base-station unit and the composite codeword signals are formed before transmission. Therefore the problem of coherence combining of users' signals does not exist here. At the receiver the estimate of the desired user's channel gain is still required for full decoding of the composite signal and hence its data.

The principles of collaborative coding is described here for the MAC and the reverse process applies equally to the downlink. Consider a system with T users, transmitting independent data on a common MAC. Each user $l, 1 \leq l \leq T$ is assigned a set of N_l codewords from the collaborative codes $\mathbf{C}_l = \{C_{l1}, C_{l2}, ..., C_{lN_l}\}$ of length n bits. The data of each user are encoded using the set of codewords from $\mathbf{C}_l$, then mapped using linear digital modulation technique. The received

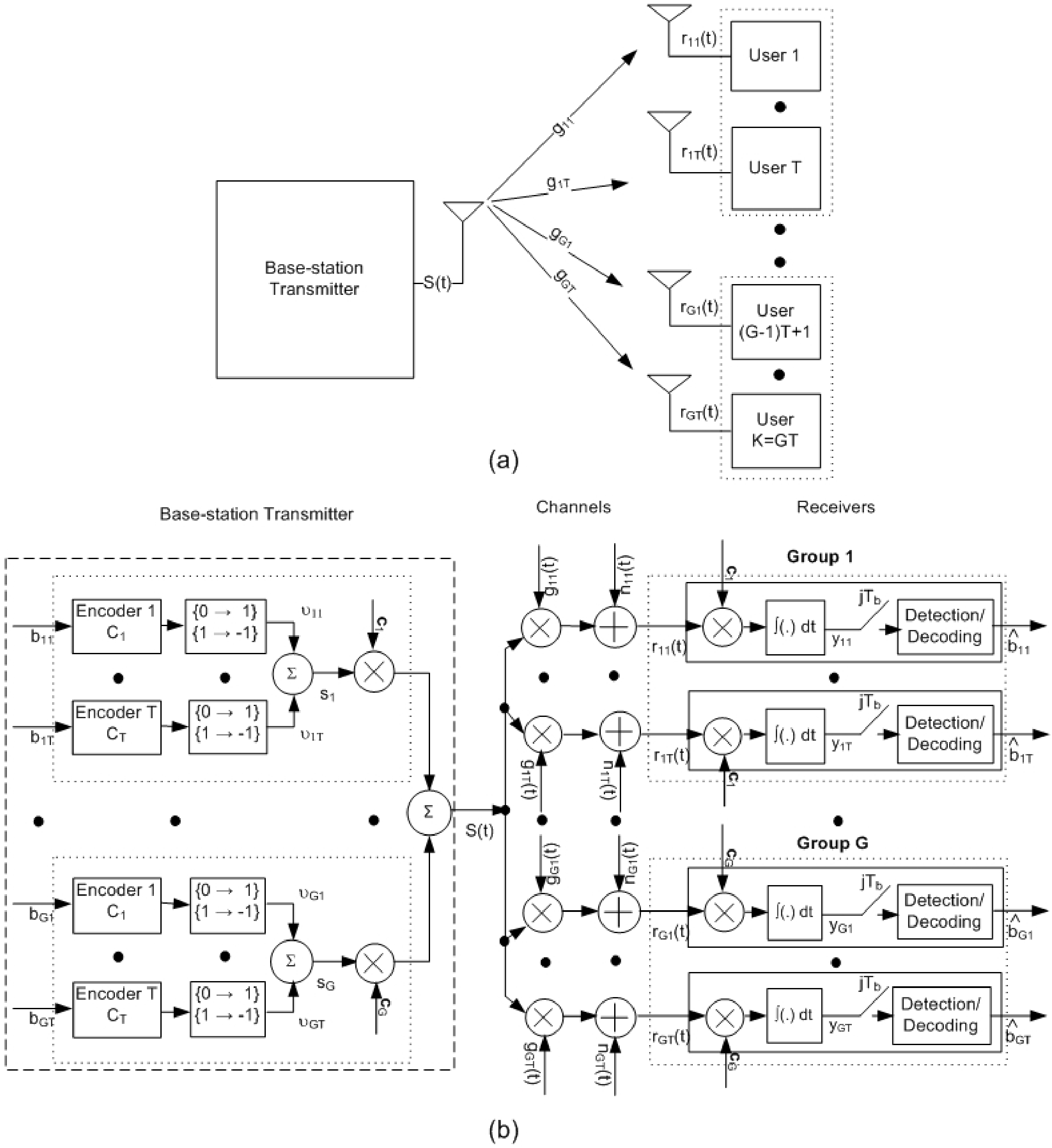


Figure 5.1: CS-CDMA-DL system model (a) generalized model (b) G groups T users system block diagram

signal is the output of MAC consisting of sum of each user's codeword signals and possibly with some added noise. It is worth noting here that the combining of signals in practical systems is a challenging problem investigated in [46], though not of an issue here since it is considered for the downlink. The total sum rate R_{sum} in bits per channel use for this coding scheme is given by:

$$R_{sum} = \sum_{l=1}^{T} \log_2(N_l)/n. \tag{5.1}$$

For example, $T = 2$ with $N_1 = 2, N_2 = 3$, is a commonly used scheme to illustrate collaborative coding with achievable sum rate of $R_{sum} = 0.5 + 0.79 = 1.29$ bits per channel use which is higher than that of conventional multiple access schemes [12]-[46]. Each user's codewords and resulting combinations are shown in Table 5.1. As can be seen, the composite codewords are unique and a single decoder can perfectly unscramble the total signal to deliver the individual user's original codewords and data. Note that multiple access function is achieved here without subdivision in time, frequency or orthogonal codes. Recently a new uniquely decodable binary code pair for the two user binary adder channel is given in [126]. It achieves highest known rate of 1.378 bits per channel use. The unique decodability property of collaborative coding is the means for achieving higher user capacity of our proposed technique. For example, using the set of codewords $\mathbf{C}_1$ and $\mathbf{C}_2$ in Table 5.1 and assigning them to two users sharing the same spreading sequence, we can easily increase the number of simultaneous users supported in the system.

C_1+C_2		(C_1) (0 0)	(C_1) (1 1)
(C_2)	(0 0)	0 0	1 1
(C_2)	(0 1)	0 1	1 2
(C_2)	(1 0)	1 0	2 1

Table 5.1: Collaborative encoding and allowable codeword combinations for two co-spread users with $R_{sum} = 1.292$ bits per channel use

The collaborative spreading and transmission is now described in details, using a baseband model of a downlink chip synchronous DS-CDMA system as shown in Figure 5.1. The data of each user in k^{th} group b_{kl} is first encoded with a collaborative codeword $C_{lx} \in \mathbf{W}_k, x \in \{1, N_l\}$ and modulated to form the symbols $\upsilon_{kl}(j), 1 \leq j \leq n$; where, b_{kl} is the user's binary data signal and of period T_b taking values $[1, 0]$ with equal probabilities, and $\mathbf{W}_k$ is the set of codewords of all users within the group. The combination of codeword symbols of the T users in the k^{th} group $s_k(j)$, can be written as:

$$s_k(j) = \sum_{l=1}^{T} \upsilon_{kl}(j), 1 \leq j \leq n. \tag{5.2}$$

Each composite codeword signal s_{kj} is spread using a distinct orthogonal spreading sequence $\mathbf{c}_k$. The signals of all G groups of users are summed to form a composite transmit signal $S(t)$, which can be written as:

$$S(t) = \sum_{k=1}^{G} s_k(j)\mathbf{c}_k, 1 \leq j \leq n. \tag{5.3}$$

The vector $\mathbf{c}_k$ repeats at every symbol period, which consists of chip values $[-1, +1]$ with rectangular pulse shaping of period T_c and spreading factor of $N = T_b/T_c$. Note that although rectangular pulse shaping is used for simplicity, the scheme can be easily generalized to use different pulse shaping methods.

It should be noted here that these composite signals prior to spreaders are multilevel, and if we look at the final spread spectrum signal of the proposed scheme, we are in effect maintaining the same multilevel signals as the conventional CDMA transmission for the same number of users. Here, rather than using separate spreading sequence for each user and then summing the spread signals, we first sum the grouped users' encoded data and then spread using a single sequence.

5.3.3 Joint Detection and Decoding

The received signal $r_{kl}(t)$ for the kl^{th} user can be written as:

$$r_{kl}(t)=g_{kl}(t)S(t) + n_{kl}(t), \tag{5.4}$$

where $g_{kl}(t) = \alpha_{kl}(t)e^{j\phi_{kl}(t)}$ is the channel gain with amplitude $\alpha_{kl}(t)$ and phase components $\phi_{kl}(t)$ and $n_{kl}(t)$ is the AWGN with two sided power spectral density $N_0/2$. It is also assumed that transmit channels of users are non-dispersive and remain constant over a codeword period (in fading case).

At the user's receiver first the composite received signal $r_{kl}(t)$ is chip matched filtered to form a received signal vector $\mathbf{r}_{kl}(j)$. Then the signal is despread using the group assigned spreading sequence $\mathbf{c}_k$ to obtain the soft estimates $y_{kl}(j)$ of the composite codeword signal s_{kj} shown as follows

$$\begin{gathered} y_{kl}(j) = \int_{(j-1)T_b}^{jT_b} \mathbf{r}_{kl}(j)\mathbf{c}_k^T \\ 1 \leq j \leq n. \end{gathered} \tag{5.5}$$

The output signal $y_{kl}(j)$ is sent to detection and decoding process of CCMA to obtain the data estimate of the desired user. In conventional CCMA scheme operating in fading channels, the employment of collaborative codes involves combining of the users' codes with their varying fading coefficients. In [15, 46], an effective and practical approach for the uplink CCMA decoding is proposed by the use of complex valued CCMA employing joint channel estimation and detection called CV-CCMA. However, in this Chapter, the downlink is considered, where all the users' data are under the control of base-station unit, where the composite codeword signals are formed and transmitted to all users. Therefore, the problem of coherence combining of users' signals does not exist, and only the estimate of the desired user's channel gain is required for full decoding of the composite signal and hence the desired user's data. The optimum (maximum likelihood) detection/decoding scheme in terms of minimizing the probability of error is derived next based on minimum distance criteria. This distance based decoding approach using the composite signal ensures full recovery of the desired user's data even though orthogonality is not maintained within the desired k^{th} group.

Without loss of generality we assume that kl^{th} user (l^{th} user in k^{th} group) is the desired user. In a T-user collaborative coding transmission, there are $L = \prod_{l=1}^{T} N_l$ allowable codeword combinations $\mathbf{A}_{klq} = \{a_{klq}(1), .., a_{klq}(n)\};\ 1 \leq q \leq L$, consisting of combinations of the T users' codewords over the corresponding users channels. Each element $a_{klq}(j)$ is the j^{th} symbol of the q^{th} allowed codeword combination and is given by

$$\begin{aligned} a_{klq}(j)=g_{kl}(j)\Big\{\sum_{l=1}^{T} \upsilon_{kl}(j)\Big\}_q, \\ 1 \leq q \leq L, 1 \leq j \leq n. \end{aligned} \tag{5.6}$$

The receiver performs joint ML detection and decoding of the users' codewords by calculating the squared Euclidian distance between the received composite codeword and all allowable combinations in the table. It is assumed that each user has a knowledge of all the allowable composite codewords of its group to decode its own data only. The squared distance metric of the despread signal $y_{kl}(j); 1 \leq j \leq n$ with each combination of codewords $a_{klq}(j); 1 \leq j \leq n$ is denoted as d_{klq}. The distance metrics are calculated by utilizing estimates of user's corresponding channel g_{kl} for each composite codeword signal as follows

$$\begin{aligned} d_{klq}^2 = \sum_{j=1}^{n} \left| y_{kl}(j) - g_{kl}\Big\{\sum_{l=1}^{T} \upsilon_{kl}(j)\Big\}_q \right|^2 \\ 1 \leq q \leq L. \end{aligned} \tag{5.7}$$

The calculated distance metric d_{klq}^2 for each allowable combination of codewords is used to perform decoding such that the one that minimizes the metric is selected as the transmitted codeword of the kl^{th} user.

$$\hat{C}_{kl} = \arg \min_{\mathbf{A}_{kl1}, .., \mathbf{A}_{klL}} d_{klq}. \tag{5.8}$$

The final data decision of the desired kl^{th} user is obtained by demapping the estimated codeword $\hat{C}_{kl}$ to the corresponding data symbol $\hat{b}_{kl}$.

5.3.4 BER Performance Analysis

For the ease of BER analysis, we consider collaborative coded and BPSK mapped user's signals $\overline{C}_l, 1 \leq l \leq T$ (overline is used to denote that these are modulated codewords) obtained from the set of original codewords $\{C_{l1}, C_{l2}, ..., C_{lN_l}\}$ and assume $N_l = 2, \forall l$. The total power for each unique composite codeword signal E_q can be found as follows:

$$E_q = \sum_{j=1}^{n} \Big[\sum_{l=1}^{T} \overline{C}_l(j) \Big]_q^2 \qquad 1 \leq q \leq L, \tag{5.9}$$

where $\overline{C}_l(j)$ is the j^{th} symbol of the modulation mapped codeword $\overline{C}_l$. Assuming that all the codewords of each user are equally likely to be transmitted, the average energy per composite codeword signal period $\overline{E}_q$, can be given by

$$\overline{E}_q = \frac{1}{L} \sum_{q=1}^{L} E_q. \tag{5.10}$$

Since each bit period carries $1/n$ portion of the composite codeword, the average energy of composite codeword per bit period is calculated as $E_b = \overline{E}_q/n$. This finally leads to our desired expression for calculating the average energy per data bit $\overline{E}_b/N_0$, which is given by

$$\overline{E}_b/N_0 = \frac{2E_b n}{TN_0}. \tag{5.11}$$

Having defined the $\overline{E}_b/N_0$, the probability of bit error in AWGN and Rayleigh flat fading channel conditions can be calculated using standard approaches as shown in [5]. For our BER analysis, we obtain the average distance similar to that used in (5.44), but between different unique composite codewords for each group. For this purpose, we calculate distance associated with each codeword combination of the transmitted signals. The absolute magnitude of squared distance between different codeword combinations are calculated and normalized by n to give m^{th} possible metric d_m^2 as follows

$$d_m^2 = \frac{1}{n} \sum_{j=1}^{n} \Big| \Big\{ \sum_{l=1}^{T} \overline{C}_l(j) \Big\}_x - \Big\{ \sum_{l=1}^{T} \overline{C}_l(j) \Big\}_y \Big|^2; \qquad 1 \leq x \neq y \leq L, 1 \leq m \leq M, \tag{5.12}$$

where $\Big\{ \sum_{l=1}^{T} \overline{C}_l(j) \Big\}_x$ and $\Big\{ \sum_{l=1}^{T} \overline{C}_l(j) \Big\}_y$ are the j^{th} symbols of any two distinct composite codewords from T users' signals, and $M = \sum_{h=1}^{L-1} h$, is the total number of possible distances between different codeword combinations. Assuming that all the codewords are equally likely to be transmitted, it is desirable to find the average error metric so that the tools designed to evaluate the error performance of a single user's signal transmission [5] can also be used here. Averaging

d_m^2 over all M possible distances between the different codeword combinations we obtain the average error metric $\overline{d}$ for the bit error performance analysis, given by

$$\overline{d}^2 = \frac{\sum_{m=1}^{M} d_m^2}{M}. \tag{5.13}$$

As noted earlier, the composite codeword signals are multilevel. The impact on the decoding due to this signal structure is that the average distance $\overline{d}^2$ between composite codeword combinations that is normalized by the codeword length n becomes smaller than that of single user BPSK modulated signals. Therefore the performance of the proposed system can be thought of as a scaled version of a BPSK modulated data signals using orthogonal spreading sequences. An approximation of the average probability of bit error Pe, for kl^{th} user in AWGN environment, i.e. $g_{kl}(t) = 1$ can be obtained by using the BER expression using standard Gaussian 'Q' function and weighting by $\overline{d}^2$ as follows:

$$Pe \approx Q\left(\sqrt{\frac{\overline{d}^2 E_b}{N_0}}\right), \tag{5.14}$$

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

The probability of bit error of collaborative coded multiuser signals using BPSK mapping in fading channels can be derived using the tools developed for single user signals as in [5]. Based on the analysis for AWGN channel and using $\overline{z}$, the probability of bit error in flat Rayleigh distributed slowly fading channels is calculated next. The BER performance $Pe(\overline{d}^2|_{\gamma(i)})$, conditioned on the fading channel of a user at each time instance i can be written as

$$Pe(\overline{d}^2|_{\gamma(i)}) \approx Q\left(\sqrt{\overline{d}^2 \gamma(i)}\right), \tag{5.15}$$

where $\gamma(i)$ is the user's instantaneous SNR, which can be written as:

$$\gamma(i) = \frac{\alpha_{kl}(i)^2 E_b n}{T N_0}. \tag{5.16}$$

To obtain the error probability $Pe(\overline{d}^2)$ in fading environment, the equation (5.15) is calculated over all the fading magnitudes of the user's channel and is given by

$$Pe(\overline{d}^2) = \int_0^\infty Pe(\overline{d}^2|_{\gamma(i)}) p(\gamma(i)) d\gamma(i), \tag{5.17}$$

where $p(\gamma(i))$ is the user's PDF of fading SNR distribution. By averaging over all the Rayleigh fading distributions of user's channel, we obtain the probability of error under the proposed technique with simple matched filter receiver front-end as follows

$$Pe \approx \frac{1}{2}\left[1 - \sqrt{\frac{\overline{d}^2 E\{\gamma\}}{1 + \overline{d}^2 E\{\gamma\}}}\right], \tag{5.18}$$

where the expectation $E\{.\}$ is taken over the channel fading magnitudes of the desired user α_{kl}.

Different from the above, an approach based on calculating minimum distance of composite codewords and simple union bound as used for M-ary phase/amplitude modulated signals in [5] can also be used to calculate the upper bound on the probability of error.

5.3.5 Quadrature Signalling for Improved Throughput

Achieving higher data throughput is very important and therefore it becomes essential to extend the proposed scheme using a more spectrally efficient data signalling methods. The scheme can be extended to increase the throughput by the factor of two using quadrature data signalling and then applying collaborative coding to user's data in both in-phase and quadrature domain. This method is termed here as 'IQ-CS-CDMA-DL'. Under this method, the previously described Collaborative Spreading technique is applied independently to both real and imaginary part of each user's data as described in 5.3.2. The receiver performs the operations detailed in 5.3.3 in both in-phase and quadrature domain to obtain codeword signals for the user. This requires correction received channel phase first. To perform the decoding of users' transmitted codewords, the soft signal $y_{kl}(j)$ is multiplied with the complex conjugate of the desired user's channel, $g^*_{kl}(j)$, which is given by

$$u_{kl}(j) = y_{kl}(j) g^*_{kl}(j). \tag{5.19}$$

The final decoding of user's transmitted codewords are then obtained by performing collaborative decoding separately on in-phase and quadrature channels as follows:

$$\begin{aligned} Id^{(q)}_{kl} &= \sum_{j=1}^{n} \left| \Re\{u_{kl}(j)\} - \Big\{|\hat{g}_{kl}(j)|^2\Big\} \Big\{ \sum_{l=1}^{T} \upsilon_{kl}(j) \Big\}_q \right|^2 \\ Qd^{(q)}_{kl} &= \sum_{j=1}^{n} \left| \Im\{u_{kl}(j)\} - \Big\{|\hat{g}_{kl}(j)|^2\Big\} \Big\{ \sum_{l=1}^{T} \upsilon_{kl}(j) \Big\}_q \right|^2 \\ & \qquad\qquad\qquad\qquad\qquad \forall q, 1 \le q \le L, \end{aligned} \tag{5.20}$$

where $\Re\{.\}$ and $\Im\{.\}$ denote real and imaginary part of a complex number. Finally, the estimates of data symbols of the users are obtained by using a lookup table decoding of the codewords to symbols as used at the transmitters by using distances $Id^{(q)}_{kl}$ and $Qd^{(q)}_{kl}$, separately. The kl^{th} user selects l^{th} bit from the data vectors that satisfy the minimum distance condition as the desired data from both I and Q domain.

5.3.6 Numerical Results

A baseband model of downlink synchronous DS-CDMA system of K users is simulated in MATLAB using unit norm Walsh-Hadamard spreading sequences with N=64 and G=64. The 2-user

C_1+C_2		(C_1)	
		(0 0)	(0 1)
(C_2)	(0 0)	0 0	0 1
	(1 0)	1 0	1 1

(a)

$C_1+C_2+C_3$					
	(C_1)	(0 1)	(0 1)	(1 0)	(1 0)
	(C_2)	(0 0)	(1 1)	(0 0)	(1 1)
(C_3)	(0 0)	0 1	1 2	1 0	2 1
	(1 0)	1 1	2 2	2 0	3 1

(b)

Table 5.2: Collaborative codes and allowable combinations for (a) two co-spread users with $R_{sum} = 1.0$ and (b) three co-spread users with $R_{sum} = 1.5$ bit per channel use, respectively

and 3-user collaborative codes given in Table 5.2 are used to give sum rates of 1.0 and 1.5 bits per channel use, respectively. The assigned codewords are BPSK modulated and transmitted over non-dispersive and slowly varying frequency flat channels (in the fading case). The system is fully synchronised and perfect knowledge of the desired user's channel are assumed at the receiver.

From practical points of view, the synchronization is easily achieved at the base-station and hence the scheme has the same synchronization requirement as downlink of non-overloaded CDMA. At the receiver side, conventional chip synchronization techniques can be applied without much difficulties. In addition, collaborative coding synchronization such as those investigated in [127] and [128], can be applied. Similarly, various techniques for obtaining channel state information (CSI) such as the pilot symbol aided [116] can be easily incorporated within the system. The effect of channel estimation error is also investigated here.

For the comparison purposes, the Non-overloaded Orthogonal CDMA, QOS CDMA [44], Random OCDMA/OCDMA [26], Improved OCDMA/OCDMA [40] and group based schemes GO-CDMA [125], and CCGO-CDMA [124] are chosen. In Non-overloaded Orthogonal CDMA, all users simply use distinct orthogonal sequences. In the schemes [44]-[40], users within two sets are assigned the available set of orthogonal sequences scrambled with modified set specific sequences. In the later schemes [125], [124] the available set of orthogonal sequences are divided into subsets and each subset of sequences are assigned to different groups. These schemes are often denoted by the parameters $(V + W, V)$, where V is the number of sequences within a subset and W is the number of additional user per group.

The main design parameters defining the overloading ratio of K/N are set to give equal sum rate for fair comparison between the schemes. The total sum rate of CDMA system can be obtained as $R_{sum-cdma} = KR_{user}$ bps, where R_{user} is the user rate given by $R_{user} = 1/N$ bps for existing CDMA since one bit is transmitted over a period of N chips. For the CS-CDMA-DL with equal rate users of codeword length n, $R_{user} = 1/nN$ bps. Two cases are considered here. ***Case 1***: The 2-user codes given in Table 5.2 (a) are used to give sum rate of 1.0 bps. As can be seen from Figure 5.2, the proposed technique provides higher number of users compared with Non-overloaded Orthogonal CDMA. However, this is achieved at the cost of 1.7 dB increase in the average SNR of the Non-overloaded Orthogonal CDMA. ***Case 2***: The sum rate of 1.5 bits/s is considered with the use of codes given in Table 5.2 (b). The number of users can be further increased using overloading ratio of 3 in comparison with other schemes, similarly at the cost of 2.2 dB increase in SNR from the orthogonal CDMA. The loss in SNR is due to the reduced distance separation with the use of composite codeword decoding method. The theoretical BER performance results of the proposed technique using the codes given in Table 5.2 (b) is also shown to closely match with the simulation result.

In Figure 5.3, we show the BER performance comparison using the codes given in Table 5.2 (b) with that of Random OCDMA/OCDMA for different number of users. It is evident that the proposed technique with $K = 64 \times 3$ users shows rapid improvement in BER as the E_b/N_0 is increased, which is in strong contrast with that of the 'Random OCDMA/OCDMA' showing inevitable BER floor at higher loads i.e. $K > 65$. In Table 5.3 the summary of performance comparisons between the proposed technique using 2-user and 3-user codes given in Table 5.2 and other schemes in AWGN environment is given. It can be clearly seen that the proposed CS-CDMA-DL has more attractive system properties compared with other considered schemes such as higher overloading ratio and improved SINR performance.

In Figure 5.4, the theoretical and simulation results in Rayleigh flat fading channel conditions are presented for $K = 64 \times 3$. It can be seen that for the same BER performance, a large increase in the number of users can be achieved with an additional SNR of around 2 dB. We also investigate the BER performance for $K = 64 \times 3$ in Rician fading channel conditions as shown in Figure 5.5 to assess the impact of severity of channel fading on the system performance. This channel model represents Rayleigh and AWGN channels as special cases, when the Rician factor approaches 0 and ∞, respectively [5]. It can be seen that the proposed technique exhibits gradual improvement in BER with the increase of Rician factor. We also analyze the sensitivity of our scheme to channel estimation errors in Figure 5.6 in Rayleigh fading conditions. For the purpose of this study, we adopt a simple model with estimation error assumed to be additive complex Gaussian random variable that is uncorrelated with the user's channel. The channel samples are obtained from complex Gaussian random variable with zero mean and unit variance $\mathcal{CN}[0, 1]$, and the estimation error signal with $\mathcal{CN}[0, \sigma_e^2]$, similarly. In this case, the increase in the estimation error i.e. σ_e^2 (measured in dB) from -30 dB to -20 dB is shown to degrade the BER performance gradually and should be taken into account in practical system design.

The error performance of IQ spreading of the proposed collaborative CDMA is presented in

Table 5.3: Comparison of different schemes for downlink of CDMA using G=64 and Walsh-Hadamard sequences of length N=64

Scheme	**Sum rate** $R_{sum-cdma}$(bps)	**K/N**	**User rate** R_{user}(bps)	**Average SINR** (dB)
Non-Overloaded Orthogonal CDMA	1.0	1	$1/64$	SNR
CS-CDMA-DL (Table 5.2 (a))	1.0	2	$1/128$	$\approx (SNR - 1.7)$
QOS CDMA [44]	1.5	1.5	$1/64$	≈ 1.5
Random OCDMA/OCDMA [26]	1.5	1.5	$1/64$	< 2
Improved OCDMA/OCDMA [40]	1.5	1.5	$1/64$	< 4.2
CS-CDMA-DL (Table 5.2 (b))	1.5	3	$1/128$	$\approx (SNR - 2.2)$

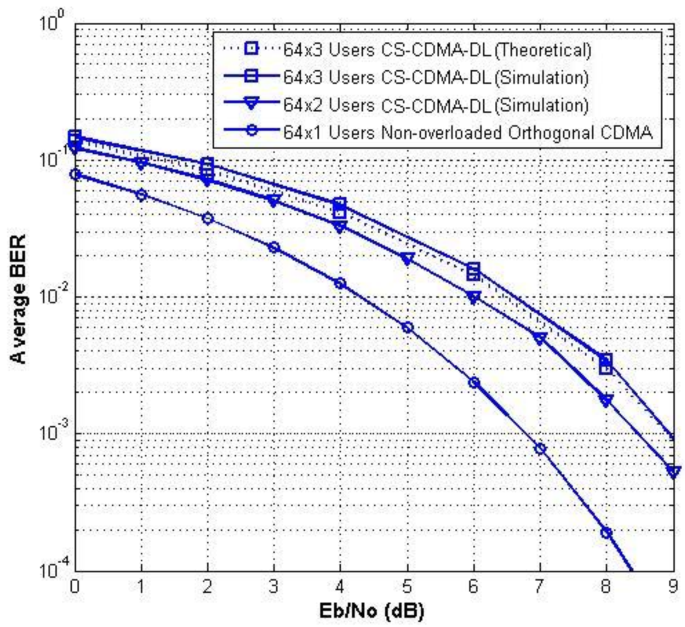


Figure 5.2: BER performance of CS-CDMA-DL in AWGN compared with Non-overloaded Orthogonal CDMA using Walsh-Hadamard sequences of N=64. (for 64×2 users, the code in Table 5.2 (a) is used and for 64×3 users, the code in Table 5.2 (b) is used)

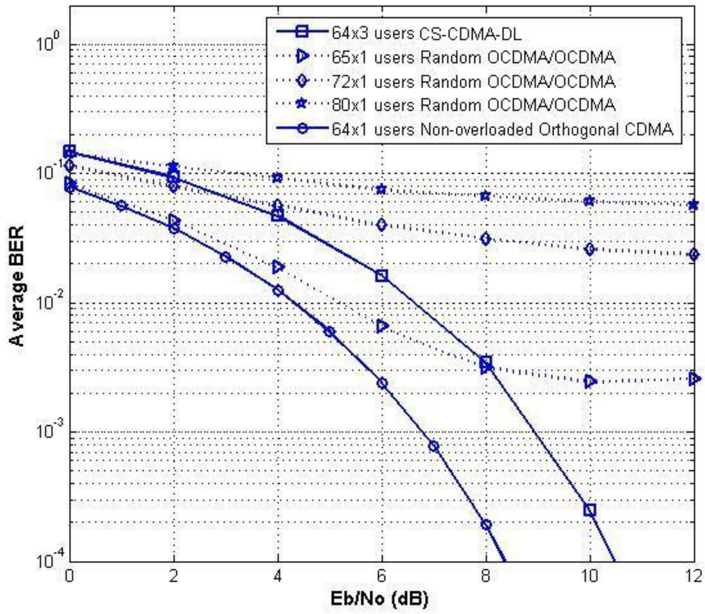


Figure 5.3: BER performance of CS-CDMA-DL compared with Overloaded and Non-overloaded Orthogonal CDMA schemes in AWGN using Walsh-Hadamard sequences of N=64. (codes in Table 5.2 (b) is used

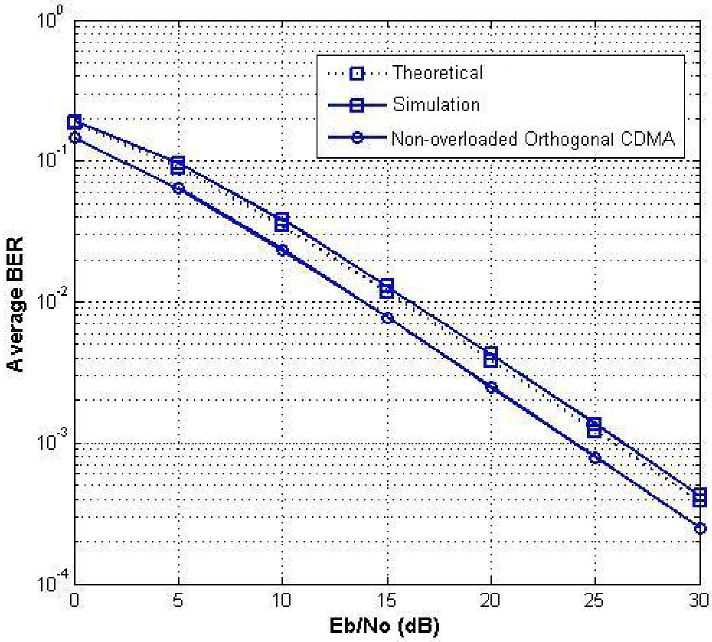


Figure 5.4: BER performance of CS-CDMA-DL in Rayleigh flat fading channels with perfect channel estimation for $K = 64 \times 3$ using Walsh-Hadamard sequences of N=64

Figure 5.7. The as expected the scheme has the same performance as that without IQ spreading in Figure 5.4. Therefore a twice the data rate for each user is achieved for the same amount of transmit power per bit at an additional receiver complexity of L computing per decoded symbol.

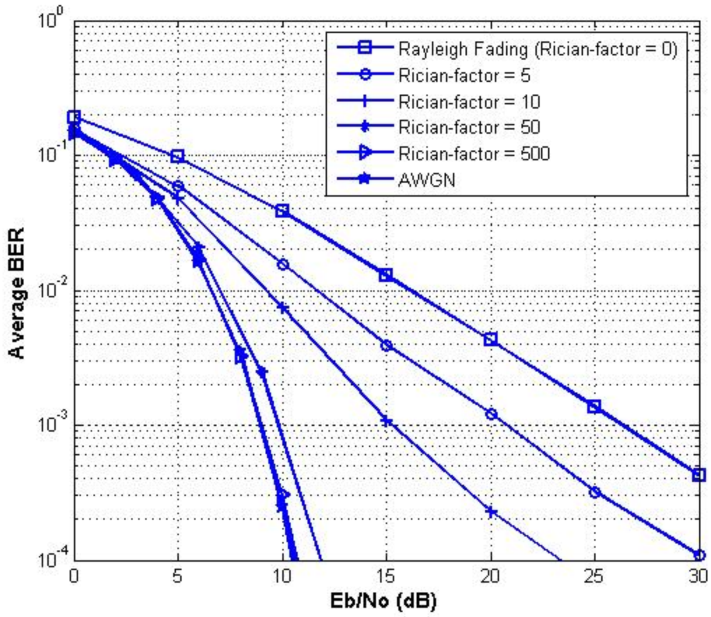


Figure 5.5: BER performance of CS-CDMA-DL in Rician fading channels for $K = 64 \times 3$ under different Rician factors using Walsh-Hadamard sequences of N=64

To provide a fair comparison with group based schemes [125], [124], an average distance measure of signal vectors is defined. Based on the minimum distance between two vectors d^2_{min}, the average distance $\overline{d}^2$ is calculated as follows:

$$\begin{aligned} (V+W)E_b &= \frac{1}{2^{(V+W)}} \sum_{v=1}^{2^{(V+W)}} i_v d^2_{min} \\ \Longrightarrow \overline{d}^2 &= \frac{2^{(V+W)}(V+W)E_b}{\sum_{v=1}^{2^{(V+W)}} i_v}, \end{aligned} \tag{5.21}$$

where E_b is the energy per data bit of a user and i_v is a ratio by which distance from the 1^{st} vector to v^{th} vector is greater than the minimum distance d^2_{min}. The $\overline{d}^2$ value is used to calculate the relative SNR loss λ in dB, compared with fully orthogonal signalling with BPSK i.e. $\lambda = \overline{d}_0^2/\overline{d}^2$ dB, where $\overline{d}_0^2 = 4$,.

In Table 5.4 key performance measures are provided for the different techniques, under the same sum rate R_{sum} and signal dimensions used. For the case of $R_{sum} = 1.5$ bps, GO-CDMA with $(3, 2)$ and CCGO-CDMA with $(3, 2)$ using densest packing of signals is assumed, while collaborative codes in Table 5.2 (b) are used for the CS-CDMA-DL. As can be seen, it has larger average distance $\overline{d}^2$ than that of GO-CDMA, but smaller than CCGO-CDMA. The sum rate and the overloading ratio of the proposed scheme can be doubled i.e. $R_{sum} = 3$ bps and $K/N = 6$,

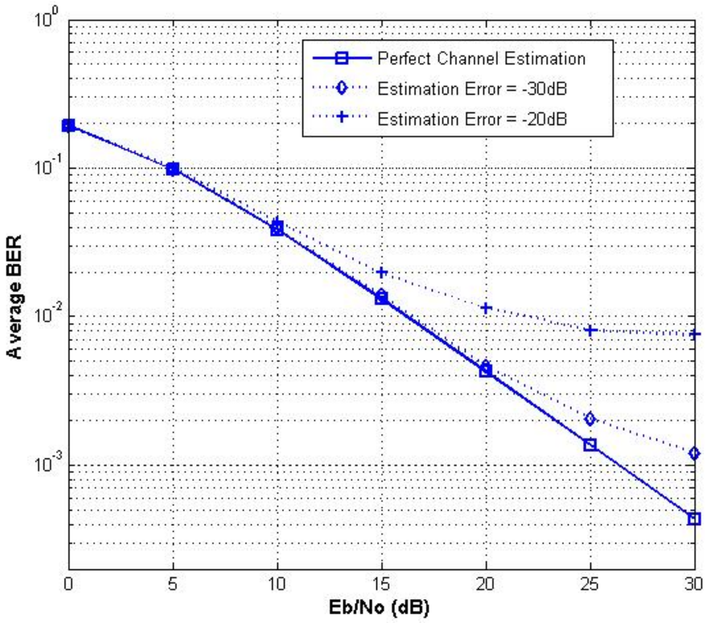


Figure 5.6: BER performance of CS-CDMA-DL in Rayleigh fading channels under imperfect channel estimation for $K = 64 \times 3$ using Walsh-Hadamard sequences of N=64

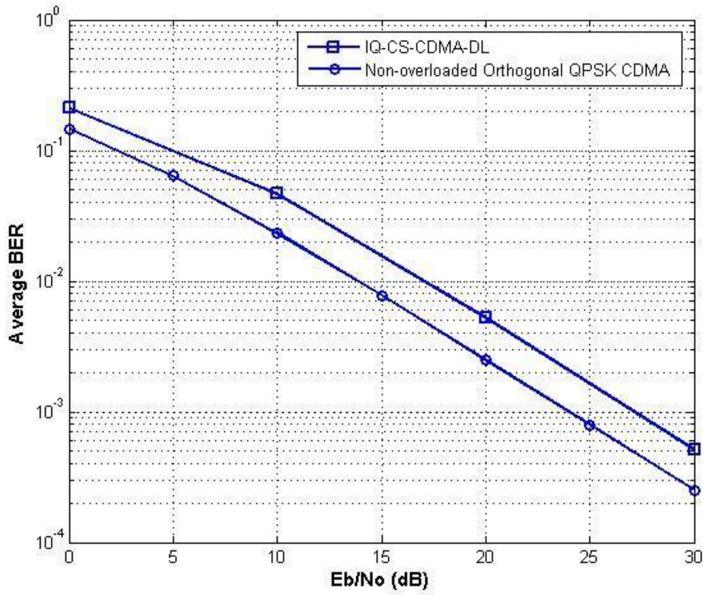


Figure 5.7: BER performance of IQ-CS-CDMA-DL under flat Rayleigh fading channels for $K = 64 \times 3$ using Walsh-Hadamard sequences of N=64

when two signal dimensions are used. This is achieved by simply forming so called in-phase (I) and quadrature (Q) groups, each consisting of sum of T users' encoded data independently on the I and Q dimensions and corresponding I and Q decoding at the receiver side and is referred to as IQ-CS-CDMA-DL. On the other hand, to achieve $R_{sum} = 3$ bps with the same two dimensional space, CCGO-CDMA needs to accommodate 4 additional users i.e $(V + W, V) = (6, 2)$ with $2^{4+2} = 64$ possible signal vectors with Gray mapping. This gives the average distance $\overline{d}^2 \approx 0.6$ corresponding to $\lambda \approx 8$ dB , which is significantly worse compared with the proposed scheme with $\lambda \approx 2.2$ dB. In addition, the proposed scheme requires less complexity receiver as it needs only a single despreading operation and lower number of search for data estimation.

Table 5.4: Comparison of different group orthogonal schemes for downlink of CDMA in AWGN environment

Scheme	**Sum rate** R_{sum}(bps)	**K/N**	**Distance** $\overline{d}^2$	**SNR loss** λ (dB)	**Search per decision**
Non-Overloaded Orthogonal CDMA	1.0	1	4	None	2
GO-CDMA (3,2) [125]	1.5	1.5	1.52	4.19	2
CCGO-CDMA (3,2) [124]	1.5	1.5	2.67	1.76	8
CS-CDMA-DL (Table 5.2 (b))	1.5	3	2.4	≈ 2.2	8
CCGO-CDMA (6,2) [124]	3	3	≈ 0.6	≈ 8	64
IQ-CS-CDMA-DL (Table 5.2 (b))	3	6	2.4	≈ 2.2	16

5.4 Collaborative Space-time Spreading (C-STS) for the CDMA Downlink

In Section 5.3, a new scheme to support overloading in downlink of CDMA also referred to as CS-CDMA-DL is presented. It is noticed that it offers a channel overloading ratio $K/N = 3$ and sum rate 1.5, while requiring only around 2 dB increase in SNR. Also it is verified that the scheme retains high SINR as long as the maximum system load does not exceed T times the number of available orthogonal spreading sequences. When it comes to practice, the performance of cellular wireless systems is effected significantly by the problem of fading, making reliable demodulation of users' data a real challenge and hence requires the use of some form of diversity. It is discussed earlier in the Chapter 2 that, among all diversity schemes, the antenna diversity provides effective solution of the fading problem while not requiring extra resources (time, frequency, and power). In a downlink cellular system, due to the size and power limitations, mobile users can not afford multiple antennas to achieve spatial receiver diversity. Therefore, it is much more efficient to provide diversity by employing multiple antennas at the base-station transmitter, where extra space and complexity is often affordable. This however requires special arrangement of transmitted signals from different antennas to avoid interference to the mobile receiver. This problem can be effectively solved using Space-Time coding technique proposed by Tarokh in [34] and Alamouti in [73]. An interesting technique for downlink of non-overloaded CDMA that uses a form of space-time coding called Space-Time Spreading is proposed in [129]. In this section, a combined collaborative spreading [45] and space-time coding [34], [73] referred to as Collaborative Space-time Spreading (C-STS) is proposed for the downlink of overloaded CDMA in a new system framework to provide spatial diversity to mobile users.

This Section is organized as follows. In subsection 5.4.1, the idea of C-STS including its system model is presented in more details. The transmission and reception methods for C-STS are presented in subsection 5.4.2 and 5.4.3, respectively. Extension to quadrature signalling is discussed in 5.4.4. The BER performance simulation results and comparisons are given in subsection 5.4.5.

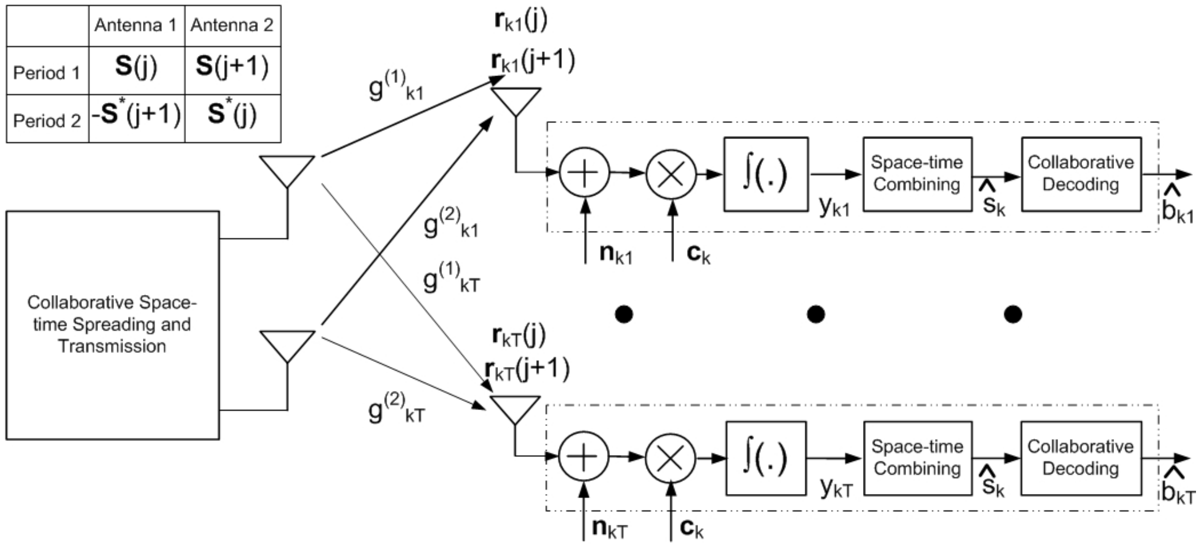


Figure 5.8: System model of C-STS for k^{th} group in downlink of CDMA

5.4.1 System Model

To provide an effective solution to the fading problem, a novel scheme is proposed here. A high level system block diagram of the C-STS transmission and reception scheme is shown in Figure 5.8. The total K users in the system is grouped into G groups and T users within each group. Each user within a group is assigned a set of collaborative codewords as shown in Table 5.5. The same codes are reused for users within all G user groups. The sum of encoded group signals are then spread using the group specific G orthogonal spreading sequences. A block diagram of CS-CDMA-DL shown in Figure 5.1 of Section 5.3 in this Chapter can be used to see these operations. The spread signals from all user groups are then summed to form a composite transmit signal $\mathbf{S}(j), 1 \leq j \leq n$ to be transmitted from i^{th} antenna where $i \in \{1, 2\}$.

The transmitted signals $\mathbf{S}(j), i \in \{1,2\}, 1 \leq j \leq n$ undergo through two independent flat fading channels and are combined at the receive antenna of a mobile user. The same signal model is applied to all mobile users with each user receiving the transmitted signals from two independent channels (This is shown for first User 1 and last User T , respectively for k^{th} group in Figure 5.8). The receiver for each user first despreads the received signal $r_{kl}(t)$ and collects the resulting signal over two symbol period to form soft estimates of users data signals $y_{kl}(j)$ and $y_{kl}(j+1)$. Next, these signals are space-time decoded to form the estimated composite codeword of kl^{th} user, $\hat{s}_{kl}(j)$ and $\hat{s}_{kl}(j+1)$ are obtained. Due to the orthogonal design of space-time coding, each users' data bit estimates are expected to achieve full two order diversity. Next, from each n estimated signals $\hat{s}_{kl}(j)$ and $\hat{s}_{kl}(j+1)$, the final estimate of the user's transmitted data bit b_{kl} is obtained using maximum likelihood method of collaborative decoding.

5.4.2 Transmitter Design

This subsection describes the set of processes involved both at the base-station transmitter and mobile users' receivers. First of all the signal model used for collaborative spreading is presented and then the signal processing stages of the proposed transmitter diversity scheme is discussed ; which is divided into four parts. The first part is space-time transmission operation performed at the transmitter of collaborative spread signals and the other three processing stages are carried at individual users' receivers.

Collaborative Spreading: The idea of collaborative spreading and transmission is revisited and employed here using a baseband model of a downlink chip synchronous DS-CDMA system (see Figure 5.1). It consists of a base-station transmitter for $K = GT$ users (equal power assumed for simplicity) employing the proposed collaborative spreading scheme communicating with K users' receivers via their fading channels $g_{kl}^{(i)}, i \in \{1,2\}$ with AWGN. The data signal of each user in k^{th} group $b_{kl} \in \mathbf{b}_k$ is first encoded with collaborative code $C_{lq} \in \mathbf{W}$ each of length n bits and BPSK mapped to form coded symbols υ_{lj}; where, $b_{kl} = \sum_{i=-\infty}^{\infty} b_k(i)p(t - iT_b)$ is the data signal, $\mathbf{W}$ is the set of codewords of all users, $b_k(i)$ is a binary data sequence taking values $[1, 0]$ with equal probabilities, $p(t)$ is a rectangular pulse with period T_b, . The combination of codeword signals of the T users in the k^{th} group can be shown as:

$$s_k(j) = \sum_{l=1}^{T} \upsilon_{kl}(j), 1 \leq j \leq n. \tag{5.22}$$

Each composite codeword signal $s_k(j)$ is spread using a distinct orthogonal spreading sequence $\mathbf{c}_k$. The signals of all G groups of users are summed to form a composite transmit signal $\mathbf{S}(j)$, which can be written as:

$$\mathbf{S}(j) = \sum_{k=1}^{G} s_k(j)\mathbf{c}_k, 1 \leq j \leq n, \tag{5.23}$$

where $\mathbf{c}_k$ is the k^{th} user group's common spreading sequence used for all the T users within k^{th} group with chip values $[-1, +1]$ and rectangular pulse $p(t)$ of period T_c. The spreading factor used is same as the number of groups and defined as $G = T_b/T_c$.

C-STS Transmission: The transmit diversity schemes employing space-time codes are very effective methods to obtain full diversity gain from multiple antenna systems [2]. The transmission scheme we propose here is based on the simple space-time Alamouti codes and is presented in matrix form as follows:

$$\mathbf{S} = \begin{bmatrix} \mathbf{S}(j) & \mathbf{S}(j+1) \\ -\mathbf{S}^*(j+1) & \mathbf{S}^*(j) \end{bmatrix},$$

where rows and columns represent signal periods and antennas, respectively and $\mathbf{S}^*$ denotes complex conjugate of $\mathbf{S}$. Under this scheme, in the first bit period, the composite signals $\{\mathbf{S}(j), \mathbf{S}(j+1)\}$ are transmitted from antenna 1 and 2, respectively. In the second period, the composite signals $\{\mathbf{S}^*(j+1), -\mathbf{S}^*(j)\}$ are transmitted from antenna 1 and 2, respectively. The received signals are wideband spread spectrum signals and hence at the receiver for each user, first the signals are detected using CDMA despreading.

5.4.3 Receiver Design

Without loss of generality, it is assumed that kl^{th} user (l^{th} user in k^{th} group) is the desired user. It is also assumed that transmit channels for users are non-dispersive and remain constant over a codeword period of n. The received signal $r_{kl}(t)$ for the kl^{th} user is written as:

$$r_{kl}(t){=}g_{kl}(t)S(t) + n_{kl}(t), \tag{5.24}$$

where $g_{kl}(t) = \alpha_{kl}(t)e^{j\phi_{kl}(t)}$ is the channel gain with amplitude $\alpha_{kl}(t)$ and phase components $\phi_{kl}(t)$ and $n_{kl}(t)$ is the AWGN with two sided power spectral density $N_0/2$. The transmit channels of users are described in details in the next subsection.

CDMA Despreading: At the user's receiver first the composite received signal $r_{kl}(t)$ is chip matched filtered to form two received signal vector $\mathbf{r}_{kl}(j)$ and $\mathbf{r}_{kl}(j{+}1)$. Then signals are despread using the group assigned spreading sequence $\mathbf{c}_k$ to obtain the soft estimates y_{kl} of the transmitted composite codeword signal $s_k(j)$ shown as follows

$$\begin{aligned} y_{kl}(j) &= \int_{(j-1)T_b}^{jT_b} \mathbf{r}_{kl}(j)\mathbf{c}_k^T; \\ y_{kl}(j+1) &= \int_{(j)T_b}^{(j+1)T_b} \mathbf{r}_{kl}(j+1)\mathbf{c}_k^T. \end{aligned} \tag{5.25}$$

The resulting signals $\{y_{kl}(j), y_{kl}(j+1)\}$ consist of transmitted signal s_k via two different channels $\{g_{kl}^{(1)}(j), g_{kl}^{(2)}(j+1)\}$ from antennas 1 and 2, respectively.

Space-Time Combining: The signals $\{y_{kl}(j), y_{kl}(j+1)\}$ are sent to a combiner that use orthogonal design of space-time decoding, which ensures that the symbol level estimates of transmitted codeword signals of the desired kl^{th} user are separated without interference effects and the copies are maximum ratio combined by following process:

$$\begin{aligned}\hat{s}_k(j) &= \left[y_{kl}(j)\left\{g_{kl}^{(1)}(j)\right\}^* + y_{kl}^*(j+1)g_{kl}^{(2)}(j)\right] \\ \hat{s}_k(j+1) &= \left[y_{kl}(j)\left\{g_{kl}^{(2)}(j)\right\}^* - y_{kl}^*(j+1)g_{kl}^{(1)}(j)\right],\end{aligned} \tag{5.26}$$

where $\{.\}^*$ denotes a complex conjugation operation.

Collaborative Decoding: The receiver performs joint detection and ML decoding of users' codewords and simultaneously provides the estimates of transmitted data signal $\hat{b}_{kl}$ of the desired kl^{th} user. The squared distance metrics of the space-time combined signals $\hat{s}_k(j); 1 \leq j \leq n$ with each combination of codeword $\upsilon_{kl}^{(q)}(j); 1 \leq j \leq n$ is denoted as d_{kq}. The distance metrics are calculated by utilizing the estimates of users' corresponding channels $\hat{g}_{kl}^{(1)}(j)$ and $\hat{g}_{kl}^{(2)}(j)$ for each combination of codeword as follows

$$\begin{aligned}d_{kl}^{(q)} = \sum_{j=1}^{n} \left| \hat{s}_k(j) - \left\{|\hat{g}_{kl}^{(1)}(j)|^2 + |\hat{g}_{kl}^{(2)}(j)|^2\right\}\left\{\sum_{l=1}^{T} \upsilon_{kl}(j)\right\}_q \right|^2 \\ \forall q, 1 \leq q \leq L.\end{aligned} \tag{5.27}$$

The calculated distance metric $d_{kl}^{(q)}$ for each combination is used to perform ML decoding such that the one that minimizes the distance metric is selected as the transmitted set of codewords of the collaborating users

$$\{\hat{C}_{kl}\} = \arg \min_{C_{1q}, C_{Tq} \in \mathbf{W}} d_{kl}^{(q)}. \tag{5.28}$$

Finally, the estimate of data symbol of the desired user $\hat{b}_{kl}$ is obtained by demapping the codeword to a symbol as used at the transmitters.

5.4.4 Extension to Quadrature Signalling

Achieving higher data throughput is very important in downlink CDMA schemes and therefore it becomes essential to extend the proposed scheme using a more spectrally efficient data signalling schemes. The proposed scheme can be extended to increase the throughput by the factor of two using quadrature data signalling and then applying collaborative coding to user's data in both in-phase and quadrature (IQ) domain. This method is termed here as IQ-C-STS to distinguish from the one dimensional or binary encoded C-STS case. Under this method, the previously described C-STS technique is applied independently to both real and imaginary part of each user's transmitted signal. The receiver for each user performs the operations detailed in 5.4.3 and then 5.4.3 to obtain complex codeword signals for the user. The signal model for the quadrature scheme can be followed easily from the equations (5.22)-(5.28) with appropriate modification to encoding and decoding schemes at the transmitter and a receiver, respectively. The final decoding of user's transmitted codewords are obtained by performing collaborative decoding separately on in-phase and quadrature channels.

$C_1+C_2+C_3$		(C_3) (0 0)	(C_3) (1 0)
(C_1+C_2)	(0 1)	0 1	1 1
	(1 0)	1 0	2 0
	(1 2)	1 2	2 2
	(2 1)	2 1	3 1

Table 5.5: Three-user uniquely decodable codes with $R_{sum} = 1.5$ bits per channel use

5.4.5 Performance Results

A baseband model of an uplink synchronous CDMA and BPSK modulated users employing Walsh Hadamard sequences with $N = 16$ are used. There are three collaborators $T = 3$ sharing a single spreading sequence and simulations are carried out for full system load conditions. The channel is a Rayleigh flat fading and assumed constant over a codeword period $n = 2$. A T-user collaborative codes as in the Table 5.5 is used for encoding users' data within each group.

In Figure 5.9, the BER performance of the proposed collaborative space-time spreading scheme is obtained and compared with conventional collaborative spreading scheme under above system settings. It can be clearly seen from the Figure that the use of transmit antenna diversity significantly improves the BER of each user. For the purpose of comparison, the BER of a single user system with Alamouti space-time scheme is also presented. It can be observed that the proposed scheme achieves full two order spatial diversity and is approximately 2 dB inferior to the Alamouti scheme.

Next, the BER of the proposed scheme is compared with that of conventional collaborative spreading, but with users employing multiple antenna receive diversity. As can be seen from the Figure 5.10, the BER of the proposed scheme is 3 dB worse than that of collaborative spreading scheme with 2 antenna receive diversity. While the proposed scheme shows a loss in SNR compared to that with multiple antenna receiver, from the view of practicality in implementation, the scheme is much more efficient and cost effective.

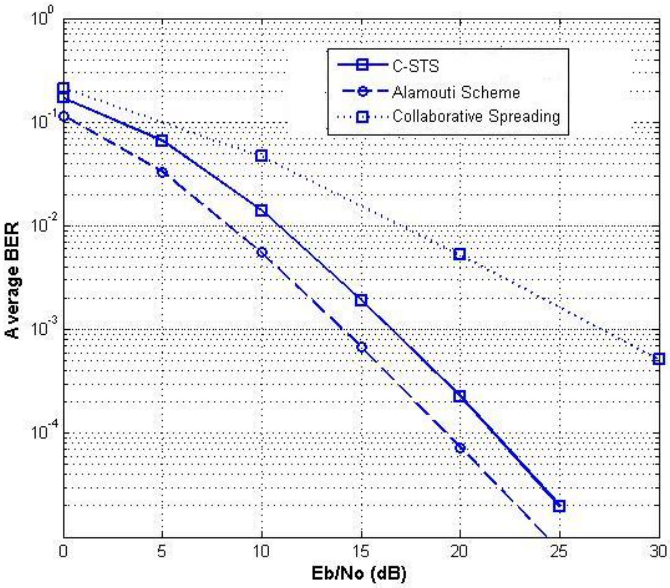


Figure 5.9: BER performance of C-STS for downlink of CDMA for $K = 16 \times 3$ users, and Walsh-Hadamard sequences of N=16

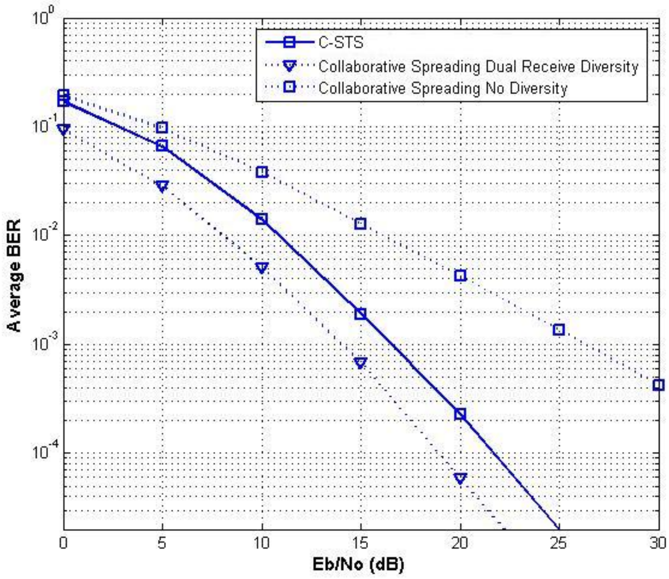


Figure 5.10: BER performance of C-STS for downlink of CDMA for $K = 16 \times 3$ users, and Walsh-Hadamard sequences of N=16

Finally, the BER performance of quadrature signalling based C-STS scheme denoted as 'IQ-C-STS' is presented in Figure 5.11. As expected the BER obtained has the same two order diversity performance and only around 2 dB worse compared with Alamouti scheme with QPSK data. The additional throughput obtained from this scheme increases the complexity of the receiver by $2L$ FLOPS and can be easily afforded for the significant performance gain of the collaborative scheme.

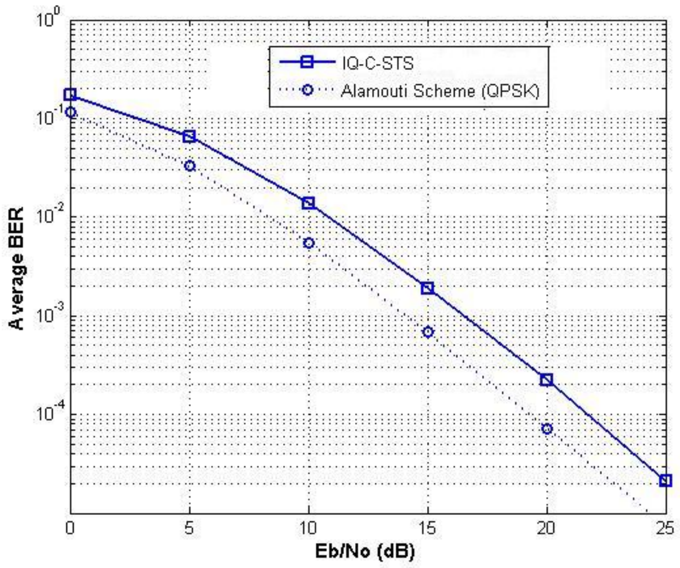


Figure 5.11: BER performance of IQ-C-STS for downlink of CDMA for $K = 16 \times 3$ users, and Walsh-Hadamard sequences of N=16

From these results, it can be noted that the CS-CDMA-DL has the same characteristics of Non-overloaded orthogonal CDMA such as maintenance of high SINR, same diversity gain under multiple antenna systems. To sum up, the contributions so far have addressed the user capacity and error performance for the downlink of CDMA. The same problems exist in the uplink and is equally interesting and challenging. In the next section, a new contribution is presented addressing these important issues.

5.5 Collaborative Spreading for CDMA Uplink (CS-CDMA-UL)

5.5.1 Introduction

In uplink of CDMA, where orthogonality among users is very difficult to maintain, MUD techniques [28] that remove detrimental effects of MAI are usually employed to improve the user capacity. It is however well known that the maximum number of users supported with the MUD is usually less than the number of available spreading sequences. Since radio spectrum is a scarce resource, there is an increasing interest in techniques to support an overloaded system. In this regard, CDMA is also known to have a soft capacity limit, such that more users than the dimension of the spreading sequences can be supported but at the expense of some error performance degradation [25]. In [130], a new method for multiuser detection of users in an overloaded CDMA system where sequences assigned to users are linearly dependant is presented in AWGN channel conditions. Performance analysis of an overloaded group orthogonal CDMA scheme that uses a form of collaborative mapping of users data in AWGN environment suitable for a downlink is investigated in [131]. Also it is shown for example in [25] and [50] that, with improved design of interference cancellation subclass of multiuser receivers, user capacity higher than the number of sequences can be achieved. The use of multiple antenna elements and beamforming at the central receiver also known as SDMA to achieve the high capacity is investigated for example in [33].

Recently, MIMO technique that uses multiple antennas both at transmitter and receiver e.g. in [35, 117, 32, 104] is becoming increasingly popular. It exploits the available spatial dimensions to improve the performance of point to point wireless transmissions. The scheme in [32], also known as VBLAST assumes full rank of MIMO channels to perform several layers of multiuser detection to achieve the high throughput. Motivated by gains promised by the MIMO, there has also been an interesting work known as MIMO-SCDMA to increase the user capacity of uplink CDMA [132]. It uses multiple antenna space-time spreading transmission at user terminals and a receiver with a single or several antennas employing a linear MUD technique. Also, Space-Time MIMO based schemes for CDMA for example in [104], [117] are proposed to improve the spatial diversity gain. Since the use of multiple antennas as in MIMO is not that practical for mobile terminals due to their size and power limitations, an idea of cooperative diversity/communications, e.g. in [7], [8] is becoming widespread recently, where transmissions of more than one independent user are coordinated to share their single antennas to achieve higher order spatial diversity in fading wireless channels. Recently, a group based scheme called GO-MC-CDMA is proposed in [133]. In this scheme users within a group share a small number of subcarriers by using individual separate spreading sequences of very short length so as to be able to afford the complexity of ML multiuser detection for each group. The scheme combines the benefits of improved diversity and MAI free performance of CDMA and OFDMA, respectively and is shown to support a number of users up to available signal dimensions (subcarriers).

It is well known that users' channels may not always be independent and can often become highly correlated, for example due to the nonzero probability of having direct line of sight conditions with the receiver [30], [134]. Under such conditions where the full rank condition is hardly satisfied, parallel transmission of independent data as in MIMO which rely solely on independent channels may result in severe detection ambiguity and hence the reduced throughput. When multiple users wish to access the same channel without any assumptions on their channels, a different approach as used in [14, 12, 126] may be used. These schemes make use of a special way of superposition of users' signals to decode their data without relying on subdivision in time, frequency or orthogonal codes. This is achieved by simply assigning individual codewords to encode data of users such that the received composite signal is uniquely decodable. The separation of users' signals is achieved from the unique decodability of the received composite signal. Also referred to as CCMA, its total sum rate and error performance under different channel conditions are investigated in [14, 12, 126].

In this contribution, we propose a new collaborative spreading transmission and reception scheme to increase the user capacity of uplink of CDMA without using multiple antennas at either sides and referred to as CS-CDMA-UL. It achieves the said gain by exploiting the differences in users' channel magnitudes or unique multiple access codes to allow more than one user (T-user) to share the same spreading sequence. To recover the data, the receiver uniquely and more efficiently exploits the capabilities of existing CDMA MUD techniques to reduce MAI from composite signals of all co-spread users. It has to be noted that the reduced MAI at the despreader output due to reuse of sequences however comes at the expense of increased CCI. The remaining MAI is further suppressed using a linear MUD e.g. a decorrelation stage for each group rather than for each user, which can eliminate the MAI for all users and in effect can save significant amount of computational complexity. Since there are only a small number of co-channel users to be detected and an ML detection scheme is easily affordable. Therefore, with this new approach $K = GT$ full rate users may be supported using G sequences and with a little increase in SNR to that of G users using the conventional CDMA approach. Based on this idea, two possible methods for using the proposed scheme are given in this section.

The first method we term as Uncoded CS-CDMA-UL, does not employ any form of collaborative coding as in [14]. Instead, it exploits independent fading magnitudes of users' channels which allows for data of co-spread users to be separated as long as their channels are not highly correlated. This condition is easily satisfied in most practical wireless environments due to independent path loss and fading of each user's channel. The proposed collaborative CDMA approach can also be thought somewhat similar to as having several independent parallel data streams originated from a single source with separate transmit antennas or spatial multiplexing of multiple antenna MIMO transmissions [102]. In contrast to the MIMO based CDMA schemes e.g. [132], [104], [117], the proposed scheme utilizes single antenna transmitters and receiver only. Unlike the cooperative schemes in [7], [8], it does not involve exchange of data between group of users (which usually requires more than one symbol period) and hence does not incur any bandwidth efficiency loss compared to a system without cooperation. In a slightly different system context but with the same objective, we also propose the second method simply termed as CS-CDMA-UL and corresponding detection scheme in different channel conditions. The scheme employs collaborative codes for each T-user group and the same codes are reused for all groups to increase the overall user capacity while also to solve the ambiguity problem. For example, by using a code with two set of codewords and assigning them separately to encode two users' data, we can significantly increase the number of simultaneous users in the system irrespective of users' channel correlation conditions. It should also be noted here that different codes can be used for different groups for further system design optimization.

The Section is organized as follows. In subsection 5.5.2, system model of the technique is presented. The joint collaborative multiuser receiver scheme is described in subsection 5.5.3 and practical issues are discussed in 5.5.4. The BER performance analysis is carried out in section 5.5.5. Subsection 5.5.6 shows the simulation results and comparisons with conventional CDMA.

5.5.2 System Model

A generalized system model of a synchronous CS-CDMA-UL with BPSK modulated users' signals with fading MAC and AWGN is shown in Figure 5.12. The assumption of a synchronous uplink is used here to simplify the key concepts and performance analysis. There are a total of K users in the system, which are divided into G groups with T users within each group using a total of G spreading sequences and the users within a group are assumed to share the same spreading sequence. The received signal that is chip matched filtered and sampled to form the received vector $\mathbf{r}(j) = [r_1(j), r_2(j), .., r_N(j)]^T$, where $[.]^T$ denotes a transpose operator. For the purpose of our study, it is sufficient to consider $\mathbf{r}(j)$ within the period of n symbols, which is given by

$$\mathbf{r}(j)=\sum_{l=1}^{T}\sum_{k=1}^{G} h_{kl}(j)(t)\mathbf{s}_{kl}(j) + \mathbf{w}(j); \qquad 1 \leq j \leq n, \tag{5.29}$$

where $\mathbf{s}_{kl}(j) = \sqrt{P_{kl}}\upsilon_{kl}(j)\mathbf{c}_k(j)$ is the transmitted signal of l^{th} user within k^{th} group with power P_{kl}, $\upsilon_{kl}(j)$ is the data signal, $\mathbf{c}_k(j)$ is the spreading sequence with chip values containing $[-1, +1]$

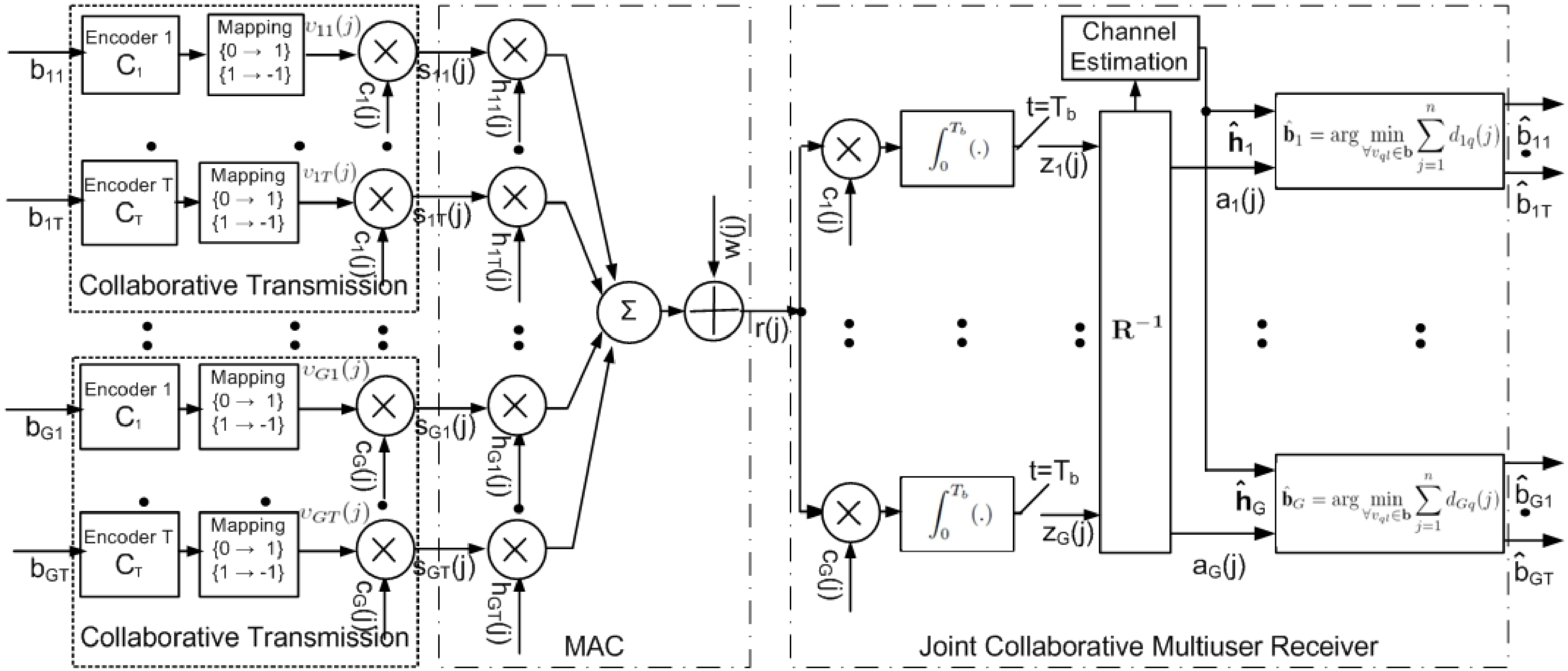


Figure 5.12: Generalized System Model for CS-CDMA-UL

C_1+C_2		(C_1) $(0\ 0)$	$(1\ 1)$
	$(0\ 0)$	0 0	1 1
(C_2)	$(0\ 1)$	0 1	1 2
	$(1\ 0)$	1 0	2 1

Table 5.6: A two-user collaborative codes and allowable codeword combinations with $R_{sum} = 1.29$ bits per channel use

using rectangular pulses of period T_c with processing gain $N = T_b/T_c$ with normalized power over the symbol period $\int_0^{T_b} c_k^2(t) = 1$ and $\mathbf{w}(j) = [w_1(j), w_2(j), .., w_N(j)]^T$ is vector of AWGN with two sided power spectral density $N_0/2$. The signals from all users are assumed to be transmitted via flat fading channels, which remain constant over at least a single codeword period n, for kl^{th} user, this can be written as

$$h_{kl}(j) = \alpha_{kl} e^{j\phi_{kl}}; 1 \leq j \leq n, \tag{5.30}$$

where $h_{kl}(j)$ is modeled as complex Gaussian random variable with a zero mean and unit variance $\sigma_h^2 = 1$ with α_{kl} the amplitude and ϕ_{kl} phase component. To aid the detection, channel of each user has to be estimated from the received signal. In this work, we do not use a specific channel estimation method, but employ a generalized channel model with the estimation error and residual interference signals assumed to be zero mean complex additive Gaussian noise. The channel estimation of kl^{th} user $\hat{h}_{kl}(j)$, is generated as follows

$$\hat{h}_{kl}(j) = h_{kl}(j) + e_{kl}(j), \tag{5.31}$$

where $e_{kl}(j)$ is the estimation error signal of zero mean and variance σ_e^2. Perfect channel estimation corresponds to the case when the error variance is zero i.e. $\sigma_e^2 = 0$. Note that although we are employing simple flat fading channel model for the purpose of clear exposition of our techniques, they can be easily extended to operate in frequency selective channels with use of multicarrier techniques which successfully convert the channels in to a number of flat fading subchannels.

Compliant with the signal model, we also briefly describe the principles of the collaborative coding and show how we are able to decode data signals of more than one user on a single channel without use of any multiple access scheme. Consider a system of T users, transmitting independent data on a common multiple access channel (MAC). Each user $l, 1 \leq l \leq T$ is assigned a set of N_l codewords from the collaborative codes $\mathbf{C}_l = \{C_{l1}, C_{l2}, ..., C_{lN_l}\}$, of equal length n symbols. The data of each user are encoded using codewords from the set $\mathbf{C}_l$, then modulation mapped using linear digital modulation technique to form encoded signals $\{\upsilon_l(1), \upsilon_l(2)..\upsilon_l(j).., \upsilon_l(n)\}$. Note that for Uncoded CS-CDMA-UL, $n = 1$ and therefore data symbols i.e. b_l rather than coded signals are transmitted i.e. $\upsilon_l(j) = b_l$. The received signal is the output of MAC consisting of sum of each user's codeword signals possibly with some added noise. The total achievable sum rate R_{sum} in bits per channel use of this coding scheme [12, 14] is given by

$$R_{sum} = \sum_{l=1}^{T} \log_2(N_l)/n. \tag{5.32}$$

For example, the channel output for $T = 2$ with $N_1 = 2, N_2 = 3, n = 2$ is shown in Table 5.6 for all codeword combinations. As can be seen from the table, the composite signals resulting from the combining of each user's codeword are unique and a single decoder can perfectly unscramble the resulting signal to deliver the users' codewords and hence their data streams. It can be noted from the Table, that the sum rate achieved for these codes is $R_{sum} = 0.5 + 0.79 = 1.29$ bits per channel use. This well known unique decodability property of collaborative coding scheme is the means for achieving higher user capacity for one of the proposed techniques.

5.5.3 Joint Collaborative Multiuser Receiver

The proposed receiver is shown in Figure 5.12, which is designed as such to allow effective recovery of users' transmitted data by gradually removing the effects of the MAI and the CCI by using multiple stages of signal processing. First, the wide-band received signal is despread at every symbol period to obtain soft estimates $z_k(j)$ of grouped users' transmitted data signals. With this process, in a system with $K = GT$ users, variance of MAI at despreader output is only $\frac{(G-1)}{N}$ compared with $\frac{(GT-1)}{N}$ of conventional CDMA, when random spreading sequences are used [36]. Next, the stage using a linear decorrelation technique [29] indicated in the figure as '$\mathbf{R}^{-1}$' is used to remove the MAI component from $z_k(j)$. It should be noted that any other type of CDMA multiuser detection technique such as e.g. SIC, PIC or DF as studied in [47] can also be employed for the removal of MAI. The signals at the decorrelator outputs $a_k(j)$, are used within joint ML detection stage for obtaining final estimates of co-spread users' data. For estimating users' encoded data i.e. under CS-CDMA-UL, the aforementioned processes are extended by detecting $a_k(j), 1 \leq j \leq n$ over a codeword period of $n \geq 2$ symbols. Detailed description of processes of the proposed multiuser receiver is described next.

Despreading: At the initial stage, bank of correlators or MF matched to G groups' sequences are employed, this is in contrast to the conventional CDMA receivers which require GT correlators, i.e. one for each user. The received signal $\mathbf{r}(j)$ is first despread with using spreading sequences $\mathbf{c}_k(j), \forall k$ to form a vector of output signals $\mathbf{z}(j) = [z_1(j), z_2(j), ..z_G(j)]^T$, which can also be written as

$$\begin{aligned} \mathbf{z}(j) = \mathbf{R}(j)\underline{\mathbf{H}}(j) + \underline{\mathbf{w}}(j) \\ 1 \leq j \leq n, \end{aligned} \tag{5.33}$$

where $\mathbf{R}$ is the cross correlation matrix of dimension $G \times G$, formed by used spreading sequences with each element denoted as $\rho_{ku} = c_k c_u, k = 1, 2, ..G; u = 1, 2, ..G$, $\underline{\mathbf{w}} = [\underline{w}_1, \underline{w}_2, .., \underline{w}_G]^T$ is the vector of the correlated noise terms with variance σ_n^2 and covariance matrix $N_0\mathbf{R}$. The product of users' data and channels in matrix $\underline{\mathbf{H}}(j)$ can be shown as follows:

$$\underline{\mathbf{H}}(j) = \begin{bmatrix} h_{11}\upsilon_{11}(j) & h_{12}\upsilon_{12}(j) & \dots & h_{1l}\upsilon_{1l}(j) & \dots & h_{1T}\upsilon_{1T}(j) \\ h_{21}\upsilon_{21}(j) & h_{22}\upsilon_{22}(j) & \dots & h_{2l}\upsilon_{2l}(j) & \dots & h_{2T}\upsilon_{2T}(j) \\ \vdots & \vdots & \ddots & \vdots & \ddots & \vdots \\ h_{k1}\upsilon_{k1}(j) & h_{k2}\upsilon_{k2}(j) & \dots & h_{kl}\upsilon_{kl}(j) & \dots & h_{kT}\upsilon_{kT}(j) \\ \vdots & \vdots & \ddots & \vdots & \ddots & \vdots \\ h_{G1}\upsilon_{G1}(j) & h_{G2}\upsilon_{G2}(j) & \dots & h_{12}\upsilon_{12}(j) & \dots & h_{GT}\upsilon_{GT}(j), \end{bmatrix}$$

where each row consists of T users' independent data and channel signals within the k^{th} group, which can also be shown as separate composite group signals $\underline{h}_k(j)$ as follows

$$\underline{h}_k(j) = \sum_{l=1}^{T} h_{kl}\upsilon_{kl}(j). \tag{5.34}$$

The data and channel signal vectors consisting of T entries each in a group sharing the k^{th} sequence, can also be written as

$$\Upsilon_k(j) = [\upsilon_{k1}(j), \upsilon_{k2}(j), .., \upsilon_{kT}(j)]^T; \mathbf{h}_k = [h_{k1}, h_{k2}, .., h_{kT}]^T.$$

Since elements in each row of $\underline{\mathbf{H}}(j)$ consist of signals of a user group forming composite signals $\underline{h}_k(j)$, the matrix $\underline{\mathbf{H}}(j)$ can be reduced in to a vector $\underline{\mathbf{h}}(j)$ given by $\underline{\mathbf{h}}(j) = [\underline{h}_1(j), \underline{h}_2(j), ...\underline{h}_G(j)]^T$, where each element is co-channel users' composite signal that consists of the sum of products of their individual data and channel gain values. Also it is noted in (5.33) is that, although there are GT users in the system, the size of noise vector is only $1 \times G$.

Decorrelation: Because the soft output signals $\mathbf{z}(j)$ generated by despreading only do not provide reliable estimates of the transmitted users' data, a group decorrelation stage is incorporated in the receiver. This is achieved by multiplying $\mathbf{z}(j)$ with the inverse of the cross correlation matrix $\mathbf{R}^{-1}$, the resulting output signal vector $\mathbf{a}(j)$, can be shown as

$$\mathbf{a}(j) = \mathbf{R}^{-1}(j)\big[\mathbf{R}(j)\underline{\mathbf{H}}(j) + \mathbf{w}(j)\big] = \underline{\mathbf{H}}(j) + \mathbf{R}^{-1}(j)\underline{\mathbf{w}}(j). \tag{5.35}$$

It can be noted that, the vector $\mathbf{a}(j)$ is free from MAI; however, it suffers from noise enhancement problem that scales linearly with increase in the number of users [29], [28]. As shown in (5.35), variance of the noise term $\mathbf{w}^{Dec}(j) = \mathbf{R}^{-1}(j)\underline{\mathbf{w}}(j) = [w_1^{Dec}(j), w_2^{Dec}(j), ..w_G^{Dec}(j)]^T$ is now greater than that of the original noise $\mathbf{w}$. We will show the degradation of output signal to noise ratio (SNR) caused by this process. The Gaussian assumption of noise and residual interference signals at the output is then used to derive BER of the proposed system.

The elements of the vector $\mathbf{a}(j) = [a_1(j), a_2(j), ..a_G(j)]^T$ are sent to G bank of collaborative joint detectors to detect and separate T co-spread users' data, the signal elements $a_k(j)$ can also be shown as the sum of the constituent users' signals and the noise term as follows

$$a_k(j) = \sum_{l=1}^{T} h_{kl}\upsilon_{kl}(j) + \eta_{kl}(j) \tag{5.36}$$

where, $\eta_{kl}(j) = w_k^{Dec}(j) + e_{kl}(j)$ is the total noise, which is still a complex Gaussian random variable consisting of decorrelator noise $w_k^{Dec}(j)$ and channel estimation error signal $e_{kl}(j)$.

ML Joint Detection: With channel estimates available, the collaborative joint detection method calculates the maximum *a posteriori* probabilities (MAP) for all the L possible transmitted data or codeword vectors. The detector provides the final estimate of users' data by selecting a vector which maximizes the probability. For the data signalling scheme with equal probability of occurrence for all possible symbols, the process is simplified and receiver makes the decision based on ML or the minimum distance criterion. For T co-channel users employing M-ary data symbols, there are $L = M^T$ possible data symbol combinations. The distance for both cases for each q^{th} possible vector is calculated as follows

$$d_{kq}^2(j) = \left| a_k(j) - \sum_{l=1}^{T} \hat{h}_{kl}\upsilon_{ql}(j) \right|^2 \qquad 1 \leq q \leq L. \tag{5.37}$$

Based on the set of calculated distances $d_{kq}^2(j)$, the receiver makes decision on the possible transmitted data as the vector $\mathbf{b}_q$ which yields the minimum sum distance shown as follows:

$$\hat{\mathbf{b}}_k = \arg\min_{\mathbf{b_q}\in\mathbf{B}} \sum_{j=1}^{n} d_{kq}^2(j); \left\{ \begin{array}{ll} n = 1 & \text{, Uncoded CS-CDMA-UL} \\ n \geq 2 & \text{, CS-CDMA-UL} \end{array} \right. \tag{5.38}$$

For uncoded case i.e. $n = 1$ and BPSK user signals with $M = 2$, there will be $L = M^T$ possible distances to calculate from the list of data vectors $\{\mathbf{b}_1, \mathbf{b}_2, ..\mathbf{b}_L\} \in \mathbf{B}$. For the coded case i.e. $n \geq 2$, there are $L = \prod_l^T \log_2(N_l)$ allowable codeword combinations. The calculated distance d_{kq}^2 for each allowable codeword combination is used to perform ML decoding such that the combination $\sum_{l=1}^{T} \left\{\upsilon_{kl} \in \mathbf{C}_l\right\}$ that minimizes the sum of distances $\sum_{j=1}^{n} d_{kq}^2(j)$, is selected to give estimates of transmitted codewords of the T collaborating users. The data estimates $\hat{\mathbf{b}}_k$ are obtained by demapping the codewords using the the same mapping table used in the users' transmitters.

5.5.4 Practical Considerations

Synchronization: For successful operation of the proposed scheme as shown in Figure 5.12, essential tasks such as chip timing synchronization, channel estimation must be carried out. The base-station receiver uses precise clock to obtain delay estimate of incoming signals from all users with respect to the reference timing say τ_0. Next, the initial spreading code acquisition or delay estimation of a user $\hat{\tau}_{kl}$ has to be carried out to synchronize the incoming signal with the locally generated copy of the sequence c_k [89]. This is generally achieved by correlating the received signal with the local sequence to first obtain the coarse delay estimation within a chip interval $\hat{\tau}_{kl} \leq T_c$. The estimate is further refined with the precision of small fraction of a chip period to lock the signal and then tracking process is carried out. The code acquisition scheme for uplink CDMA using difference signature sequences and in practical fading environments are investigated in [135], [89], [90] and can also be used within our system. It has to be noted that the imperfect delay estimation can however severely effect the performance of linear MUD receivers [136].

When $K > N$ i.e. as system becomes overloaded, the available spreading sequence has to be shared by $T > 1$ users. To successfully acquire and lock to users' signals, the base-station can assign individual sequence with good autocorrelation property such as Barker sequence [90] and then use periodically inserted orthogonal pilots sequences $u_{kl}, 1 \leq l \leq T$ of length F within each user's data so that the effects of $T - 1$ co-channel users can be minimized [103]. By correlating the incoming signal with $\mathbf{c}_k$ as shown earlier as in (5.33), the receiver can extract the desired kl^{th} user's transmitted signal $\tilde{\mathbf{s}}_{kl}$ from the correlator output as follows

$$\tilde{\upsilon}_{kl}\tilde{h}_{kl} = \frac{1}{F}\sum_{f=1}^{F}\left\{\mathbf{r}^T(f)\mathbf{c}_k(f)\right\}u^*_{kl}(f) \Longrightarrow \hat{\tau}_{kl} = \max\left\{\delta_k(\tilde{\tau}_{kl})\upsilon_{kl}h_{kl} + MAI + noise\right\}^2, \tag{5.39}$$

where $\delta(.)$ is the Kronecker delta , $\{\}^*$ is the complex conjugation operator, MAI is the multiuser interference from other groups to the k^{th} group and $noise$ is the AWGN. From (5.39) the receiver can obtain a rough estimate of the delay $\tilde{\tau}_{kl}$ by selecting the value that gives the maximum. The acquisition process is then accompanied by the code tracking to facilitate subsequent signal processing tasks.

Since decision of T users' data is obtained using their composite signal, synchronous reception of the users is very desirable to retain the maximum signal energy. This requires a mechanism to adjust timing of users' transmissions based on feedback from the receiver on their arrived signals. A useful algorithm investigated in [135] may be used here for this purpose. The technique allows users to either delay or progress their transmissions within a small fraction of chip period by sending timing control bits (TCB) based on received delay estimates. After acquiring the codes and T-user synchronization for all G groups, the usual CDMA detection including the linear MUD can take place.

Channel Estimation: Linear receivers such as correlators and decorrelators have the useful property of not requiring explicit channel estimation [28]. The ML joint detection stage however requires channel estimates of T users to recover their data. For the collaborative coded case, it also requires synchronization to be maintained over the codeword length n. This should not be a problem as the chip and symbol synchronization is already achieved before the CDMA detection. Moreover techniques for CCMA codeword and symbol synchronization investigated in [127], [128] can be used without difficulties. The channel estimates are usually obtained after MAI suppression at the output of MUD [47] and with the use of periodically inserted pilot or training sequences [116], [103]. The quality of channel estimation i.e. its error variance is usually dependent on system parameters such as pilot to data power ratio, user loading environments, type of multiuser detection schemes [137], [47]. As noted earlier, the channel estimate of kl^{th} user can also be obtained at the correlator output as in (5.39). However, for the improved performance the estimate $\hat{h}_{kl}$ is obtained by collecting signals at the decorrelator output over the period of F symbols and followed by averaging, given by

$$\hat{h}_{kl} = \frac{1}{F}\sum_{f=1}^{F} a_k u_{kl}^{*}(f) = h_{kl} + \eta_{kl}, \tag{5.40}$$

where η_{kl} is the total estimation error signal as in (5.36). The estimate $\hat{h}_{kl}$ is not effected by $T-1$ users' signals but it is effected by the decorrelator noise and possible error due to non-ideal channel estimation filter. The effect of imperfect channel estimation is an important practical issue and will be presented later in the results section.

Computational Complexity: It is noted that ML joint detection requires search over $L = M^T$ possible combinations of data vectors. Since both M and T are very small numbers the additional complexity is still very low. Under collaborative coded case, the output signal a_k is the superposition of co-spread user' codewords which is multilevel as can be seen from Table 5.7. The decoding of codewords involves search over a small number $L = \prod_{l=1}^{T} \log_2(N_l)$ of allowable composite codewords of length $n, n \geq 2$. As far as the overall receiver complexity is concerned, although there are $K = GT$ users, the proposed scheme requires despreading and decorrelation operations for only G groups. Assuming a decorrelator MUD, approximate computational complexity is $O\big[GN + G^2N + LG\big]$ FLOPs per symbol decision compared with $O\big[GTN + (GT)^2N + GT\big]$ using conventional CDMA. The FLOPs counts are taken for despreading, decorrelation and final symbol decision operations, respectively. It can be noted here that the proposed technique can provide higher user capacity also at a much lower computational complexity.

5.5.5 BER Analysis

In this Section, we derive the BER for a user with the proposed scheme in Rayleigh fading environment by using the Gaussian Approximation method [36] for interference and noise signals at the output of the decorrelator as proposed in [28]. Without any loss in generality we assume that kl^{th} user is the desired user. Since the combined MAI and noise signals are jointly Gaussian distributed, the probability of bit error of any user within k^{th} group, $Pe(a_k|_{h_k})$ conditioned on channel realisation $h_{kl}(j)$ can be obtained from the SINR of signals for each time instance $SINR_k(j)$, which can be written as

$$Pe(a_k|_{h_{kl}}) = Q\left(\sqrt{SINR_k(j)}\right), \tag{5.41}$$

where $Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt$. The average error probability $P_e(a_k)$ under the channel fading is then obtained by calculating (5.41) over all the fading SNR distribution of the user, which is given by

$$Pe_k = \int_0^\infty Pe(a_k|_{h_{kl}}) p(\gamma_k) d\gamma_k, \tag{5.42}$$

where $p(\gamma_{kl})$ is the PDF of SINR distribution conditioned on the fading magnitude of the desired user. Next, we derive BER for the two cases of CS-CDMA-UL with $n = 1$ and $n \geq 2$.

Case 1: Uncoded CS-CDMA-UL ($n = 1$): The average SINR at the output of the decorrelator $SINR_k^{Dec}$, for a G user system compared to the single user SNR, $SNR_0 = E\{\alpha_{kl}^2\}/N_0$, can be obtained using a method given in [28] as follows

$$\overline{SINR_k^{Dec}} = \left[\frac{E^2\{|a_k|/T\}}{var\{a_k\}}\right] = E\Big\{SNR_0\big(1 - \mathbf{f}_k \mathbf{R}_k^{-1} \mathbf{f}_k^T\big)\Big\}, \tag{5.43}$$

where $E\{.\}$ is the expectation operator to account for possible random distribution of users' amplitude and cross correlation of sequences, $\mathbf{f}_k$ is the k^{th} column of $\mathbf{R}$ without the diagonal element and $\mathbf{R}_k^{-1}$ is obtained from $\mathbf{R}^{-1}$ by discarding from it the elements of k^{th} row and column. The average BER at the decorrelator output can be obtained using the standard Gaussian Q-function for the signal at each time instance j i.e. $Pe(j)$. The average BER for the kl^{th} user at the output of joint detector can be obtained using the SINR at the output of the decorrelator as in (5.43), where the SNR loss due to $T - 1$ co-spread users in the desired group has to be considered.

For this purpose, an average distance measure of data vectors is defined. Based on the minimum distance between two vectors d_{min}^2, the average distance $\overline{d}^2$ is calculated as follows:

$$\begin{aligned} TE_b &= \frac{1}{2^T} \sum_{v=1}^{2^T} i_v d_{min}^2 \\ \Longrightarrow \overline{d}^2 &= \frac{2^T T E_b}{\sum_{v=1}^{2^T} i_v}, \end{aligned} \tag{5.44}$$

where E_b is the energy per data bit of a user and i_v is a ratio by which distance from the 1^{st} vector to v^{th} vector is greater than the minimum distance d_{min}^2. The $\overline{d}^2$ value is used to calculate the relative SNR loss λ in dB, compared with fully orthogonal signalling with BPSK i.e. $\lambda = 10\log\left\{\overline{d}_0^2/\overline{d}^2\right\}$ dB, where $\overline{d}_0^2 = 4$. This distance measure can be used along with the SINR at the decorrelator output to obtain an approximate probability of error for a user. By averaging $Pe(j)$ over all the fading distribution $p(\gamma_k)$, the average probability of error Pe_{kl}^{Dec} can be approximated as follows

$$Pe_{kl}^{Dec} \approx \frac{1}{2}\left[1 - \sqrt{\frac{\left(\overline{d}^2/\overline{d}_0^2\right)\overline{SINR_k^{Dec}}}{1+\left(\overline{d}^2/\overline{d}_0^2\right)\overline{SINR_k^{Dec}}}}\right]. \tag{5.45}$$

Case 2: CS-CDMA-UL ($n \geq 2$): The probability of bit error analysis of the CS-CDMA-UL signals using BPSK mapping is considered here using Euclidian distance between different allowable composite codewords. This is developed from our previous analysis for uncoded case and the effect of collaborative coded signals on the BER can be assessed by doing some modifications to the tools developed for existing single user per sequence approach. For this purpose, first we derive distance metric associated with each distinct codeword combination of T users within a group. The absolute magnitude of squared distance between each unique codeword combinations are calculated and normalized with n to give a distance metric d_i^2 as follows

$$d_v^2 = \frac{1}{n}\left|\left\{\sum_{l=1}^{T}\overline{C}_l\right\}_x - \left\{\sum_{l=1}^{T}\overline{C}_l\right\}_y\right|^2 \tag{5.46}$$
$$1 < v < V; 1 \leq x \neq y \leq L,$$

where $\left\{\sum_{l=1}^{T}\overline{C}_l\right\}_x$ and $\left\{\sum_{l=1}^{T}\overline{C}_l\right\}_y$ are any two distinct modulated codeword combinations (overlines are used here to denote the modulated codewords) of T users, and $V = \sum_{m=1}^{L-1} m$ is the total number of possible distance paths between two unique codeword combinations. Since all codewords are equally likely, an average distance metric of all unique codeword combinations can be obtained and which can be used within tools to analyse the performance of standard single user per sequence based transmissions. Averaging d_i^2 over all V possible distances, we obtain the average distance $\overline{z}$, given by

$$\overline{d}^2 = \frac{\sum_{v=1}^{V} d_v^2}{V}. \tag{5.47}$$

Using $\overline{z}$, we can now calculate the probability of bit error of CS-CDMA-UL system over fading channels with the $SINR_k^{Dec}(j)$ obtained from (5.45). The probability of error Pe_{kl}^{CCol} for a user under CS-CDMA-UL can be obtained as follows

$$Pe_{kl}^{CCol} \approx \frac{1}{2}\left[1 - \sqrt{\frac{\left(\overline{d}^2/\overline{d}_0^2\right)\overline{SINR_k^{Dec}}}{1+\left(\overline{d}^2/\overline{d}_0^2\right)\overline{SINR_k^{Dec}}}}\right]. \tag{5.48}$$

Note the approximation used in (5.48) is due to the calculation of Pe_{kl}^{CCol} based on the average distance of all allowable codeword combinations. A simple minimum distance and union bound based approach as in [30] can also be used to calculate the upper bound on the BER.

5.5.6 Performance Results

In this Section, we present simulation performance results of the proposed CS-CDMA-UL schemes and compare under the same system settings and channel conditions with Conventional CDMA approach of using single sequence for each user. Synchronous uplink with K equal power users and Gold sequences with spreading factor of $N = 31$ is assumed. The results for the two schemes under the proposed system are divided into two parts: Uncoded CS-CDMA-UL and CS-CDMA-UL, respectively. Two well known CDMA detection techniques within the proposed receiver schemes are investigated : simple matched filters and decorrelators. The results of CS-CDMA-UL using these receiver techniques are denoted as CS-CDMA-UL (MF) and CS-CDMA-UL (DCOR), respectively and the same notations applies also to the uncoded case.

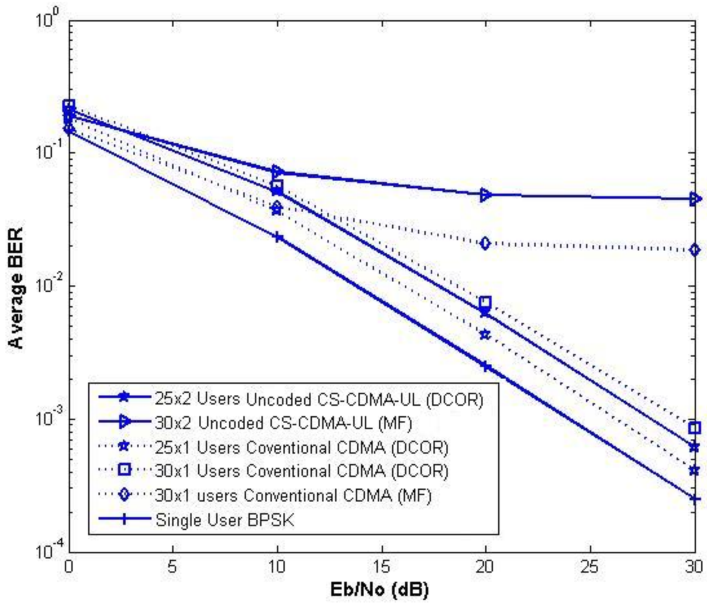


Figure 5.13: BER performance of Uncoded CS-CDMA-UL and Conventional CDMA in Rayleigh fading channels with perfect channel estimation for $G \times T$ users and Gold sequences of $N = 31$

5.5.6.1 Uncoded CS-CDMA-UL

User capacity: Figure 5.13 shows the BER performance of the proposed technique and conventional CDMA under the same settings in flat Rayleigh fading channels with perfect channel estimation conditions. The BER simulation performances of conventional CDMA with 25 and 30 users employing a decorrelator are also shown. In the proposed scheme, $G = 25$ and $T = 2$ are chosen corresponding to total number of users $K = G \times T = 50$. The results clearly show the superior BER performance of the proposed Uncoded CS-CDMA-UL (DCOR) compared with Conventional CDMA (DCOR). It can support 50 users at the same BER performance compared with approximately 27 users. This improved performance comes at no extra power or computational cost and with significantly less number of spreading sequences. The result can also be verified from the theoretical analysis, which predicts a SNR loss $\lambda = d_0^2/\bar{d}^2 = 4/2.66 \approx 1.8$ dB compared to that with single user detection for $K = 25$ users. When matched filters (MF) only (i.e. without the MUD) are employed, the BER of both schemes are significantly worse as expected. The gap in the BER of conventional and the proposed technique is wider in this case compared to that with decorrelators. This is because MF outputs generate much poorer estimates for the collaborative joint detection of T users' data.

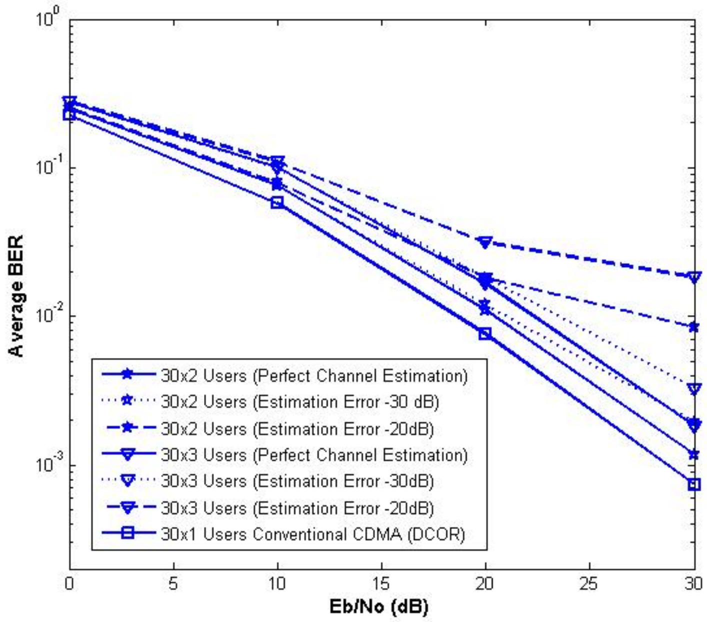


Figure 5.14: BER performance of Uncoded CS-CDMA-UL (DCOR) in Rayleigh fading channels with perfect channel estimation (solid lines) and with estimation errors (broken lines) for $G \times T$ users and Gold sequences of $N = 31$

The effect of group size T and imperfect channel estimation: In Figure 5.14, we first show the BER of Uncoded CS-CDMA-UL (DCOR) using a fixed $G = 30$ and with different values of T under perfect channel estimation conditions (shown in solid lines). The BER curves are obtained for: $a) \quad T = 1, b) \quad T = 2, c) \quad T = 3$, respectively. As can be seen from the figure, the proposed technique exhibits satisfactory BER performance even when T is increased from 2 to 3 and shows only a modest SNR penalty of $\approx 2dB$ for each additional co-spread users in a group. The performance under imperfect channel estimation conditions (shown in dotted lines) are also shown for $G = 30$ and $T = 2, 3$ with error variances $\sigma_e^2 = -30dB$ and $-20dB$, respectively. It can be noted that, the proposed technique shows quite robust performance under low estimation error variances. The performance is shown to degrade gradually, for example when the variance increases from $-30dB$ to $-20dB$. Nevertheless, it can be observed that the receiver shows the robust performance under high system loading even with considerable estimation error.

For example, with $\sigma_e^2 = -20dB$ and $K = 30 \times 3$ users, the receiver maintained a reasonable BER of 0.02 at the $E_b/N_0 = 20dB$. The BER can easily be improved using error correction coding techniques.

The fading channels of the collaborating users may not always be i.i.d. Rayleigh distributed which was assumed in the simulations so far. In practice some amount of fading correlation may exist which can lead to reduced distance measure for ML joint estimation of the co-spread users' data in (5.38). The correlated fading envelopes of the users can be obtained by partially shifting the waveform of a user and use that waveform for the other user. Also there exist several techniques for generating such waveforms e. g. [138, 139]. However, in this work a simple Rician channel model is used to assess the effects of fading correlation and also impact of static channel conditions.

Performance under Rician fading and synchronization error conditions: As a part of this performance study, we model the users' channels as complex random variables with Rice distribution, while leaving rest of the system settings unchanged. Under such channel conditions, the intensity of fading is also known as the Rician factor κ, which is defined for kl^{th} user as $\kappa_{kl} = \mu_{kl}^2/2\sigma_y^2$, where μ_{kl}^2 is the power of direct line of sight component and $2\sigma_y^2$ is the total power of many weaker diffused components. The Rician model is known to represent Rayleigh fading and AWGN channels as special cases with $\kappa = 0$ and $\kappa = \infty$, respectively.

Figure 5.15 shows the BER performance under different Rician factors κ with perfect channel estimation. It is noted that the BER degrades gradually as κ increases i.e. $\kappa > 0$. The performance results under such conditions can be explained as follows: Under low or zero values of κ, the transmitted signals of each users are highly likely to undergo through distinct channels, thus maintaining a satisfactory distance measures for the collaborative joint detectors. Conversely, when the channel is AWGN like, due to lack of scattering the detectors suffer from ambiguity due to higher probability of having equal distance metrics. This phenomenon is noted to be very similar to the effect on the degraded capacity of MIMO wireless communications [134] in Rician channels due to lack of scattering. To address this problem, we investigate the system employing collaborative coding and its performance under different channel conditions.

Shown also in the Figure are the BER of the scheme under Rayleigh fading channels $\kappa = 0$ and with imperfect synchronization conditions. The error signal is assumed to be uncorrelated with the received signal and modeled as an additive complex Gaussian random variable with standard deviation in fraction of a chip period [47]. As can be seen from the figure, the synchronization has significant impact on the BER of the proposed scheme. As expected, proposed technique similar to linear MUD schemes is also sensitive to the synchronization error [136].

Impact of receive diversity: The performance of Uncoded CS-CDMA-UL with diversity recep-

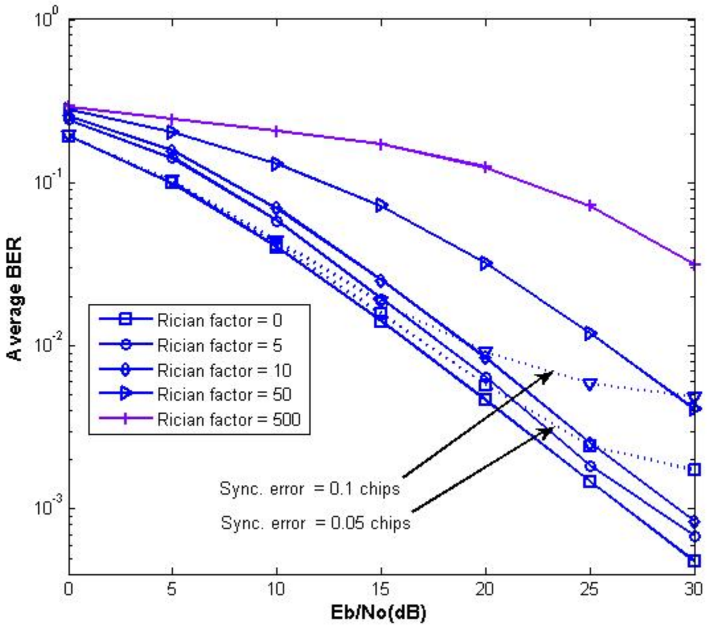


Figure 5.15: BER performance of Uncoded CS-CDMA-UL (DCOR) in fading channels with different Rician factors (κ) for 15×2 users and Gold sequences of $N = 31$

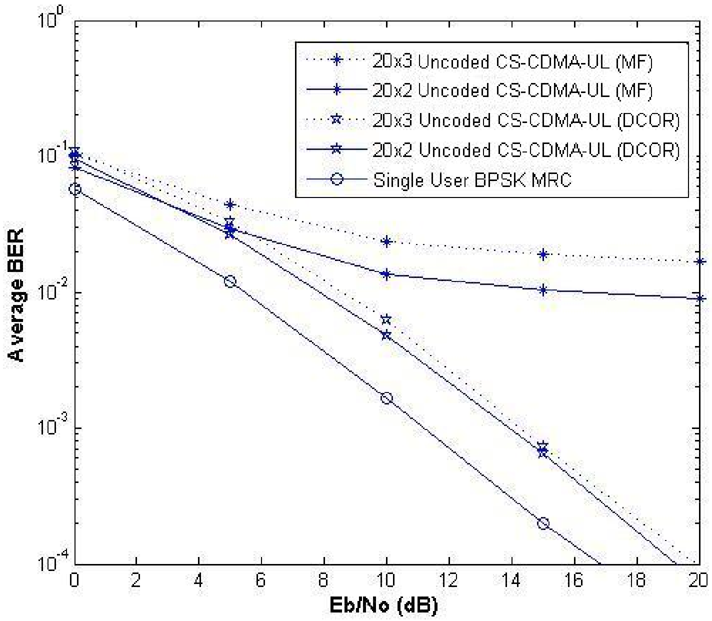


Figure 5.16: BER performance of Uncoded CS-CDMA-UL (DCOR) with dual receiver diversity in Rayleigh fading channels with perfect channel estimation for $K = 20$ and [T=2 (solid lines), T=3 (dotted lines)] users and Gold sequences of $N = 31$

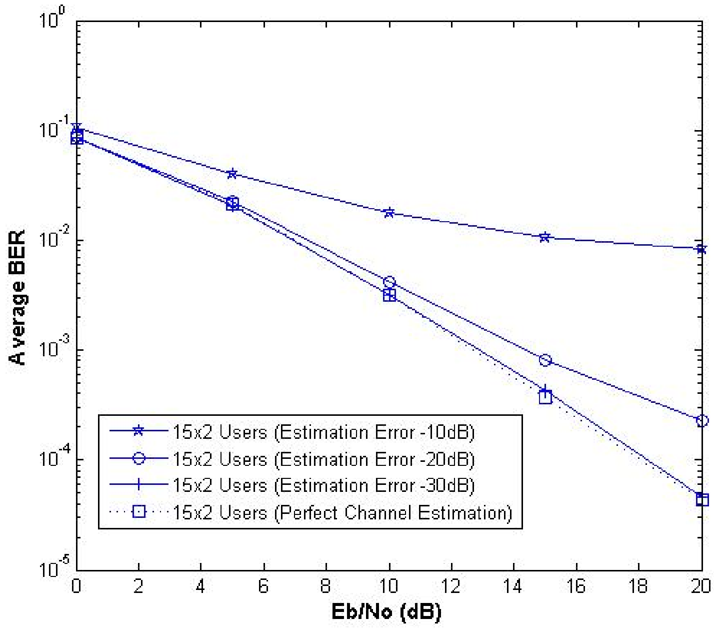


Figure 5.17: BER performance of Uncoded CS-CDMA-UL (DCOR) with dual receive diversity in Rayleigh fading channels with imperfect channel estimation for $K = 15$ and $T = 2$ users and Gold sequences of $N = 31$

tion with perfect channel estimation is investigated. The BER performance for the case of $K = 20$ and $T = 2, 3$ are shown in the Figure 5.16. As expected the conventional MF receivers that do not consider the effect of MAI show degraded BER performance even with diversity reception. The employment of decorrelator frontend significantly improves the BER performance under the same system loading conditions. In fact, with diversity reception, it can be seen that the BER performances of Uncoded CS-CDMA-UL with 30×2 and 30×3 users shows almost identical performance. This implies that the gap in BER due to additional users sharing the same sequences can be narrowed by using diversity reception. The use of transmit diversity can also be applied equally well and the similar BER results are expected.

Effect of channel estimation on the diversity reception: The introduction of diversity showed very robust performance in fading channels in Figure 5.17. The BER performance of the system under imperfect channel estimation condition is now investigated. The channel estimation model used is as given in (5.31). It is evident that the performance of the system degrades as the variances of channel estimation error increases.

5.5.6.2 CS-CDMA-UL

User capacity: Figure 5.18 shows the BER performance of CS-CDMA-UL and Conventional CDMA under the same system settings in flat Rayleigh fading channels with perfect channel estimation conditions. For the proposed scheme, $G = 25$ and $T = 3$ are chosen, where BPSK modulated collaborative codes of length $n = 2$ as shown in Table 5.7 are used. This also corresponds to total number users $K = G \times T = 75$ each of $1/n$ rate. For fair comparison with Conventional CDMA, this effectively translates to $K = (G \times T)/n = 37.5$ full rate users. The BER performances of Conventional CDMA for 25 and 30 users are also shown. Simulation results clearly show that at the same BER, the proposed CS-CDMA-UL (DCOR) supports significantly higher number of users. In terms of gain in full rate users, it can support 37 users with lower BER performance than that of Conventional CDMA with 30 users. This improved performance comes at very little added computational cost while requiring much less i.e. only $G = 25$ sequences. Again, the BER result can also be verified from the theoretical analysis, which predicts a SNR loss $\lambda = d_0^2/\overline{d}^2 = 4/2.4 \approx 2$ dB compared with that of conventional scheme with 25 users and can also be observed from the figure.

$C_1+C_2+C_3$		(C_3)	
		(0 0)	(1 0)
(C_1+C_2)	(0 1)	0 1	1 1
	(1 0)	1 0	2 0
	(1 2)	1 2	2 2
	(2 1)	2 1	3 1

Table 5.7: Three-user uniquely decodable codes with $R_{sum} = 1.5$ bits per channel use

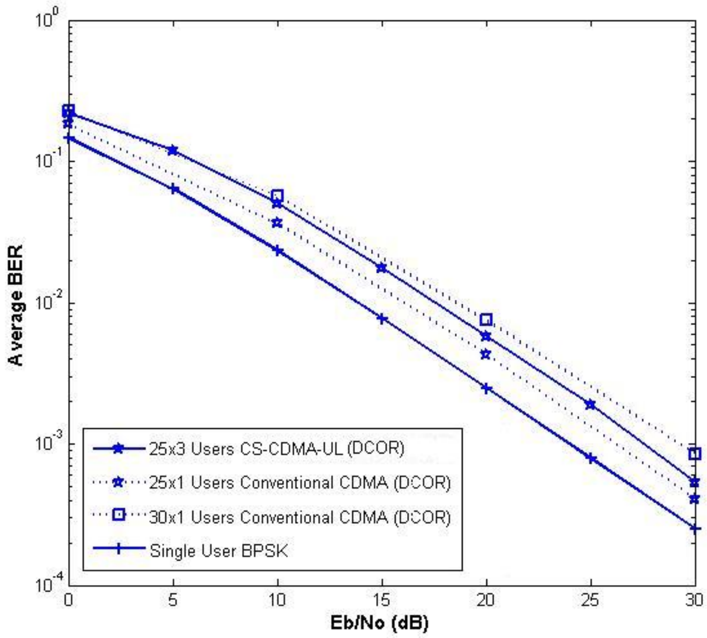


Figure 5.18: BER performance of CS-CDMA-UL in Rayleigh fading channels with perfect channel estimation for $G \times T = 25 \times 3$ user and Gold sequences of $N = 31$ using collaborative codes from Table 5.7

Performance under Rician fading and imperfect channel estimation: Figure 5.19 shows the

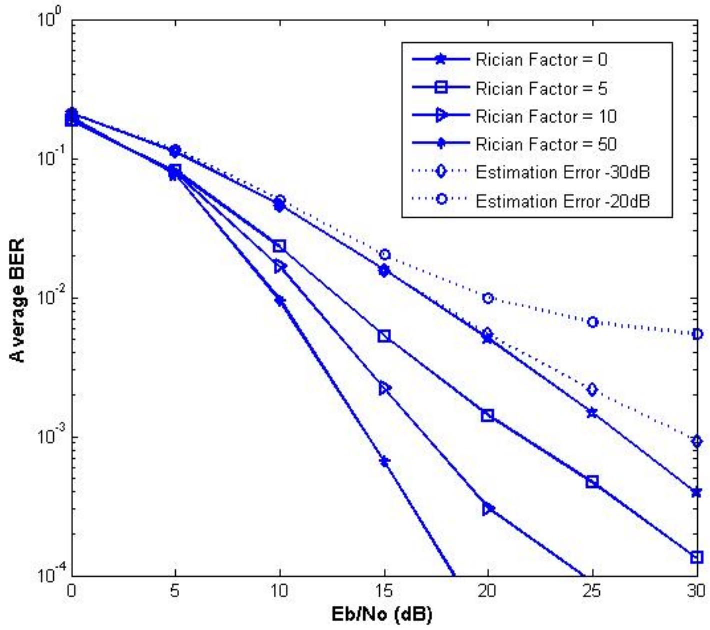


Figure 5.19: BER performance of CS-CDMA-UL (DCOR) in different Rician factors κ under perfect channel estimation and with different channel estimation errors variances (σ_e^2) in Rayleigh fading $\kappa = 0$ for $G \times T = 20 \times 3$ and Gold sequences of $N = 31$ using collaborative codes from Table 5.7

BER performance of the proposed scheme under Rician channels with different κ and with perfect channel estimation (shown in solid lines). As expected, it shows much improved BER when the Rician factor increases $\kappa > 0$. We have already mentioned the ambiguity problem of uncoded transmission under the correlated fading channels and argued that collaborative coding solves the ambiguity. It is evident from the figure that under the CS-CDMA-UL, users' data can be recovered regardless of channel correlation or Rician conditions. Also in the same Figure, we show the BER performance of the scheme in Rayleigh fading ($\kappa = 0$) with imperfect channel estimation conditions (shown in dotted lines) under estimation error variances $\sigma_e^2 = -30dB$ and $-20dB$. As can be noted, the proposed system shows robust BER performance under low estimation error conditions, which gradually degrades when error variance increases from $-30dB$ to $-20dB$. Nevertheless, the receiver showed quite reasonable BER under high system load with moderate estimation errors. For example with $\sigma_e^2 = -20dB$ and $K = 20 \times 3$ users, it maintained the BER of ≈ 0.01 at $E_b/N_0 = 20dB$.

Impact of receive diversity: The performance of CS-CDMA-UL (DCOR) is investigated with diversity reception with perfect channel estimation. The BER performance for the case of $K = 20$ and $T = 3$ are shown in the Figure 5.20. The employment of decorrelator frontend significantly improves the BER performance under the same system loading conditions. As can be noted from the figure, the SNR gain of $\simeq 12dB$ can be obtained at the BER of 10^{-3} with dual diversity

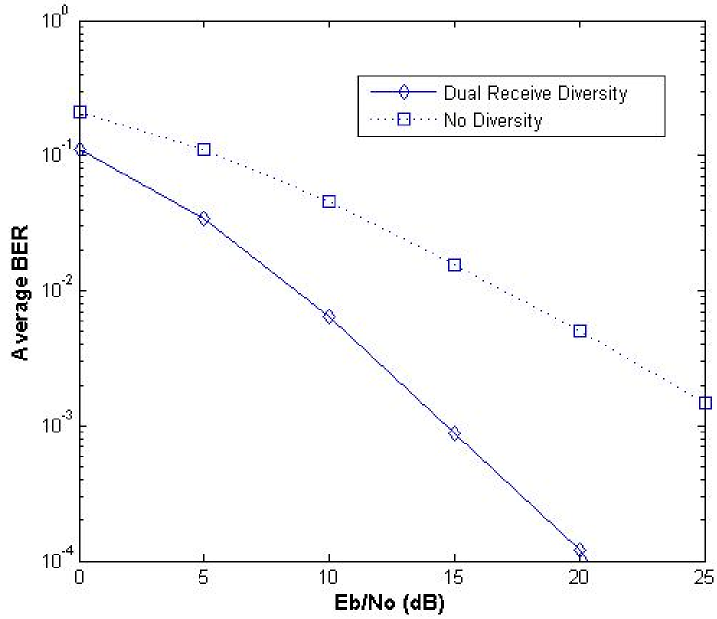


Figure 5.20: BER performance of CS-CDMA-UL (DCOR) in Rayleigh fading channels under perfect channel estimation with dual receive diversity for $K = 20$ and $T = 3$, using Gold sequences of $N = 31$ and collaborative codes from Table 5.7

reception.

Effect of channel estimation of the diversity reception: The introduction of diversity showed very robust performance in fading channels in Figure 5.21. The BER performance of the system is now investigated under imperfect channel estimation conditions. The channel estimation model used is as given in (5.31). It is noticed that the performance of the system degrades as the variances of channel estimation error increases. It is also noted that the system is more resilient to the channel estimation error compared to no-diversity case as in figure 5.19 for the same system loading. It can be concluded that the important system capacity (number of users supported) is retained for the proposed system with some degradation in BER when non-ideal channel estimation techniques are used.

Effect of diversity reception under Rician fading: The introduction of diversity showed very robust performance in Rayleigh fading channels. Figure 5.22 shows the BER performance of the CS-CDMA-UL (DCOR) system under Rician fading channel conditions with different K-factors. It is observed that the increase in K-factor i.e. increase in the power of line of slight component shows much improved BER performance compared that under pure Rayleigh fading (K-factor=0). For K-factor=50, the SNR gain of around 9 dB is observed compared to that in Rayleigh fading.

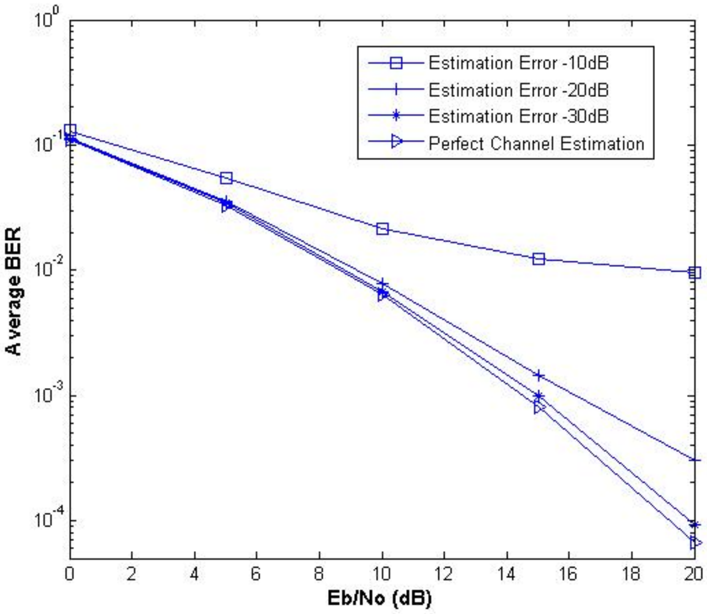


Figure 5.21: BER performance of CS-CDMA-UL (DCOR) in Rayleigh fading channels under imperfect channel estimation with dual receive diversity for $K = 20$ and $T = 3$, using Gold sequences of $N = 31$ and collaborative codes from Table 5.7

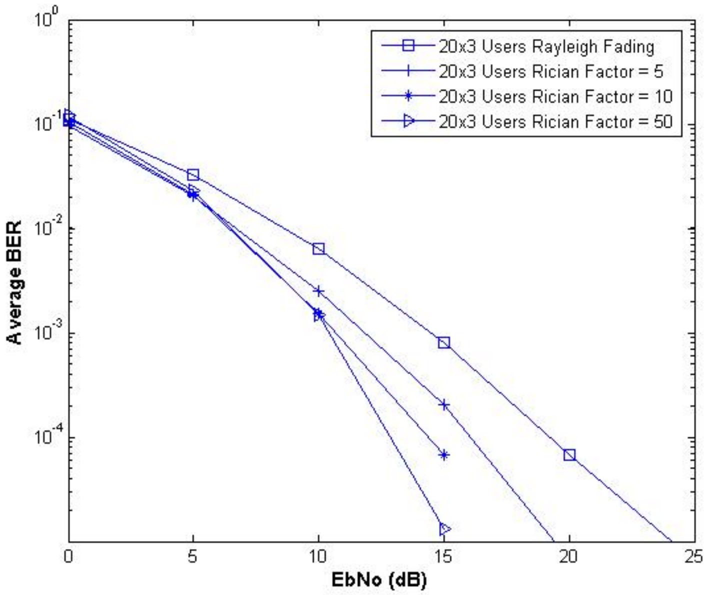


Figure 5.22: BER performance of CS-CDMA-UL (DCOR) in Rician Fading channels under different K-factors (κ) with dual receive diversity for $K = 20$ and $T = 3$ using Gold sequences of $N = 31$ and collaborative codes from Table 5.7

5.6 Chapter Summary

This Chapter focused on investigating new techniques to improve the performance of downlink and uplink of CDMA. First, the basic concepts of user collaboration to achieve higher number of users for CDMA uplink is presented in Section 5.2. It is discussed that independent channels of users can be used for collaborative spreading to save considerable amount of system resources with such scheme compared with conventional single sequence per user approach. Furthermore, how collaborative coding and decoding techniques can be employed to resolve the detection ambiguity under high correlation of co-spread users' channels is also briefly noted.

In Section 5.3, a new collaborative spreading technique for downlink of overloaded CDMA termed as CS-CDMA-DL is proposed and analysed. It is shown to achieve higher number of users using very simple encoding and decoding methods. Theoretical and simulation analysis of the BER, user capacity performance of this scheme is obtained and compared with various techniques. For example, using orthogonal sequences with spreading factor of 64, a total of 192 half rate users can be simultaneously supported at ≈ 2.2 dB increase in SNR compared with fully orthogonal CDMA. It can achieve higher SINR compared with other schemes such as OCDMA/OCDMA, and supports much higher overloading ratio with lower complexity receiver than group orthogonal CDMA schemes. Furthermore, the BER performance is also evaluated in Rayleigh flat fading environment with imperfect channel estimation.

Although, the performance the CS-CDMA-DL shows a similar BER performance as of non overloaded orthogonal CDMA, its performance in fading channel conditions is not satisfactory for many demanding applications such as multimedia information transfers or realtime applications. To improve the performance, a new technique called Collaborative Space-Time Spreading or C-STS is introduced by using multiple antennas at the base-station in Section 5.4, to provide diversity gain for mobile users. The BER performance of the scheme is also compared to that with non-overloaded fully orthogonal transmission scheme and that of dual receive diversity under the same channel conditions. For example, the technique is shown to achieve the same BER of the Alamouti scheme for only ≈ 2 dB increase in SNR giving substantial gain under same transmitted power compared with a single antenna scheme.

Finally, in Section 5.5 we proposed a new transmission and receiver technique referred to as CS-CDMA-UL to increase the user capacity and support an overloaded uplink CDMA and with a low complexity receiver. By grouping a small number of users to share the same spreading sequence and performing CDMA detection per group rather than user followed by an ML joint detection of grouped users allowed us to achieve higher number of users also with low computational requirements. Extensive simulation results show that, the Uncoded CS-CDMA-UL technique offers substantial increase in user capacity compared with conventional CDMA under the same power and BER requirements. For example, using Gold sequences of length 31, it supported 50 full rate users compared with approximately 27 users. The use of collaborative coding i.e. CS-CDMA-UL is shown to give similar user capacity gain and significantly improved BER particularly under highly correlated channel conditions. It is also shown that most of the capacity gain can still be retained under moderate channel estimation error conditions. Investigations are underway to design a new receiver scheme using a RAKE structure. Also, the use of multicarrier based CDMA transmission and corresponding multiuser detection scheme is left as an interesting topic for further research.

Chapter 6

Conclusions and Future Work

Higher spectral efficiency and user capacity are the main objectives for the design of practical multiuser wireless systems. The thesis has made a number of new contributions towards meeting these objectives for system employing CCMA and CDMA techniques. Following this course, new schemes that combine the benefits of the two techniques are proposed and evaluated. An outline of the main conclusions and recommendations for future investigations are given below:

6.1 Conclusions

Chapter Three: This Chapter addressed the problem of estimation and cancellation of MAI in uplink of CDMA. Three new contributions are presented using a blind adaptive approach. A robust and high capacity successive interference cancellation scheme for DS-CDMA referred to as CMA-SIC is proposed in Section 3.2. It provides improved MAI supersession for the detection and more accurate amplitude estimation for the interference cancellation. An interesting feature of blind adaptive CMA algorithm is exploited in the CMA-SIC and the despreader weights are used in optimum way to reduce the error propagation problem of conventional SIC. The accuracy of channel estimation of SIC is evaluated in terms of MSE with respect to the users' true channels. Significant reduction in MSE compared with conventional SIC is also noted. The BER analysis is conducted assuming Gaussian distribution of MAI signals. The spectral efficiency analysis of the CMA-SIC in Rayleigh flat fading channel condition is also presented. It is revealed that approximately twice the number of bits per channel use can be achieved with the CMA-SIC compared with conventional SIC.

In Section 3.3, a blind adaptive PIC receiver termed as BA-PIC for DS-CDMA is proposed. It has been identified that with the new design of estimation and cancellation processes, the performance of the BA-PIC is significantly better compared with conventional PIC. In the proposed BA-PIC, initial detection of users' signal is carried out by performing adaptive despreading and further improvement is achieved by optimal interference cancellation in MMSE sense using the weights obtained due to constant modulus property of users transmitted signals. It has been demonstrated that the BA-PIC offers significant improvement in BER with just a single stage of interference cancellation in contrast to conventional PIC, which requires three stages of cancellation to achieve the same performance. Extensive simulation results of the BA-PIC under different nearfar condition with fading and non-fading provides a firm basis to conclude that PIC receivers are not vulnerable to nearfar condition as it has been thought commonly. A very interesting result of this thesis is that, the normalized spectral efficiency of the PIC with multiple stages of cancellation can be much higher than SIC receivers. Spectral spectral efficiency comparison of CMA-SIC and BA-PIC is carried out in flat Rayleigh fading channel condition. With three stages of cancellation, BA-PIC is shown to achieve a rate of ≈ 7.5 bits/s compared with ≈ 4.3 bits/s of CMA-SIC.

In 3.4, we introduced a new subcarrier combining scheme for detection of multiuser signals using MC-CDMA transmission scheme under multipath fading and multiuser interference. A low complexity blind adaptive algorithm for subcarrier combining is proposed and evaluated. It is shown to suppress interference and give significant improvement in the detection performance compared to that with MRC and EGC techniques.

Chapter Four: The Chapter focused on achieving spatial diversity from collaboration between users in the uplink of CCMA and CDMA. First in Section 4.2, a simple two antenna transmitter diversity scheme is proposed and analysed for CCMA. Each multiple-antenna user transmits its codewords by using one antenna at a time slot. It is shown that full diversity can be achieved by this scheme; however there is 3 dB penalty compared with receive diversity with the same number of antennas. BER results using a simple analysis based on calculation of average distances of composite codewords is also shown to closely match with simulations. Next, in Section 4.3, a new scheme to provide spatial diversity for CCMA with user collaboration is investigated. BER performance bounds and comparison with CCMA without collaboration and with receive diversity is given. When the degree of user collaboration or the inter-user channel power is increased, significant gain in BER compared with CCMA without collaboration is achieved. BER of the new scheme is shown to approach that of dual receive diversity performance under higher ratio of inter-user channels.

In Section 4.4, two new schemes using collaborative diversity and successive interference cancellation referred to as C-SIC and C-BASIC are proposed for uplink of CDMA. An asymptotic analysis of achievable rate of the C-SIC is also given. It is noted that the C-SIC offers much improved performance compared with C-MF, however it suffers from degraded BER as the system load increases due to residual interference. The C-BASIC scheme using of the improved CMA-SIC technique is then investigated to alleviate this problem. The C-BASIC shows higher attainable diversity and robust performance under high loading situations. Nearfar condition which usually leads to degraded performance in CDMA is found to be beneficial for the both collaborative schemes.

In Section 4.5, a novel bandwidth efficient collaborative diversity (BECD) scheme for multiple access channels is proposed and in particular, evaluated for uplink of CDMA. The scheme uses the concept of collaborative spreading by a small group of users to share the same spreading sequence and joint detection of grouped users' data at the base-station. Two approaches to BECD using simple signal superposition and space-time coding techniques are presented. With each group consisting of two users, it is shown that the BECD scheme can achieve full second order diversity at no extra bandwidth cost with increase in the SNR ratios of inter-user channels. In contrast to other collaborative schemes the BECD has an added advantage in that overloaded system can be supported by trading off some diversity gain.

Chapter Five: This Chapter introduced a new approach of 'User Collaboration' to increase the number of users in CDMA. In Section 5.3, a new scheme called as CS-CDMA-DL is proposed to support an overloaded downlink of orthogonal CDMA by using collaborative coding to allow the use of single sequence for more than one user. The CS-CDMA-DL scheme, compared to some other downlink CDMA schemes, has been shown to provide higher overloading ratio and SINR at a small increase of SNR compared with fully orthogonal CDMA. Simplified BER analysis using average distances of collaborative codewords is used along with simulations to corroborate the results. For example, using Walsh Hadamard sequences of length of 64, CS-CDMA-DL is shown to support 192 half rate users at only a 2.2 dB increase in SNR compared with fully orthogonal CDMA. To increase the data throughput, CS-CDMA-DL is also investigated using quadrature signalling. In Section 5.4, Collaborative Space-Time Spreading (C-STS) is investigated for the CS-CDMA-DL by incorporating the space-time coding technique. BER performance of the C-STS is evaluated and shown to give that of full second order diversity. An extension to quadrature data signalling is presented and evaluated to enhance the throughput.

In Section 5.5, a novel scheme for increasing the user capacity of uplink CDMA referred to as CS-CDMA-UL is proposed and analysed. The scheme allows a small number of users to share the same sequence and perform group MUD and low complexity ML joint detection to separate the co-spread users. It is demonstrated that number of full rate users can be increased significantly at no extra bandwidth, power or antennas with this approach. For example, with a spreading length of N=31, it can support 50 users at the same performance of 27 users of conventional one sequence per user approach. The theoretical BER analysis based on average distance of users' data is shown to match the simulated BER within 0.4 dB. With the use of collaborative coding the system is able to resolve the ambiguity problem in detection of the co-spread users in highly correlated (Rician) channel conditions. The practical problem of channel estimation error is also investigated and found to gracefully effect the system performance.

6.2 Future Work

Chapter Three: The blind adaptive approach proposed in this thesis can also be applied in many other applications such as adaptive delay estimation, synchronization etc. and will be useful to explore in the future. Also, there has been increasing interest in using Multicarrier CDMA techniques for operation under hostile multipath fading channels. One of the practical problems in such environment is imperfect channel estimation which may have severe effect on the detection performance and system capacity. It will be interesting to use this approach for interference estimation as well as subcarrier combining to fully exploit the benefits of multicarrier CDMA. It is expected that this blind method will yield improved estimates of users signals compared to conventional fixed MF correlation based detection methods. It would also be worthwhile to note that blind channel estimation and interference cancellation approach can be combined with turbo multiuser detection schemes [93], that use forward error correction with joint interference cancellation and iterative decoding to further exploit the temporal diversity of AWGN channels.

Chapter Four: Transmitter diversity techniques for CCMA and BER analyses are provided. An interesting configuration for this scheme will be to investigate a joint space-time and collaborative coding. Regarding the user collaborative diversity for CCMA, soft decision decoding methods for estimating partners' transmitted codewords can also be explored. C-SIC and C-BASIC are shown to be effective techniques to address the MAI problem and improving the collaborative diversity in uplink of CDMA. Theoretical analysis of BER performance of the schemes can be carried out in the future for different optimization purposes. Investigation of the scheme using different partner selection criteria such as minimum distance, maximum SNR etc. [109] would be useful to investigate. Study of other users' channels and interference statistics to mutual detection of collaborating users' data are also interesting to pursue in future work.

Chapter Five: The collaborative spreading introduced in this Chapter has demonstrated the effectiveness of the new collaborative approach in CDMA. Future work can be carried out to further investigate the synchronization and channel estimation aspects of the scheme to carry out different optimizations such as pilot to data power ratio, acquisition time etc.. User selection criteria for grouping based on channel state of each user to minimize the total transmitted power can also be explored. Design of receivers to exploit multipath using RAKE structure or multicarrier CDMA techniques are natural to further exploit the frequency diversity of wideband CDMA transmissions. A very useful study will be to see if we can achieve diversity while retaining the high user capacity by incorporating the ideas such as cooperative diversity [7], multicarrier transmission [140]. These techniques are becoming widespread for improving the error performance of users under various fading channel environments by using space and frequency dimensions. Since we have already demonstrated high number of users with the collaborative spreading in frequency flat channels, it is expected that the frequency diversity of channels could be well used to achieve the gain without additional bandwidth. New schemes for joint synchronization, channel estimation and multiuser user detection [89] are other interesting areas for future studies.

Appendix A: Simple and Power Efficient Semi-Complex Spreading Scheme for BPSK DS-CDMA

The appendix presents an additional work conducted during the early period of this research. A simple and power efficient spreading scheme for CDMA using simple correlation receivers is described next. The performance of technique is verified by simulations as well as theoretical analysis.

A.1 Introduction

One of the most important aims of CDMA is to increase the number of simultaneous users supported with acceptable error performance while keeping the transmitted power as low as possible. This is generally challenging in a system using single user Matched Filter (MF) receivers due to the presence of MAI causing graceful degradation of error performance. It is known that any technique that reduces the effect of the MAI can increase the user capacity of CDMA systems [37]. The multiuser receivers [28] that jointly detect all the users while suppressing the effects of MAI, is a very effective approach to the problem and has been the intensive area of research for last two decades. These receivers while show significantly improved performance compared to MF receivers[141], are also highly computationally complex. The problem of MAI can also be alleviated to great extent by design and employment of improved spreading sequences. Besides pseudo noise (PN) sequences used in the industry currently, different varieties of short complex (polyphase) spreading sequences [142], [143] with improved auto and cross-correlation properties are proposed in the literature [38].

In DS-CDMA, user data are usually spread using the real binary [141] or complex spreading schemes[144, 142]. The complex spreading scheme [143] usually employs two binary spreading sequences to spread the data in real and imaginary (quadrature) axises, or different phases for each real sequence. This work proposes a new spreading scheme for BPSK DS CDMA, where users' data are spread in either real or imaginary axis during each symbol period according to the precoding sequence known both to transmitter and receiver. This scheme while requires the same amount of power as the real binary spreading scheme in a single user channel, allows higher number of users to share the same multiple access channel (MAC) at the same target error performance. The scheme may also lead to further increase in the user capacity when multiuser receivers that further suppress the MAI [28], are employed.

The work is organized as follows. In section A.2, general system and channel model is described. The precoding scheme for the spreading user' data is presented in section A.3 and the bit error probability analysis is carried out in section A.4. The simulation results under AWGN and fading channel conditions are presented in section A.5.

A.2 System model

An uplink synchronous DS-CDMA system of K equal power BPSK users is considered. A base-band model of the system with the proposed scheme with K users' transmitters, the MAC with AWGN and a receiver for k^{th} user is shown in Figure A.1. The channel model used in this work is given by

$$\beta_k(t) = \alpha_k(t)e^{j\phi_k(t)} \tag{1}$$

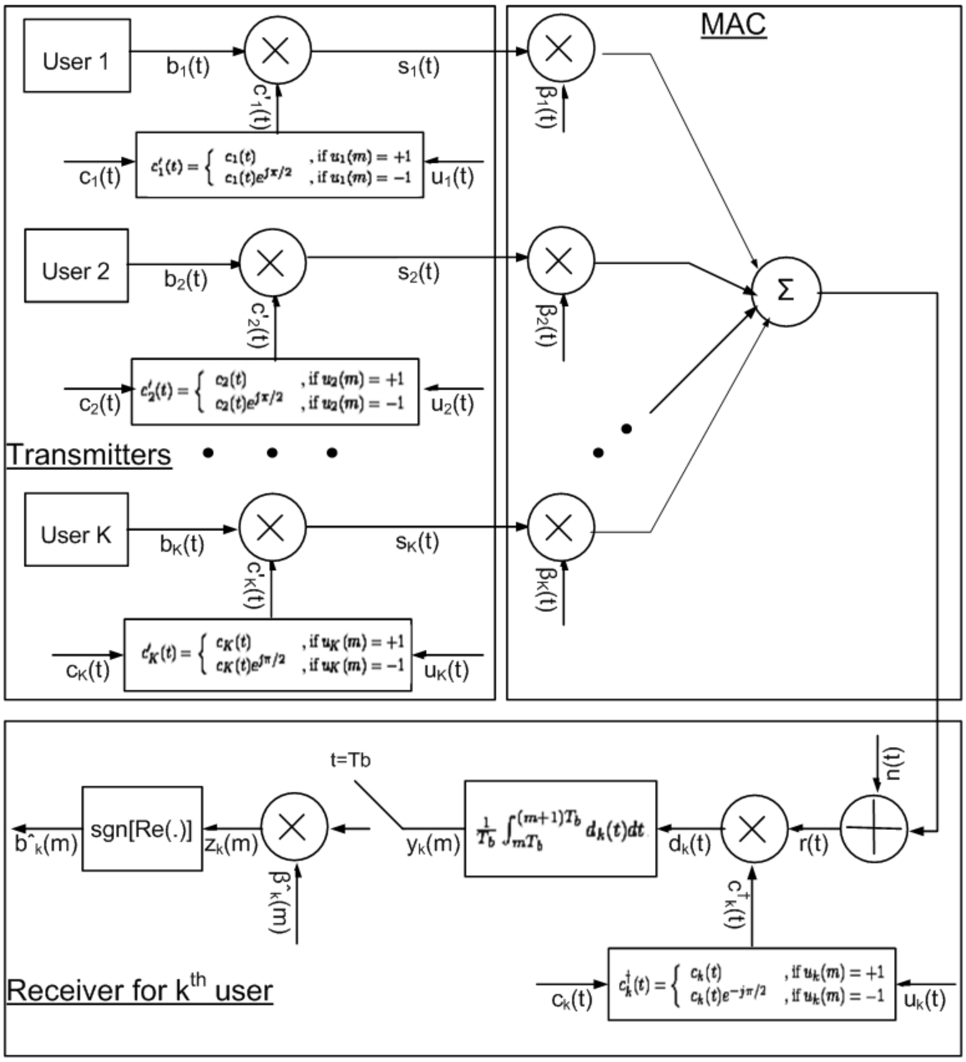
where $\alpha_k(t)$ and $\phi_k(t)$ denote amplitude and phase component, respectively. For $\beta_k(t) = 1, \forall k$ this corresponds to AWGN only channel with no fading. The received composite signal $r(t)$ from K users' transmission can be written as:

$$r(t)=\sum_{k=1}^{K} \beta_k(t)s_k(t) + n(t), \tag{2}$$

where, $s_k(t)$ is the transmitted signal of k^{th} user and is given by

$$s_k(t) = \sqrt{P_k}b_k(t)c_k'(t) \tag{3}$$

where, P_k is the signal power, $b_k(t) = \sum_{m=-\infty}^{\infty} b_k(m)p(t - mT_b)$ is the data signal and $b_k(m)$ is a binary sequence taking values $[-1, +1]$ with equal probabilities, $p(t)$ is a rectangular pulse with period T_b, $c_k'(t)$ is the precoded spreading sequence, which is a function of $c_k(t)$ and $u_k(t)$), where $c_k(t) = \sum_{n=-\infty}^{\infty} c_k(n)p(t - nT_c)$ is the original spreading sequence of k^{th} user with chip values $[-1, +1]$ using rectangular pulse $p(t)$ with period T_c and $u_k(t)$ is the k^{th} user's precoding sequence. The spreading factor used is $N = T_b/T_c$, and $n(t)$ is the AWGN with two sided power spectral density $N_0/2$.



A.1: System model of precoding based spreading

A.3 Precoding based semi-complex spreading and despreading

Under this scheme, at the transmitter, each user is assigned a repeating unique random binary precoding sequence $u_k(t)$ with period of L symbols, along with the user's individual spreading sequence $c_k(t)$. The minimum length is L and has to satisfy $L_{min} = \lceil \log_2(K) \rceil$, so that each sequence $u_k(t)$ is unique, where $\lceil x \rceil$ denotes an integer value greater than x. At each symbol period mT_b, the user data signal $b_k(m)$ is multiplied with the precoded spreading sequence $c'_k(t)$, which is chosen according to the following rule:

$$c'_k(t) = \begin{cases} c_k(t) & \text{, if } u_k(m) = +1 \\ c_k(t)e^{j\pi/2} & \text{, if } u_k(m) = -1 \end{cases} \tag{4}$$

Without loss of generality, it is assumed that the desired user is the k^{th} user, and that the receiver has acquired perfect synchronisation of spreading and precoding sequence of the user. In the case of transmission through complex fading channels , the receiver has to estimate the channel parameter (phase) and coherently despread the chip matched filtered received signal. The despreading of $r(t)$ for k^{th} user is done by multiplying $r(t)$ with known spreading sequence $c_k(t)$ to form a complex valued signal as follows

$$d_k(t) = r(t)c^{\dagger}_k(t), \tag{5}$$

where multiplication by $c^{\dagger}_k(t)$ is used to constrain the despread decision variable signal to be in real axis, $c^{\dagger}_k(t)$ is obtained as follows:

$$c^{\dagger}_k(t) = \begin{cases} c_k(t) & \text{, if } u_k(m) = +1 \\ c_k(t)e^{-j\pi/2} & \text{, if } u_k(m) = -1 \end{cases} \tag{6}$$

The output of the $d_k(t)$, which consists of contributions of the desired user and interfering user signals, is integrated over the symbol period to form the soft estimate of data signal for the k^{th} user given as:

$$y_k(m) = \frac{1}{T_b} \int_{mT_b}^{(m+1)T_b} d_k(t)dt. \tag{7}$$

The desired decision variable $z_k(m)$ is obtained from $y_k(m)$ by multiplying the latter with the complex conjugate of the estimated channel, $\tilde{\beta}_k(m)$ given by

$$z_k(m) = \{y_k(m)\tilde{\beta}^*_k(m)\}. \tag{8}$$

Finally, the symbol decision for the desired user is obtained by taking the sign of the real part of $z_k(m)$ as follows:

$$\hat{b}_k(m) = sgn\Big\{\Re\{z_k(m)\}\Big\}. \tag{9}$$

A.4 BER Performance Analysis

From the probability theory, it is known that a long random binary sequence can also be considered as an independent and identically distributed (i.i.d.) random variable with equal probability of containing -1 or $+1$. In a random binary sequence of short length, the probabilities are approximately equal. From this approximation, it can be generalized that in a system with K users each employing random sequences, at a symbol period mT_b, there are K/2 users with $u_i(m) = -1$ and other $K/2$ users with $u_i(m) = +1$. The cross correlation of any user's spreading sequence $c_i, i \neq k$ to the desired k^{th} user's sequence can be written as

$$\begin{aligned} \rho_{k,i} &= c\dagger_k c_i' \cos(\angle c\dagger_k - \angle c_i'); \\ 0 &\leq \angle c\dagger_k, \angle c_i' \leq 2\pi. \end{aligned} \tag{10}$$

where $\angle$ denotes the phase. If k^{th} user is assumed with $u_k(m) = 1$, then $(K-1)/2$ signals with $u_i(m) = -1, i \neq k,$, will have relative phase difference of $\pi/2$ and hence in an AWGN channel, their cross correlation becomes zero

$$\cos(\angle c\dagger_k - \angle c_i') = 0 \longrightarrow \rho_{k,i} = 0. \tag{11}$$

The vector of decision variable signal of users $z_k \in \mathbf{z} = [z_1, z_2, ..z_K]^T$ at any time instant can also be shown in matrix notation as

$$\mathbf{z} = \Re\{\mathbf{RAb} + \mathbf{n}\}. \tag{12}$$

For the purpose of visualization, the cross correlation matrix $\Re\{\mathbf{R}\}$ for $K = 5$ is shown with users assigned the unique sequences $u_k(m), \forall k$ such that the cross correlation of users' spreading sequences have equal number of $\rho_{k,i} = 0$ and $\rho_{k,i} \neq 0$ on the off-diagonal positions shown as

$$\begin{aligned} \Re\{\mathbf{R}\} &= \begin{bmatrix} 1 & 0 & 0 & \rho_{14} & \rho_{15} \\ \rho_{21} & 1 & 0 & 0 & \rho_{25} \\ 0 & \rho_{31} & 1 & \rho_{34} & 0 \\ \rho_{41} & 0 & \rho_{43} & 1 & 0 \\ \rho_{51} & 0 & \rho_{53} & 0 & 1 \end{bmatrix} ;, \text{if } u_k(m) = +1 \\ \Re\{\mathbf{R}\} &= \begin{bmatrix} 1 & \rho_{12} & \rho_{13} & 0 & 0 \\ 0 & 1 & \rho_{23} & \rho_{24} & 0 \\ \rho_{31} & 0 & 1 & 0 & \rho_{35} \\ 0 & \rho_{42} & 0 & 1 & \rho_{45} \\ 0 & \rho_{52} & 0 & \rho_{54} & 1 \end{bmatrix} ;, \text{if } u_k(m) = -1. \end{aligned} \tag{13}$$

The diagonal matrix of combinations of users' amplitude and channel gains **A** and the data vector **b** in (12) are shown as

$$\mathbf{A}_{1\times K} = diag[\sqrt{P_1}\beta_1, \sqrt{P_2}\beta_2, .., \sqrt{P_K}\beta_K]. \tag{14}$$

$$\mathbf{b}_{1\times K} = [b_1, b_2, .., b_K]^T. \tag{15}$$

The correlated, circularly symmetric complex Gaussian noise vector **n** is given by

$$\mathbf{n}_{1\times K} = [n_1, n_2, .., n_K]^T. \tag{16}$$

From (4) and using (6)-(9), the 'desired' signal part, which consists of the contribution from the desired user data signal and approximately half, i.e. $(K-1)/2$ sum of interfering user signals and half the receiver noise only, is extracted. The remaining $(K-1)/2$ user signals and the noise do not effect the decision variable signal because of their zero cross correlation. This leads to the expression of the decision variable of the desired user $z_k(m)$ shown as

$$z_k(m) = \sqrt{P_k(m)}b_k(m) + n_k/2 + \sum_{i=1, i\neq k}^{\approx(K-1)/2} \sqrt{P_i(m)}\rho_{k,i}b_i(m). \tag{17}$$

For DS-CDMA systems, there exists several methods for calculations of probabilities of bit error (BEP), the simplest and the most common being the GA method used in e.g. [36]. It is assumed that the combined noise and interference signal terms under the proposed scheme are Gaussian distributed. The signal to interference and noise ratio (SINR) is used to obtain the BEP of such systems, which is given by

$$Pe = Q\left(\sqrt{SINR}\right) \tag{18}$$

where $Q(x) = \frac{1}{\sqrt{2\pi}}\int_x^\infty e^{-t^2/2}dt$ is the well-known Gaussian error function. The BEP for a K user synchronous CDMA system with real binary and quadrature spreading using random sequences are derived in [141]. The BEP for the system with the proposed scheme Pe^{pre} can be obtained by knowing the fact that at any time instant, there are effectively only $(K-1)/2$ interferering user signals in the system (see (17) above). The value of $\rho_{k,i}$ can be used in SINR expression to calculate BEP Pe^{pre} for the proposed system using different sequences from (18) as follows:

$$Pe^{pre} \simeq Q\left(\left[\frac{N_0}{2E_b} + \sum_{i=1}^{\simeq(K-1)/2} \rho_{k,i}^2\right]^{-1/2}\right). \tag{19}$$

When long random sequences are used for spreading users' data, the MAI power due to the cross correlation can be approximated $\rho_{k,i}^2 \simeq 1/N$ [36]. Thus BEP of the proposed system using random sequences Pe_{pn}^{pre} is given by

$$Pe_{pn}^{pre} \simeq Q\left(\left[\frac{N_0}{2E_b} + \frac{K-1}{2N}\right]^{-1/2}\right) \tag{20}$$

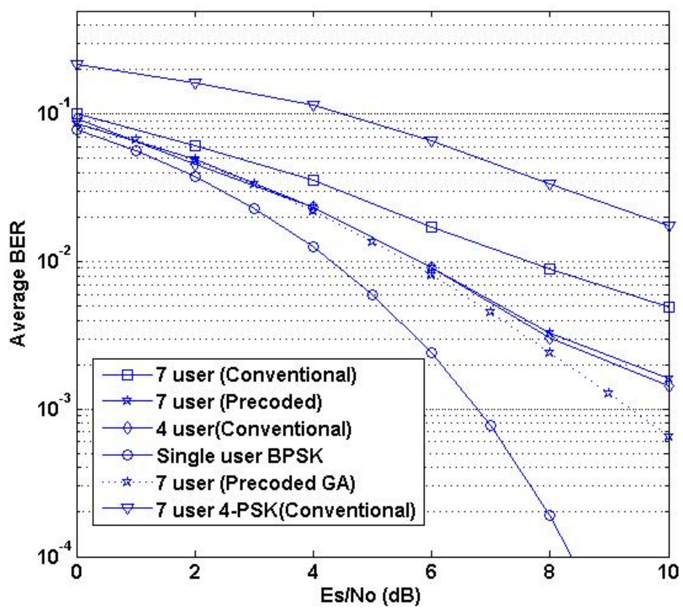
where E_b/N_0 is the signal to noise ratio per bit of desired user. The BEP analysis can also be extended for other sequences such as Gold, Kasami [89]. Since these are short sequences that repeat at every symbol period, the cross correlation values $\rho_{k,i}$ are fixed, they can be used in (19) directly to obtain the BEP.

The performance of the proposed scheme in fading channels is evaluated using computer simulations. The channel gains are modeled as unit variance complex Gaussian random variables following Rician distribution. The intensity of fading is also known as the Rician K-factor, given by

$$\kappa = \mu^2 / 2\sigma^2 \tag{21}$$

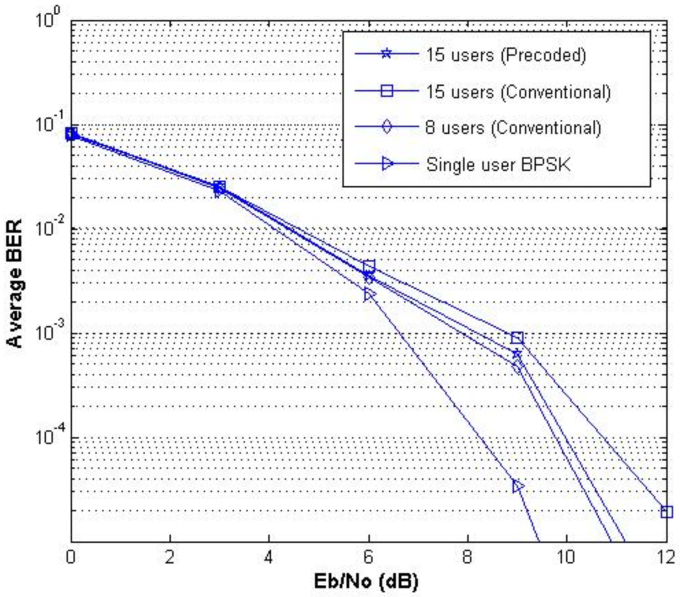
where μ^2 is the power of direct line of sight component and $2\sigma^2$ is the total power of many weaker diffused components. This channel model represents Rayleigh fading and AWGN channels as special cases with $\kappa = 0$ and $\kappa = \infty$, respectively.

A.5 Numerical Results



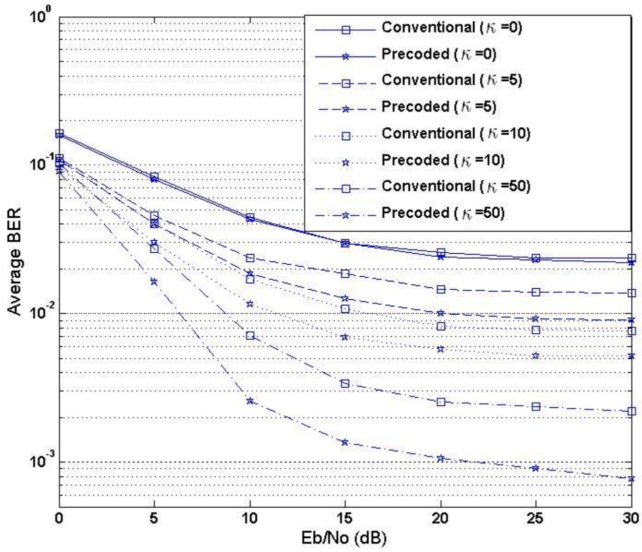
A.2: Performance of the proposed semi-complex spreading scheme in AWGN channel for 7 users and PN sequences of N=64

A baseband model of K user uplink synchronous DS-CDMA system with BPSK in AWGN channel is simulated in MATLAB. Long random (PN) sequences with N=64 and random precoding sequences with L=64 are used for all users. Figure A.2 shows the BER performance results of a 7 user system using the real binary spreading scheme [141] (Conventional) and the proposed scheme (Precoded). Also BER of systems with the conventional scheme with 4 users with BPSK and 7 user with 4-PSK data signaling are shown for the comparison. As expected, the system with the proposed scheme with 7 users has comparatively better BER, approaching that of 4 BPSK user system with conventional scheme, as predicted by the analysis from (20). Therefore, at the same



A.3: Performance of the proposed semi-complex spreading scheme in AWGN channel for 15 users and Gold sequences of N=31

BER, the proposed scheme can support approximately $K-1$ additional users. In terms of power efficiency, the proposed scheme compared with the conventional scheme requires approximately 2 dB of less power than 7 users with BPSK and $\approx 5dB$ less than 7 users with unity power 4-PSK data signalling (2 bits) to achieve the same BER of 0.01. The BER obtained from the GA of the proposed scheme (Precoded GA) is optimistic as expected [36], [141].



A.4: Performance of the proposed semi-complex spreading scheme in Rician channels for 7 users and PN sequences of N=64

The performance of the system with Gold spreading sequences is also investigated. These sequences exhibit improved cross correlation properties and thus show better error performances compared to random PN sequences [89]. Short Gold sequences with N=31 and random precoding sequences with L=31 are used for all users. The initial code acquisition for such sequences are investigated in [90]. The effect of precoding based spreading for users employing Gold sequences on the BER performance is shown in Figure A.3 for 15 users. As expected, the BER of 15 users with precoding shows improved BER compared to 15 users with conventional binary spreading. It is also noted that BER of 15 users with precoding is very close to that of conventional spreading with 8 users, confirming our analysis.

The BER simulation results of a 7-user system, employing PN spreading sequences $N = 64, L = 64$ in fading channels with different κ are shown in Figure A.4. It is evident that the proposed scheme (Precoded) shows gain in the BER compared to real binary (Conventional) scheme over the range of κ values, with the gain increasing proportionally to the increase of κ. With $\kappa = 0$, their relative performances remained identical. This is expected, as in Rayleigh fading channel all user signals have statistically identical phase distribution between $[0, 2\pi]$, in which case all user signals effect the desired decision variable equally in the both schemes.

A.6 Conclusion

A simple and power efficient precoding based spreading scheme for BPSK DS-CDMA is proposed. The scheme is shown to reduce MAI by half and hence much improved BER for the same number of users is achieved. Its performance is evaluated using different spreading sequences and channel conditions and verified from both analysis and simulations in AWGN. In fading environments, the scheme exhibited good BER performance particularly when power of diffused components are low.

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